Study of dielectric-loaded waveguide slot antennas

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Study of Dielectric-Loaded Waveguide Slot Antennas

Zheng Boyu

School of Electrical & Electronic Engineering

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in fulfillment of the requirement for the degree of
Doctor of Philosophy

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Summary

Waveguide slot antennas have found wide applications in modern communication systems such as aircraft, spacecraft, radar and television. Nowadays, the waveguide slot antenna plays an important role in wireless communication systems due to its attractive features such as low loss, high efficiency, high power handling capacity and ease of fabrication. However, with the rapid progress of space applications, higher requirements have risen, which require that waveguide slot antennas have more compact sizes and a lighter weight as well as a higher gain and lower side lobes. The size and weight of waveguide slot antennas can be reduced by loading dielectric materials in the conventional air-filled waveguides. The guided wavelength in the dielectric-loaded waveguide can then be shortened, which is very useful in array applications. These antennas are called dielectric-loaded waveguide slot antennas and they are the main subjects of study in this thesis.

Slots cut on the wall of an air-filled rectangular waveguide have been extensively investigated by many researchers. However, only a few research works have been reported on dielectric-loaded waveguide slot antennas. In this thesis, a thorough study has been conducted on the scattering and circuit parameters, radiation patterns and mutual couplings of dielectric-loaded waveguide slot antennas and ar-

rays.

A hybrid mode-matching/moment technique has been developed for analyzing both transverse and longitudinal slots cut on the broad wall of a rectangular waveguide partially loaded with an H-plane dielectric slab. The slot structure has been treated as two parts, a dielectric-loaded waveguide T-junction and an open-ended waveguide radiating into half space. The two blocks has been formulated by mode-matching and moment method, respectively. Cascading these two blocks has given the scattering parameters of the waveguide slot structure. Comparison between our calculated data and those obtained by commercial software has demonstrated the accuracy and efficiency of our new full-wave method. Parametrical studies of the dielectric-loaded waveguide slot have shown the effect of the slab's thickness and dielectric constant as well as the waveguide wall thickness on the slot antenna's impedance.

The radiation patterns of the dielectric-loaded waveguide slot antennas have been investigated by the combination of the proposed hybrid method and the Geometrical Theory of Diffraction. The effect of a finite size of the waveguide's broad wall on the radiation patterns of the slot antennas has also been studied.

The mutual coupling between rectangular slots cut on a partially dielectric-loaded rectangular waveguide has been examined. The internal and external mutual coupling has been theoretically taken into consideration in our approach. It has been demonstrated that our analyzing method is also accurate and efficient in predicting mutual coupling between dielectric-loaded waveguide slots.

In order to appreciate the advantage of the compact size of dielectric-loaded waveguide slot antenna, an 8-element nonuniform slot array has been designed, fabricated and tested. The structural parameters of the slot element as well as the inter-element spacings have been determined based on our rigorous full-wave analysis. The measured return loss of the slot array has shown that the 10 dB bandwidth is about 17% at the operating frequency of 10 GHz. It has also been observed that the main-to-side lobe ratio is better than 15 dB in the frequency range of 1 GHz, which meets the objective of our research.

Table of Contents

Acknowledgments	i
Summary	ii
List of Figures	ix
List of Tables	xviii
List of Acronyms	xix
1 Introduction	1
1.1 Motivation	1
1.1.1 Basics of Slot Antennas	1
1.1.2 History of Slot Antennas	3
1.1.3 Waveguide Slot Antennas	5
1.2 Objectives	8
1.3 Major Contributions of the Thesis	10
1.4 Organization of the Thesis	13
2 Literature Review of Waveguide Slot Antennas	15
2.1 Introduction	15

2.2	Air-filled Waveguide Slot Antennas	16
2.3	Dielectric-loaded Waveguide Slot Antennas	20
2.4	Analysis Methods	21
3	S-parameters of Dielectric-loaded Waveguide Slot Elements	24
3.1	Introduction	24
3.2	Analysis Model	27
3.2.1	Problem Geometry	27
3.2.2	Methodology	28
3.3	Formulation	29
3.3.1	Modes in a Waveguide Partially Filled with a Dielectric Slab	30
3.3.2	Three-port T-junction	32
3.3.3	Open-ended Waveguide	44
3.3.4	Cascading Technique	54
3.4	Numerical Results	55
3.4.1	Air-filled Waveguide Slots	56
3.4.2	Dielectric-loaded Waveguide Slots	59
3.4.3	Parametrical Study	63
3.5	Conclusion	74
4	Radiation Patterns of Dielectric-loaded Waveguide Slots	76
4.1	Introduction	76
4.2	Field Distribution along the Waveguide Slot	79
4.3	Calculation of Radiation Patterns	81
4.3.1	GO Fields	81
4.3.2	GTD Formulation	82

4.4	Numerical Results	89
4.4.1	Air-filled Rectangular Waveguide Slot Antennas	89
4.4.2	Partially Dielectric-loaded Rectangular Waveguide Slot Antennas	93
4.4.3	Effect of Different Wall Sizes on Waveguide Slots' Radiation Patterns	98
4.5	Conclusion	101
5	Mutual Coupling between Dielectric-loaded Waveguide Slots	102
5.1	Introduction	102
5.2	Analysis Model	103
5.3	Numerical Results	108
5.3.1	Mutual Coupling between Two Open-ended Waveguides	108
5.3.2	Mutual Coupling between Three or More Open-ended Waveguides	112
5.3.3	Two Slots Cut on Partially Dielectric-loaded Rectangular Waveguide	116
5.3.4	Circuit Parameters of Two Slots Structure	119
5.3.5	An Array of Three Slots	123
5.4	Conclusions	128
6	Dielectric-Loaded Waveguide Slot Antenna Arrays	130
6.1	Introduction	130
6.2	Design Considerations	132
6.2.1	Choosing the Slot Element	132
6.2.2	Slot Element Characteristics	136

6.2.3	Mutual Coupling between Slot Elements	143
6.3	Geometry of the Antenna Array	146
6.3.1	Optimizing the Inter-element Spacings	146
6.3.2	Feeding Structure	153
6.4	Measured Results	156
6.5	Conclusions	163
7	Conclusion and Recommendations	165
7.1	Conclusion	165
7.2	Recommendations for Further Research	169
	Author's Publications	171
	Bibliography	173

List of Figures

1.1	Different types of slot antennas.	3
2.1	Air-filled rectangular waveguide slot antenna.	17
2.2	Dielectric-loaded rectangular waveguide slot antenna.	20
3.1	Dielectric-loaded rectangular waveguide slot.	28
3.2	Decomposition of the dielectric-loaded waveguide slot; (a) physical structure, and (b) decomposed into two parts.	29
3.3	Cross-section of a dielectric-loaded rectangular waveguide.	31
3.4	(a) Normalized phase constants β/k_0 , and (b) Attenuation constants α in an H-plane dielectric slab loaded waveguide ($a = 22.86 \text{ mm}$, $b = 10.16 \text{ mm}$, $t = 7.16 \text{ mm}$, $\epsilon_{r1} = 1.0$ and $\epsilon_{r2} = 2.5$).	33
3.5	(a) Normalized phase constants β/k_0 , and (b) Attenuation constants α in an E-plane dielectric slab loaded waveguide ($a = 10.16 \text{ mm}$, $b = 22.86 \text{ mm}$, $t = 10 \text{ mm}$, $\epsilon_{r1} = 2.5$ and $\epsilon_{r2} = 1.0$).	34
3.6	Waveguide T-junction loaded with a dielectric slab.	35
3.7	Scattering parameters of an air-filled rectangular waveguide T-junction with a side arm of reduced sized waveguide ($a = 34.85 \text{ mm}$, $b = 15.799 \text{ mm}$, $w = 22.86 \text{ mm}$, $l = 10.16 \text{ mm}$).	41

3.8	Scattering parameters of a dielectric-loaded rectangular waveguide T-junction with a side arm of reduced size waveguide ($a = 34.85$ mm, $b = 15.799$ mm, $w = 22.86$ mm, $l = 10.16$ mm, $t = 7.5$ mm, $\epsilon_{r1} = 1$, $\epsilon_{r1} = 2.5$).	42
3.9	Relative error for $ S_{11} $ of a dielectric-loaded waveguide T-junction (Dimensional parameters are the same as in Fig. 3.8).	43
3.10	The field equivalent principle applied for the open-ended waveguide; (a) open-ended waveguide, and (b) equivalent problem.	45
3.11	Mapping of the range of the 4-D integral.	50
3.12	Reflection coefficient of an open-ended rectangular waveguide (WR-90: $22.86\text{mm} \times 10.16\text{mm}$); (a) magnitude, and (b) phase.	53
3.13	Frequency dependence of normalized series resistance and reactance for a centered transverse slot in an air-filled waveguide ($a = 22.86$ mm, $b = 10.16\text{mm}$, $w = 14\text{mm}$, $l = 3\text{mm}$, $s = 1\text{mm}$).	56
3.14	Normalized shunt (a) conductance and (b) susceptance versus the length of the slot for a longitudinal slot in an air-filled waveguide ($a = 22.86\text{mm}$, $b = 10.16\text{mm}$, $w = 1.5875\text{mm}$, $d = 1.8796\text{mm}$, $s = 1.27\text{mm}$, frequency is 9.375 GHz).	58
3.15	Scattering parameters of a transverse slot in a rectangular waveguide partially filled with a dielectric slab ($a = 15\text{mm}$, $b = 6\text{mm}$, $t = 3\text{mm}$, $w = 15\text{mm}$, $l = 6\text{mm}$, $s = 1\text{mm}$, $\epsilon_{r1} = 6$, $\epsilon_{r2} = 1$).	59
3.16	Normalized series resistance and reactance for a centered transverse slot in a rectangular waveguide partially filled with a dielectric slab (Parameters as in Fig. 3.15).	60

3.17	Scattering parameters of a longitudinal slot in a rectangular waveguide partially filled with a dielectric slab ($w = 3mm$, $l = 14mm$, $d = 3.75mm$; Other parameters the same as in Fig. 3.15).	61
3.18	Normalized shunt admittance for a longitudinal slot in a rectangular waveguide partially filled with a dielectric slab ($w = 3mm$, $l = 14mm$, $d = 3.75mm$; Other parameters the same as in Fig. 3.15).	62
3.19	Normalized series resistance and reactance for a partially dielectric-loaded waveguide transverse slot with different permittivities of the dielectric slab ($a = 15mm$, $b = 6mm$, $t = 3mm$, $w = 15mm$, $l = 6mm$, $s = 1mm$, $\varepsilon_{r2} = 1$).	64
3.20	Normalized series resistance and reactance for a partially dielectric-loaded waveguide transverse slot with different thicknesses of the dielectric slab ($a = 15mm$, $b = 6mm$, $w = 15mm$, $l = 6mm$, $s = 1mm$, $\varepsilon_{r1} = 6$, $\varepsilon_{r2} = 1$).	65
3.21	Normalized shunt conductance and susceptance for a partially dielectric-loaded waveguide longitudinal slot with different permittivities of the dielectric slab ($a = 15mm$, $b = 6mm$, $t = 3mm$, $w = 3mm$, $l = 14mm$, $d = 3.75mm$, $s = 1mm$, $\varepsilon_{r2} = 1$).	66
3.22	Normalized shunt conductance and susceptance for a partially dielectric-loaded waveguide longitudinal slot with different thicknesses of the dielectric slab ($a = 15mm$, $b = 6mm$, $w = 3mm$, $l = 14mm$, $d = 3.75mm$, $s = 1mm$, $\varepsilon_{r1} = 6$, $\varepsilon_{r2} = 1$).	67

3.23	Normalized series (a) resistance and (b) reactance for a partially dielectric-loaded waveguide transverse slot when the dielectric slab is at different layers of the waveguide ($a = 15mm$, $b = 6mm$, $w = 15mm$, $l = 6mm$, $s = 1mm$).	68
3.24	Normalized series resistance and reactance for a transverse slot cut on an air-filled rectangular waveguide ($a = 22.86mm$, $b = 10.16mm$, $w = 14.1mm$, $l = 5mm$, $s = 1mm$).	69
3.25	Normalized series resistance and reactance for a transverse slot cut on a partially dielectric-loaded rectangular waveguide ($a = 15mm$, $b = 6mm$, $t = 3mm$, $w = 14.5mm$, $l = 5mm$, $s = 1mm$, $\varepsilon_{r1} = 6$, $\varepsilon_{r2} = 1$).	70
3.26	Normalized series resistance and reactance of a dielectric-loaded waveguide slot with different thicknesses of the waveguide wall (Parameters are the same as in 3.25 except the thickness s of the waveguide wall).	72
3.27	Relative error for $ S_{11} $ of a dielectric-loaded waveguide slot (Parameters are the same as in 3.25 except the thickness s of the waveguide wall; Mode numbers for Regions 1 and 4 in Fig. 3.6 are 50 and 200, respectively).	73
4.1	Waveguide slot antenna on a finite ground plane.	83
4.2	E-plane radiation and diffraction mechanisms.	84
4.3	H-plane radiation and diffraction mechanisms.	86
4.4	Equivalent magnetic current for the H-plane radiation pattern. . . .	88

4.5	Radiation patterns of a longitudinal slot antenna in an air-filled rectangular waveguide ($a = 22.86mm$, $b = 10.16mm$, $w = 3mm$, $l = 14.3mm$, $d = 4.6mm$, $s = 1mm$, $HD1 = HD2 = 60mm$, $HC1 = 17.03mm$, $HC2 = 7.83mm$, $f = 10$ GHz).	91
4.6	Radiation patterns of a transverse slot antenna in an air-filled rectangular waveguide ($a = 22.86mm$, $b = 10.16mm$, $w = 14.1mm$, $l = 5mm$, $d = 3.15mm$, $s = 1mm$, $HD1 = HD2 = 60mm$, $HC1 = 9.28mm$, $HC2 = 15.58mm$, $f = 10$ GHz).	92
4.7	Radiation patterns of a longitudinal slot antenna in a partially dielectric-loaded rectangular waveguide ($a = 15mm$, $b = 6mm$, $t = 3mm$, $w = 3mm$, $l = 14mm$, $d = 3.75mm$, $s = 1mm$, $\epsilon_{r1} = 6$, $\epsilon_{r2} = 1$, $HD1 = HD2 = 60mm$, $HC1 = 22.75mm$, $HC2 = 30.25mm$, $f = 10$ GHz).	95
4.8	Radiation patterns of a centered transverse slot antenna in a partially dielectric-loaded rectangular waveguide ($a = 15mm$, $b = 6mm$, $t = 3mm$, $w = 15mm$, $l = 6mm$, $s = 1mm$, $\epsilon_{r1} = 6$, $\epsilon_{r2} = 1$, $HD1 = HD2 = 60mm$, $HC1 = HC2 = 26.5mm$, $f = 10$ GHz).	96
4.9	E-plane radiation pattern of a centered transverse slot in a partially dielectric-loaded rectangular waveguide (Parameters are the same as in Fig. 4.8 except the frequency f).	97
4.10	Radiation patterns of a longitudinal slot antenna in a partially dielectric-loaded rectangular waveguide with different sizes of ground plane ($a = 15mm$, $b = 6mm$, $t = 3mm$, $w = 3mm$, $l = 14mm$, $d = 3.75mm$, $s = 1mm$, $\epsilon_{r1} = 6$, $\epsilon_{r2} = 1$, $f = 10$ GHz).	99

4.11	Radiation patterns of a centered transverse slot antenna in a partially dielectric-loaded rectangular waveguide with different sizes of ground plane ($a = 15mm$, $b = 6mm$, $t = 3mm$, $w = 15mm$, $l = 6mm$, $s = 1mm$, $\varepsilon_{r1} = 6$, $\varepsilon_{r2} = 1$, $f = 10$ GHz).	100
5.1	Dielectric-loaded rectangular waveguide slots.	104
5.2	Equivalent model of the dielectric-loaded waveguide slots.	105
5.3	Geometry of two coupled rectangular waveguides.	109
5.4	Mutual coupling between two rectangular waveguides ($a_1 = a_2 = 22.86$ mm, $b_1 = b_2 = 10.16$ mm, $D_x = 25.4$ mm, $D_y = 0$ mm).	110
5.5	Mutual coupling between two rectangular waveguides ($a_1 = a_2 = 10.16$ mm, $b_1 = b_2 = 22.86$ mm, $D_x = 12.7$ mm, $D_y = 0$ mm).	111
5.6	(a) Geometry and (b) coupling coefficients of three collinear rectangular waveguides with narrow side coupling ($a = 22.86mm$, $b = 10.16mm$ and $D = 25.4mm$).	113
5.7	(a) Geometry and (b) coupling coefficients of three collinear rectangular waveguides with broad side coupling ($a = 10.16mm$, $b = 22.86mm$ and $D = 12.7mm$).	114
5.8	(a) Geometry and (b) coupling coefficients of four rectangular waveguides ($a = 22.86mm$, $b = 10.16mm$, $D_1 = 25.4mm$ and $D_2 = 12.7mm$).	115
5.9	Scattering parameters of two longitudinal slots in a rectangular waveguide partially loaded with a dielectric slab ($a = 15mm$, $b = 6mm$, $t = 3mm$, $s = 1mm$, $\varepsilon_{r1} = 6$, $\varepsilon_{r2} = 1$, $w_1 = w_2 = 3mm$, $l_1 = l_2 = 14mm$, $d_1 = d_2 = 3.75mm$); (a) $D = 5mm$, and (b) $D = 10mm$.	117

5.10	Scattering parameters of two transverse slots in a rectangular waveguide partially loaded with a dielectric slab ($w_1 = w_2 = 15mm$, $l_1 = l_2 = 6mm$, $d_1 = d_2 = 0mm$; other parameters are the same as in Fig. 5.9); (a) $D = 5mm$, and (b) $D = 10mm$	118
5.11	Block diagram of two slot antennas in the broad wall of a partially dielectric-loaded rectangular waveguide.	119
5.12	Circuit model of two waveguide slot antennas.	120
5.13	Frequency dependency of the mutual impedance between two transverse slots in the broad wall of a partially dielectric-loaded waveguide(Distance between the two slots is 20 mm).	121
5.14	Frequency dependency of the mutual impedance between two transverse slots in the broad wall of a partially dielectric-loaded waveguide(Distance between the two slots is 60 mm).	122
5.15	Magnitude and phase of the mutual impedance versus slot separation (Structural parameters are $a = 15mm$, $b = 6mm$, $s = 1mm$, $t = 3mm$, $\epsilon_{r1} = 6$, $\epsilon_{r2} = 1$, $w_1 = w_2 = 15mm$, $l_1 = l_2 = 6mm$). . . .	124
5.16	Geometry of a linear array consisting of three slots.	125
5.17	Scattering parameters of three transverse slots cut on a partially dielectric-loaded rectangular waveguide ($D_1 = D_2 = 11mm$). . . .	126
5.18	Scattering parameters of three transverse slots cut on a partially dielectric-loaded rectangular waveguide ($D_1 = D_2 = 19mm$). . . .	127
5.19	E-plane radiation patterns of three transverse slots cut on a partially dielectric-loaded rectangular waveguide ($f = 10$ GHz).	128
6.1	(a) Standing-wave and (b) traveling-wave longitudinal slot arrays. .	133

6.2	Transverse slot element in the broad wall of a rectangular waveguide partially loaded with a dielectric slab.	136
6.3	Equivalent circuit of a transverse slot in the wall of a partially dielectric-loaded rectangular waveguide.	137
6.4	Normalized resistance and reactance components of a transverse slot on the broad wall of a partially dielectric-loaded rectangular waveguide ($w = 13.6mm$, $l = 6mm$, $a = 15mm$, $b = 7mm$, $t = 4.8mm$, $s = 1mm$, and $\epsilon_r = 3.5$).	139
6.5	A transverse slot represented by a π -network. (Parameters are the same as in Fig. 6.4.)	140
6.6	Return loss of the slot element on the wall of a partially dielectric-loaded rectangular waveguide with one end shorted ($d_1 = 11.6mm$ and $d_2 = 0mm$. Other parameters are the same as in Fig. 6.4). . . .	141
6.7	Input impedance of a single slot element (Parameters are the same as in Fig. 6.6).	142
6.8	Normalized guided wavelength of the dominant mode in a partially dielectric-loaded rectangular waveguide. ($a = 15mm$, $b = 7mm$, $t = 4.8mm$, and $\epsilon_r = 3.5$).	143
6.9	Geometry and equivalent circuit of two coupling slot elements. . . .	144
6.10	Scattering parameters of the structure consisting of two slot elements ($a = 15mm$, $b = 7mm$, $t = 4.8mm$, $s = 1mm$, $\epsilon_r = 3.5$, $w = 13.6mm$, $l = 6mm$, $d_1 = 11.6mm$, $d_2 = 0mm$, $d = 18mm$). . . .	145
6.11	Mutual impedance between two slot elements (Parameters are the same as in Fig. 6.10).	146
6.12	Geometry of the 8-element slot array.	147

6.13	Normalized array factor pattern of the 8-element nonuniform linear array ($d_1 = 0.57\lambda_0$, $d_2 = 0.57\lambda_0$, $d_3 = 0.7\lambda_0$ and $d_4 = 0.7\lambda_0$, $f = 10$ GHz).	148
6.14	Simulated H-plane radiation patterns of the 8-element nonuniform slot array at 10 GHz.	149
6.15	Simulated H-plane radiation patterns of the 8-element nonuniform slot array at 10 GHz (Mutual coupling is not included).	151
6.16	Simulated return loss and coupling coefficients of (a) Port 1 and (b) Port 5 ($d_1 = 17mm$, $d_2 = 17mm$, $d_3 = 24mm$ and $d_4 = 26mm$).	152
6.17	Geometry of the SMA to waveguide transition section.	153
6.18	Simulated scattering parameters of the SMA to waveguide transition (single transition, $t_1 = 5.8mm$ and $t_2 = 4.8mm$).	154
6.19	Photograph of the fabricated one-to-eight Wilkinson power divider.	155
6.20	(a) Top and (b) bottom views of the fabricated nonuniform waveguide slot array.	157
6.21	Measured scattering parameters of the slot elements.	158
6.22	Measured return loss of the 8-element slot array.	159
6.23	Measured radiation patterns of the 8-element nonuniform slot antenna array at 10 GHz.	160
6.24	Measured H-plane radiation patterns of the 8-element nonuniform slot antenna array at (a) 9.5 GHz and (b) 10.5 GHz.	162

List of Tables

4.1	Front-to-back ratio of slot antennas on the broad wall of a partially dielectric-loaded waveguide with different wall sizes.	98
6.1	Comparison between calculated and measured gain and main-to-side lobe ratio of the 8-element nonuniform slot array.	161

List of Acronyms

ABBREVIATIONS	FULL EXPRESSIONS
1-D, 2-D, 3-D, 4-D	One dimensional, two dimensional, three dimensional, four dimensional
BW	Bandwidth
CBS	Cavity-backed slot
dB	Decibel
EM	Electromagnetic
FD	Finite-difference
FDFD	Finite-difference frequency domain
FDTD	Finite-difference time domain
FEM	Finite-element method
GO	Geometrical optics
GPS	Global positioning system
GTD	Geometrical theory of diffraction
HF	High frequency
HFSS	High frequency structure simulator
HPBW	Half-power beamwidth
LSE	Longitudinal section electric

LSM	Longitudinal section magnetic
MM	Mode-matching
MOM	Method of moment
PCB	Printed circuit board
RF	Radio frequency
SIW	Substrate integrated waveguide
SMA	Sub-miniature type A
TE	Transverse electric
TM	Transverse magnetic
TEM	Transverse electric and magnetic
UHF	Ultra high frequency
UTD	Uniform theory of diffraction
VHF	Very high frequency
VNA	Vector network analyzer

Chapter 1

Introduction

1.1 Motivation

1.1.1 Basics of Slot Antennas

An antenna is defined as ‘a means for radiating or receiving radio waves’ by the *IEEE Standard Definitions of Terms for Antennas* (IEEE Std 145-1983)¹. In other words, the antenna in a wireless system is an electromagnetic transducer, which is designed to convert guided waves within a guiding device to radiated free-space waves, or to convert free-space waves to guided waves [1]. It is a transmitting antenna in the former case or a receiving antenna in the latter.

In wireless communication systems, the antenna is one of the most important and critical components. A good design of the antenna can relax system requirements and improve overall system performance. In addition to transmitting and receiving energy, an antenna in a modern wireless system usually also acts as a

¹*IEEE Transactions on Antennas and Propagation*, vol. 31, no. 6, Part II, November 1983.

1.1 Motivation

directional device to enhance the transmitted or received energy in some directions while suppressing it in others [2].

An antenna can take various forms and can be classified in many ways. For example, according to their structures, we have wire antennas, aperture antennas, microstrip antennas, reflector antennas, lens antennas and array antennas. Different types of antennas can be employed in different systems to satisfy particular requirements of the radiation characteristics.

Among different types of antennas, aperture antennas are very common at microwave frequencies. Aperture antenna is, as its name implies, an antenna that utilizes an aperture as a radiator. The apertures can be covered with a dielectric material to protect them from a hazardous environment. Aperture antennas are preferred for space applications because they can be flush-mounted on the surface of the spacecraft and aircraft without disturbing their aerodynamic performance, which is critical in high speed applications.

There are many kinds of aperture antennas, such as slits, slots, open-ended waveguides and horns, etc. Slot antenna is a widely used type of aperture antenna. It has been widely used in high-performance radar and communication systems. Slot antenna has several forms such as microstrip-fed slot antennas, cavity-backed slot (CBS) antennas and waveguide slot antennas, as illustrated in Fig. 1.1.

1.1 Motivation

Figure 1.1: Different types of slot antennas.

1.1.2 History of Slot Antennas

The history of antennas [3] can be traced back to James Clerk Maxwell (1831 - 1879) who unified the theories of electromagnetics, and represented their relations through a set of well known *Maxwell's Equations* [4]. In 1886, Professor Heinrich Rudolph Hertz established the first complete radio system, which had an end-loaded half-wave dipole as transmitting antenna and a resonant square loop antenna as receiver. However, it was not until 1901 that Guglielmo Marconi was able to send signals over large distances. He used wires as transmitting and re-

1.1 Motivation

ceiving antennas to perform the first transatlantic transmission from Poldhu in Cornwall, England, to St. John's Newfoundland, Canada.

From Marconi's experiment through the 1940's, antenna technology was mostly focused on wire related radiating elements and frequencies up to about ultra high frequency (UHF). It was not until World War II that modern antenna technology was introduced and new antenna elements (such as waveguide apertures, horns, reflectors) were developed. Many microwave sources (such as the klystron and magnetron) were also invented with frequencies of 1 GHz and above.

Cavity-backed slot (CBS) antennas had been widely studied since the 1960's. Theoretical and experimental investigations had been conducted by many researchers [5]–[8]. The CBS antenna was constructed by placing a metal cavity at one side of the slot so that the energy can only be radiated from the other side. Cavity-backed slot antennas were primarily fed by a two-wire line at the center of the slot. Probe feeding was also introduced later. In this configuration, a probe is protruded into the cavity and connects the slot antenna to the standard sub-miniature type A (SMA) connector. Since the 1970's, microstrip-fed slot antennas have become very popular and have been extensively studied [9]–[11] with the rapid progress of printed circuit board (PCB) manufacturing technology. From then on, microstrip-fed CBS antennas were widely adopted [12] because this kind of CBS antennas can be fabricated with ease, low cost and high accuracy.

Waveguide slot antennas had found applications on aircraft and radar systems in the World War II. During this period, most of the research works were based

1.1 Motivation

on experimental data [13, 14] and had been considered almost a “cut and try” operation, though in the later 1940’s theoretical analysis method had been developed for resonating slots [15]. Waveguide slot antennas had also been employed to construct leaky (fast) wave antennas [16]–[20] in the 1950’s. The slot could be a longitudinal slot or an array consisting of closely spaced non-resonant transverse slots. A traveling wave is transmitted in the waveguide, which could be either circular or rectangular. Using a leaky wave slot antenna, one could adjust the dimensions of the slots and the waveguide to control the leakage and the main radiation direction of the energy. Besides these applications, waveguide slot antennas had also been adopted for satellite TV reception [21]–[23] to obtain a high gain as well as a high efficiency. Nowadays, the waveguide slot antenna has been employed in various areas due to its attractive features such as high efficiency, high power capacity and good solidity compared to microstrip antennas. Hence, our research work is mainly focused on waveguide slot antennas.

1.1.3 Waveguide Slot Antennas

The waveguide slot antenna consists of a waveguide and a slot cut on the wall of the waveguide. The waveguide can be rectangular or circular and the slot can be longitudinal, transverse or inclined. Because the slot cuts the current flow on the waveguide wall, electric field is generated along the slot and energy can be transmitted or received by the slot. Nowadays, waveguide slot antennas have found many applications in wireless communication systems. For example, longitudinal slot array antennas have been utilized as UHF band television antennas. The waveguide slot antenna is playing an very important role in modern space and marine applications because it can be flush-mounted on surfaces.

1.1 Motivation

In the 1940's, waveguide slot antennas were first introduced. At the first stage, theoretical studies were based on many approximations [15] and experimental data [13, 14, 24]. After several years, more accurate analysis methods were developed [25]–[27]. The waveguide slots could be then analyzed both at resonant and non-resonant frequencies. The circuit model was presented and the slot admittance could have been computed. Design procedures had also been presented for antenna engineers. However, the analysis methods were generally rough and the design procedures still relied on the previously obtained experimental data.

During the 1960's through the 2000's, the advances made in computer technology have had a major impact on the modern antenna technology. And they will undoubtedly have an even greater influence on antenna engineering in the twenty-first century. Since early 1960's, numerical methods have been introduced that can analyze and design waveguide slot antennas accurately. These methods included both low-frequency (e.g., method of moments [28]–[33], finite-difference [34, 35], finite-element [36, 37]) and high frequency (e.g., geometrical theories of diffraction [38]) techniques. These full-wave numerical methods made it possible to predict the performance of waveguide slot antennas with remarkable accuracy. In the 1980's, slots cut on an air-filled rectangular waveguide were thoroughly studied by several researchers. The mutual coupling effect in waveguide slot array antennas was also included in their calculations. These research works provided a guideline for the analysis and design of slot antennas cut on air-filled waveguides.

With the rapid progress of space applications, higher requirements have risen which need the antennas to be more compact as well as achieving a higher gain and

1.1 Motivation

lower sidelobes. Waveguide slot arrays are often adopted to meet these requirements. Longitudinal slots are generally used in the conventional air-filled rectangular waveguide. The slot elements should have an inter-element spacing of half a guided wavelength and be interlaced along the waveguide to avoid grating lobes. However, longitudinal slots also have some disadvantages. For a standing-wave longitudinal slot array, the array has a relatively small return loss bandwidth due to the rapid change of the inter-element spacing with frequency. For a traveling-wave longitudinal slot array, the main radiation beam is shifting when the operating frequency changes. Transverse slots in the broad wall of a rectangular waveguide are not prevalently used in array applications although their radiation patterns are usually symmetric and stably pointing to the broadside. This is because the inter-element spacing (one guided wavelength) for transverse slot array is much larger than that for longitudinal slots, which may result in grating lobes and larger array size.

Since the 1990's, a few researchers presented a means of inserting a dielectric slab in the rectangular waveguide, which provided a solution to the aforementioned problems. For slots cut on the broad wall of a waveguide partially loaded with an E-plane dielectric slab, the method of moments (MOM) was employed along with the Green's function in the waveguide [39, 40]. However, the computation is extremely time consuming even for modern computers due to the complexity of the Green's function of multi-layered structure. In addition, the electric field along the transverse slot is distorted due to the discontinuity near the slot, which may result in an asymmetric radiation pattern. In recent years, slots cut on the broad wall of a rectangular waveguide partially loaded with an H-plane dielectric slab have

1.2 Objectives

also been presented [41]–[43]. Simulations of the slot antenna using commercial software have been conducted and a small planar array have been successfully constructed. However, there still lacks theoretical analysis of the slot antennas cut on the H-plane dielectric-loaded rectangular waveguide.

From the above discussion, it is understood that although tremendous research works have been done on waveguide slot antennas, there is still plenty of room for making improvements, especially on dielectric-loaded waveguide slot antennas and arrays. This thesis aims to develop new techniques to analyze these problems, which is the motivation of this Ph.D thesis.

1.2 Objectives

In responding to the aforementioned problems, the goal of this thesis is to investigate dielectric-loaded waveguide slot antennas and design both slot elements and arrays for modern wireless communications. To fulfil the goal, the following tasks have been envisaged.

1. We begin our study with a comprehensive literature review of the structures, applications and analyzing methods of waveguide slot antennas. This exercise creates a strong foundation and background for further exploration.
2. It is noted that dielectric-loaded waveguide slot antennas have many advantages compared to the conventional air-filled waveguide slot antennas. When the waveguide is partially filled with a dielectric material, the dimension of the waveguide can be reduced and the guided wavelength can also be

1.2 Objectives

shortened, which can avoid grating lobes in array applications and realize a compact antenna array. We begin our study by developing an accurate and efficient method for analyzing different kinds of dielectric-loaded waveguide slot antennas.

3. Satisfied with the developed analyzing technique, we proceed to solve the whole waveguide slot structure. From the obtained scattering parameters of the waveguide slot antennas, we extract the circuit parameters of different slots to gain the necessary physical insight to design slot antennas. We also study the effect of the structural parameters on the characteristics of dielectric-loaded waveguide slot antennas. We also extend our study to the radiation performance of a single dielectric-loaded waveguide slot antenna element. The radiation patterns of slot antennas cut on both dielectric-loaded waveguide and conventional air-filled waveguide are studied and compared.
4. The mutual coupling effect is very important when we assemble an array using slot elements. Our work also extends to examine the mutual coupling between different slot antennas on the wall of a dielectric-loaded rectangular waveguide.
5. To design a dielectric-loaded waveguide slot antenna array, we combine the analyzed single slot with the mutual coupling effect between the antenna elements. We also design the inter-element spacing to achieve a high gain and low sidelobes by fabricating a uniformly excited waveguide slot array with non-uniform spacing.

The outcomes of the Ph.D study have produced some fundamental contributions and developments over the existing research works. They are outlined in the

1.3 Major Contributions of the Thesis

next section.

1.3 Major Contributions of the Thesis

In the Ph.D research project on dielectric-loaded waveguide slot antennas, six major contributions have been made. Some of them have been published in international journals and conferences (Please see author's publications on page 171).

The major contributions of the thesis are listed as follows:

1. We have carried out a thorough literature review of the previous research works on waveguide slot antennas, through which we conclude that waveguide slot antennas are playing important roles in the modern communication systems. We have found that slot antennas on the waveguide partially loaded with an H-plane dielectric slab have many advantages compared to conventional air-filled waveguide slot antennas.
2. We have developed a new hybrid full-wave method for analyzing slot antennas cut on both E- and H-plane dielectric-loaded waveguide. This hybrid method combines mode-matching (MM) and method of moments (MOM) to solve the slot antenna problem. The waveguide slot structure is decomposed into two parts: a dielectric-loaded waveguide T-junction and an open-ended waveguide radiating into half space. The T-junction is analyzed by the mode-matching method, while the open-ended waveguide is characterized by the moment method with the modal functions in the slot being the full domain basis functions. This proposed hybrid technique takes the advantage of the mode-matching method's high efficiency in analyzing the waveguide

1.3 Major Contributions of the Thesis

T-junction problem and the moment method's simplicity for the open-ended waveguide. It has a great advantage over the conventional moment method since the complicated Green's functions of a rectangular waveguide partially loaded with a dielectric slab is avoided.

3. In the formulation of the open-ended waveguide, we proposed a new approach for computing multidimensional integrals, which greatly reduces the computation effort. This new approach has been utilized to simplify the 4-dimensional (4-D) integral in the moment method analysis of the open-ended waveguide to a 1-D one, which has no singularity and can be efficiently calculated by Gaussian quadrature.
4. Parametrical studies have been conducted to investigate the effect of the dielectric slab's thickness and relative permittivity on slot antenna's circuit parameters. These studies provide us with the physical insight into the radiation of dielectric-loaded waveguide slots. Radiation characteristics of dielectric-loaded slot antennas have been examined. With the developed new hybrid technique, the whole slot structure has been solved. After that, we calculated the electric field distribution along the slot. The geometrical theory of diffraction (GTD) has been applied to evaluate the diffraction from all the waveguide wall edges. Radiation patterns were finally obtained by the vectorial summation of the geometrical optics (GO) field and the diffracted fields. The backward radiation was also included in our calculation. The effect of different sizes of waveguide wall on the radiation patterns of dielectric-loaded waveguide slot antennas are also studied. Good agreement is observed between the results calculated by our method and those obtained

1.3 Major Contributions of the Thesis

by commercial software (Ansoft's HFSS).

5. We have also studied the mutual coupling between two slot elements cut on the same dielectric-loaded rectangular waveguide. Each slot element has been decomposed into a dielectric-loaded waveguide T-junction and an open-ended waveguide. A four-port network has been composed in the T-junction cascading and the internal mutual coupling between the two slots was included automatically. The external mutual coupling was calculated by full-wave integration. The whole waveguide slot structure has been solved by combining the scattering parameters of the four-port network and the external mutual coupling.
6. A non-uniformly spaced 8-element slot array has been designed, fabricated and tested. The sidelobes were at least 15 dB lower than the mainlobe. Generally, a uniform antenna array with elements excited non-uniformly is employed to achieve desired radiation patterns and high main-to-side lobe ratio. However, in our study, we have found that simply adjusting the inter-element spacing can also provide similar radiation characteristics. The transverse slot cut on an H-plane dielectric-loaded rectangular waveguide has been chosen as the array element because it has a stable radiation pattern with mainlobe pointing to the broad side. After that, we use the standing wave slot element since it has a wider bandwidth than the traveling wave slot antenna. Because dielectric-loaded waveguide has a smaller size than conventional air-filled one, we were able to adjust the spacings of the slot elements and design the non-uniformly distributed 8-element slot antenna array. Experimental measurements show that the sidelobe level is as low as 15 dB, as we expected.

1.4 Organization of the Thesis

1.4 Organization of the Thesis

The remaining part of the thesis is organized as follows. Chapter 2 presents a literature review of the waveguide slot antennas. Firstly, the previous research works on conventional air-filled waveguide slot antennas are investigated. Theoretical and experimental studies on slot elements and arrays are described. Then we concentrate on the existing research on dielectric-loaded waveguide slot antennas in recent years. Recent progress in this area is also discussed.

We begin our study with analyzing different slot elements cut on the wall of a rectangular waveguide partially loaded with a dielectric slab. In Chapter 3, the analysis and formulation of partially dielectric-loaded waveguide slot antennas are given. A hybrid method for analyzing waveguide slot antennas is proposed, which is validated to be very accurate and efficient. After that, the circuit parameters of the slot antennas are extracted. Parametrical studies are also conducted to study the effect of the structural dimensions on the circuit model of dielectric-loaded waveguide slot antennas.

The radiation patterns of dielectric-loaded waveguide slot antennas are presented in Chapter 4. The radiation performances of both conventional air-filled waveguide slot antennas and dielectric-loaded waveguide slot antennas are presented. The effect of different ground plane size on the radiation characteristics of waveguide slot antennas is also examined.

Chapter 5 investigates the mutual coupling between waveguide slot elements.

1.4 Organization of the Thesis

The effect of the dimensions and orientations of the waveguide slot antenna element and the distances between them on the mutual coupling characteristic is examined in detail.

An 8-element dielectric-loaded waveguide slot antenna array is presented in Chapter 6. The slot antenna array is designed to achieve a low level of sidelobes. Non-uniform distribution of the elements that are excited uniformly is employed to meet our goal. The mutual coupling between slot elements is automatically taken into consideration. The designed slot array is fabricated and tested. Experimental results show that sidelobes below 15 dB are achieved, as we expected. Good agreement is also observed for the radiation patterns between measured data and those calculated by our method.

Finally, Chapter 7 concludes the work and gives recommendations for further research.

Chapter 2

Literature Review of Waveguide Slot Antennas

2.1 Introduction

Aperture antennas are very common at microwave frequencies. There are many different forms of an aperture antenna, such as slits, slots, open-ended waveguides and horns, etc. The apertures may be square, rectangular, circular, elliptical, or any other configuration, and can be covered with a dielectric material to protect them from environmental conditions. One kind of these antennas is the waveguide slot antenna, which is a slot radiator cut in the wall of a waveguide. Because waveguide slot antenna can be conveniently flush-mounted on the surface of the device, it is very useful for aircraft and spacecraft applications. It is also a suitable candidate for planar array applications with its great advantages of narrow-beam or shaped-beam radiation patterns, high efficiency and power capacity and light weight.

2.2 Air-filled Waveguide Slot Antennas

It is well known that the radiation pattern of a single waveguide slot is relatively wide and each element provides low value of directivity. Hence waveguide slot array antennas are usually employed to achieve very directive characteristics (very high gain) to meet the requirements of long distance communications. The arrangement of the array may be such that the summation of the radiation from all the elements gives a maximum in a particular direction or directions, minima in others, or otherwise as desired. Through proper configuration of the amplitude and phase of the excitation of individual element as well as the relative displacement between the elements, not only a high value of directivity but also scanning ability of the main radiation beam of an antenna system can be achieved theoretically. Therefore, waveguide slot antennas are ideally suited for modern requirements, involving reduced-height waveguides due to constraints in size and weight.

In this chapter, we first introduce the previous research works on conventional air-filled waveguide slot antennas. After that, recent studies on dielectric-loaded waveguide slot antennas are given. Finally, some of the numerical methods that have been employed in analyzing waveguide slot antennas are presented.

2.2 Air-filled Waveguide Slot Antennas

Waveguide-fed slot antennas have been widely investigated since the pioneer work of Stevenson [15] and Watson [13] during World War II. The early experimental work on waveguide slots for radiation and coupling applications was carried out by Watson [13]. In 1948, Stevenson [15] established the internal Green's functions for a rectangular waveguide and used the functions to analyze the scattering of

2.2 Air-filled Waveguide Slot Antennas

Figure 2.1: Air-filled rectangular waveguide slot antenna.

a longitudinal slot excited by an incident TE_{10} mode. In his work, he made the following assumptions: the waveguide was air-filled, the slot length was $\lambda_0/2$, the electric field in the slot was transverse and had equiphase half-cosinusoid amplitude distribution, the waveguide walls were ideally thin, and the broad wall containing the slot was imbedded in an infinite ground plane. With these assumptions, Stevenson developed integral equations for the electric field along the slot and showed that the slot could be represented by a shunt element on an equivalent transmission line and therefore, the dominant TE_{10} mode scattering by the slot was symmetrical. Using a power balance technique, he obtained a formula for the normalized resonant conductance of a single longitudinal slot as a function of its offset from the centerline of the rectangular waveguide. He attempted to generalize his result to model a non-resonant slot, but failed to find a similar expression for slot susceptance.

In 1951, Stegen [24] performed a series of experiments on longitudinal slot cut

2.2 Air-filled Waveguide Slot Antennas

in a broad wall of a standard X-band waveguide. He used round-ended slots, 0.063 inch wide, of various offsets and lengths, and measured relative backscattering at a frequency of 9.375 GHz. Stegen found that the admittance data of an isolated slot could be presented in a universal form that was extremely useful for computational purposes. He assembled the measured data with slot offset and length as parameters and established a pair of curves, which were in good agreement with the experimental data. His data served as the standard for 30 years and were reprinted in [44]. Additionally, Stegen found that resonant length, contrary to Stevenson's assumption, was offset dependent.

Working in parallel with Stegen, Oliner [26, 27] developed the first theory for longitudinal slots that would permit calculation of the susceptance as well as the conductance. He achieved this by formulating a stationary expression for reactive power and found good agreement with Stegen. Oliner's significant contributions were on the determination of the slot impedance or admittance away from a resonant frequency, based on the variational formulation, with the effects of wall thickness taken into account. The variational formulation has been improved in subsequent work [45, 46].

Khac and Carson [47]–[49] undertook the next important theoretical research. Under the assumption of a 'zero' wall thickness and purely transverse electric field in the slot, they used Stevenson's internal Green's function and the external half-space Green's function to formulate integral expressions for the internal and external scattered fields. This resulted in an integral equation involving the unknown slot aperture electric field, with the longitudinal component of the in-

2.2 Air-filled Waveguide Slot Antennas

cident TE_{10} mode being the driving function. Then the method of moments was employed, with the pulse function as the basis function and the point matching, to solve for the slot electric field. Other researchers had improved this approach later in [50]–[52].

After that, Elliott and others developed a design procedure for slot arrays [44, 53] in 1978 and improved it [54] in 1983. This design procedure also included external mutual coupling between all elements in the array [55], and internal higher-order-mode coupling through the TE_{20} mode [56]. It was shown that the TE_{20} mode accounted for a substantial part of the higher order mode coupling and that such coupling, though small, might be significant in high performance slot arrays. Briefly stated, the design procedure required theoretical or experimental data on the scattering properties of an isolated slot, external mutual coupling between slots, and internal higher-order-mode coupling between adjacent elements. Of all these parameters, the scattering characteristics of an isolated element were the most important and difficult to obtain. The method, employed in this design procedure, to obtain scattering characteristics of a single element was mainly from the work of Khac and Carson, which was mentioned in the preceding paragraph.

From 1980's, Rengarajan, based on Stevenson's Green's functions in conjunction with the method of moments (MOM), carried out a thorough investigation on waveguide longitudinal slots, transverse slots and inclined slots as well as the coupling junctions used in planar slot arrays, with the mutual coupling effects taken into consideration [28]–[33]. Extensive theoretical and numerical results were provided in his papers [28]–[33].

2.3 Dielectric-loaded Waveguide Slot Antennas

Figure 2.2: Dielectric-loaded rectangular waveguide slot antenna.

Most of these studies focused on rectangular slots cut on an air-filled waveguide. However, various shapes have also been used for the waveguide slot antennas to obtain the desired properties. For example, cross slot [57] has been employed to achieve a circularly polarized radiation pattern. Another example is T-slot opened in the wall of a rectangular waveguide [58] that can enhance the performance of the transverse slots.

2.3 Dielectric-loaded Waveguide Slot Antennas

Joubert and McNamara [39] studied the radiation performance of slots on the broad wall of a rectangular waveguide inhomogeneously loaded with a dielectric slab. In fact, as early as in 1983, Elliott presented a procedure for designing a dielectric-loaded small shunt slot array [54] with the external mutual coupling effects taken into consideration.

2.4 Analysis Methods

Recently, a slot array on the broad wall of a rectangular waveguide partially filled with a dielectric slab was presented [42]. The dielectric-loaded waveguide slot has many advantages compared to the conventional air-filled waveguide slot. When the waveguide is partially filled with a dielectric material, the dimensions of the waveguide can be reduced because the guided wavelength can be shortened, which can avoid grating lobes in array applications and realize a compact antenna array. Compared to the slot antenna in a waveguide fully filled with a dielectric material, the partially dielectric-filled slot structure can provide extra flexibilities to optimize the antenna performance so that a compact antenna array can be obtained with wide bandwidth.

2.4 Analysis Methods

In the previous studies, most of the researchers employed the internal waveguide Green's function and the external half-space Green's function to model the whole slot antenna structure. However, when the rectangular waveguide is partially loaded with a dielectric slab, the interior region of the waveguide turns into a multi-layered structure and the Green's function for the waveguide becomes very complicated and its numerical computation is extremely time consuming.

With the rapid progress in modern computers in the past twenty years, different full-wave methods, besides the conventional method of moments, have been employed to solve the waveguide slot problem, such as the finite-element method (FEM) [36, 37], the finite-difference time-domain method (FDTD) [34, 35], the mode matching (MM) method [59], the hybrid method [60], etc. Among these

2.4 Analysis Methods

methods, the FEM method and FDTD method are more popular than other methods. For example, Ansoft's HFSS is a mature software using the finite element method. XFDTD is another commercial software that can solve various 3-D structures using the FDTD method.

However, these full-wave methods normally require intensive computer memory and CPU time, especially for dielectric-loaded structures that are usually employed in the design to reduce the size of many microwave components. If the method of moments is adopted in analyzing slot antennas cut on a partially dielectric-loaded rectangular waveguide, the complicated internal Green's function is inevitable and the calculation will take a very long time. As for FEM or FDTD, in both cases the whole slot antenna structure will need to be divided into many meshes, which will be refined in several iterations to achieve a good accuracy. The major difference between FEM and FDTD is that one is in the frequency domain and the other is in the time domain. The mesh creation, refinement and the electromagnetic field calculation require a lot of CPU time for both methods. In order to record all the information of the solved structure for further use such as calculation of the radiation patterns, the field in each mesh must be saved in storage devices. Generally, these saved data are tremendous and occupy a large amount of disk spaces. Nevertheless, FEM and FDTD have their own advantages that they can deal with various types of structures, especially for irregular devices.

When considering slot antennas cut on a partially dielectric-loaded waveguide, the method of moments [39] has been employed for E-plane dielectric-loaded rectangular waveguide. However, the efficiency is not high enough to provide a di-

2.4 Analysis Methods

rection for real design process of such structures. Although a slot array cut on an H-plane dielectric-loaded rectangular waveguide has been designed and tested in [42] based on the simulation using commercial software, the accuracy and the efficiency of the design can hardly be guaranteed. For this reason, the design of waveguide slot antenna arrays for modern requirements is still a tough task. Thus, a rigorous and efficient method for analyzing waveguide slot antennas is urgently desired.

Chapter 3

S-parameters of Dielectric-loaded Waveguide Slot Elements

3.1 Introduction

The waveguide slot antenna has been widely used in high-performance radar and communication systems. It is also a suitable candidate for aircraft and spacecraft applications because it can be conveniently flush-mounted on their surfaces. Waveguide slots have been extensively investigated since World War II. Stevenson [15] established the internal Green's functions for a rectangular waveguide and formulated the general broad wall slot problem in the form of an integral equation. He also determined the resistance/conductance for longitudinal shunt slot, inclined series slot, and transverse series slot when the slots are at their first resonance. Oliner [26, 27] calculated the slot impedance/admittance based on a variational formulation taking the effect of finite wall thickness into account. His formulation was later improved in the subsequent work [46]. Recently, Rengarajan [28]–[32] solved

3.1 Introduction

the compound slot problem, which is more flexible and complicated. Although waveguide slots have been widely studied, most of the previous work focused on air-filled waveguide slots or homogeneously loaded waveguide slots. Joubert and McNamara [39] studied the radiation performance of slots on the broad wall of a rectangular waveguide inhomogeneously loaded with a dielectric slab. Elliott presented efficient procedures for designing both air-filled waveguide slot array [53] and dielectric-loaded small shunt slot array [54], with the external mutual coupling effects taken into consideration. In the previous studies, most of the researchers [29], [39], [49]–[52] employed the internal waveguide Green’s function and the external half-space Green’s function to model the slot antenna structure. However, the Green’s function in the internal region is complicated and its numerical computation is very time consuming when the rectangular waveguide is partially loaded with a dielectric slab.

Recently, a slot array on the broad wall of a rectangular waveguide partially filled with a dielectric slab was presented [42]. Dielectric-loaded waveguide slot has many advantages compared to the conventional air-filled waveguide slot. Compared to the slot antenna in a waveguide fully filled with a dielectric material, the partially dielectric-filled slot structure can provide extra flexibilities to optimize the antenna performance so that a compact antenna array can be obtained with wide bandwidth.

In the chapter we present a new hybrid technique combining the mode-matching method and moment method to analyze various slots cut in the wall of a rectangular waveguide partially filled with a dielectric slab. In our formulation, the original

3.1 Introduction

slot structure is divided into two parts: a rectangular waveguide T-junction partially filled with a dielectric slab and an open-ended waveguide radiating into the half space. Following the same procedure in [61], the dielectric-loaded waveguide T-junction is treated as a three-port network and its scattering parameters are formulated by the mode-matching method. The open-ended waveguide, which has been extensively studied by many researchers [62]–[67], is analyzed by the moment method. By applying Schelkunoff’s field equivalence principle [2], an equivalent magnetic current is placed on the end surface of the open-ended waveguide. The eigen modes in the slot are used as the full domain basis functions. A new analytical approach is proposed to simplify the 4-dimensional (4-D) integral in the moment method analysis of the open-ended waveguide to a 1-D one, which has no singularity and can be efficiently calculated by Gaussian quadrature [68]. The scattering parameters of the dielectric-loaded waveguide slot are then obtained by cascading the waveguide T-junction and the open-ended waveguide.

This proposed hybrid technique takes advantage of high efficiency of the mode-matching method in analyzing the waveguide T-junction problem and the simplicity of moment method for the open-ended waveguide. Numerical results for the scattering parameters and circuit parameters of various slots are compared to measured data and simulated ones obtained by HFSS. Good agreement is observed for all the cases considered. The effects of structural parameters on the slot’s circuit parameters are studied, which directly provides the physical insight into the radiation of dielectric-loaded waveguide slots.

The remaining parts of this chapter are organized as follows. Section 3.2 de-

3.2 Analysis Model

describes the problem geometry and the methodology that we use for characterizing a slot cut on the wall of a rectangular waveguide partially filled with a dielectric slab. Section 3.3 provides the formulation for the modes in a rectangular waveguide partially filled with a dielectric slab, the three-port T-junction, the open-ended waveguide and the cascading technique. Section 3.4 presents numerical results of scattering parameters and circuit parameters for various slots cut on the wall of a rectangular waveguide partially filled with a dielectric slab. The parametrical study is also performed. Conclusions are provided in Sec. 3.5.

3.2 Analysis Model

There are several different types of waveguide slots, such as transverse slot, longitudinal slot, center-inclined slot and compound slot. However, longitudinal and transverse slots are predominantly used in most practical applications. Therefore, our work focuses on these two kinds of slots, though the method described can be readily extended to compound slot and other slots.

3.2.1 Problem Geometry

A slot cut in the broad wall of a dielectric-loaded rectangular waveguide is illustrated in Fig. 3.1. The waveguide is of width a and height b . There are two layers of dielectric materials in the waveguide with the relative permittivities being ε_{r1} and ε_{r2} , respectively. The thickness of the lower layer is t . The slot cut in the wall of the waveguide is of width w and length l , respectively. The thickness of the waveguide wall is s . The displacement of the slot from the center of the waveguide is d . The reference plane T is located at the center of the slot.

3.2 Analysis Model

Figure 3.1: Dielectric-loaded rectangular waveguide slot.

It is known that a waveguide slot can be represented by a two-port network. Simplified circuit model is widely used, which consists of only one series impedance $Z = R + jX$ for transverse slot and one shunt admittance $Y = G + jB$ for longitudinal slot, respectively [26, 69]. In the thesis, a full-wave analysis is performed based on the hybrid mode-matching/moment method and the scattering parameters of the waveguide slot are then obtained, from which the circuit parameters can be readily extracted.

3.2.2 Methodology

Since the finite thickness of the waveguide wall has noticeable effect on the slot impedance/admittance, it should therefore be theoretically taken into account in our analysis. Figure 3.2(a) is the side view of the slot structure. The waveguide slot can be considered as a cascaded structure, which consists of a dielectric-loaded waveguide T-junction and an open-ended waveguide radiating into the half

3.3 Formulation

Figure 3.2: Decomposition of the dielectric-loaded waveguide slot; (a) physical structure, and (b) decomposed into two parts.

space, as illustrated in Fig. 3.2(b). The length of the rectangular side-arm of the T-junction is set to s , which is the thickness of the waveguide wall. The cross-sectional dimensions of the side-arm of the T-junction and the open-ended waveguide are equal to those of the original slot.

3.3 Formulation

In our hybrid mode-matching/moment method, the dielectric-loaded waveguide slot structure is decomposed into two parts: a three-port waveguide T-junction and an open-ended waveguide. The two parts are solved separately and then cascaded together to obtain the scattering parameters of the waveguide slot. In order to formulate the dielectric-loaded T-junction by the mode-matching method, the eigen modes in a waveguide partially filled with a dielectric slab are studied initially.

3.3 Formulation

3.3.1 Modes in a Waveguide Partially Filled with a Dielectric Slab

It is known that the eigen modes in a rectangular waveguide partially filled with dielectric slab can be derived from magnetic and electric types of Hertzian potential functions [70]. These derived modes are called longitudinal section electric (*LSE*) and longitudinal section magnetic (*LSM*) modes, which can be written in the following forms.

LSE modes:

$$\Psi_h = \begin{cases} C_h \cos \frac{n\pi x}{a} \sin(k_{y1}y), & 0 \leq y \leq t \\ D_h \cos \frac{n\pi x}{a} \sin[k_{y2}(b-y)], & t \leq y \leq b \end{cases} \quad (3.1)$$

LSM modes:

$$\Psi_e = \begin{cases} C_e \sin \frac{n\pi x}{a} \cos(k_{y1}y), & 0 \leq y \leq t \\ D_e \sin \frac{n\pi x}{a} \cos[k_{y2}(b-y)], & t \leq y \leq b \end{cases} \quad (3.2)$$

where $\Psi_{h,e}$ are the potential functions for *LSE* and *LSM* modes, respectively. $C_{h,e}$ and $D_{h,e}$ are the modal amplitude coefficients.

$$\begin{cases} k_{y1}^2 = k_0^2 \varepsilon_{r1} + \gamma^2 - \left(\frac{n\pi}{a}\right)^2 \\ k_{y2}^2 = k_0^2 \varepsilon_{r2} + \gamma^2 - \left(\frac{n\pi}{a}\right)^2 \end{cases} \quad (3.3)$$

where γ is the propagation constant, k_0 is the wave number in free space. The

3.3 Formulation

Figure 3.3: Cross-section of a dielectric-loaded rectangular waveguide.

electromagnetic fields in Fig. 3.3 can then be obtained from

$$\begin{aligned}\mathbf{E} &= -j\omega\mu_0 \sum_{n=1}^{N_h} A_n^h \nabla \times \mathbf{\Pi}_{nh} + \sum_{n=1}^{N_e} A_n^e \nabla \times \nabla \times \mathbf{\Pi}_{ne} \\ \mathbf{H} &= \sum_{n=1}^{N_h} A_n^h \nabla \times \nabla \times \mathbf{\Pi}_{nh} + j\omega\varepsilon_{1,2} \sum_{n=1}^{N_e} A_n^e \nabla \times \mathbf{\Pi}_{ne}\end{aligned}\quad (3.4)$$

where $\mathbf{\Pi}_h = \hat{y}\Psi_h(x, y)e^{-\gamma z}$, $\mathbf{\Pi}_e = \hat{y}\Psi_e(x, y)e^{-\gamma z}$. Enforcement of the boundary condition at the interface ($y = t$) results in the characteristic equations of *LSE* modes and *LSM* modes.

LSE modes:

$$k_{y2} \tan(k_{y1}t) + k_{y1} \tan[k_{y2}(b - t)] = 0; \quad (3.5)$$

LSM modes:

$$\varepsilon_{r2}k_{y1} \tan(k_{y1}t) + \varepsilon_{r1}k_{y2} \tan[k_{y2}(b - t)] = 0. \quad (3.6)$$

3.3 Formulation

Equations (3.5) and (3.6) can then be solved to determine the propagation constants of both *LSE* and *LSM* modes, which will be used in the mode-matching formulation for the dielectric-loaded waveguide T-junction.

In order to verify our formulas, an H-plane dielectric slab loaded rectangular waveguide is considered first. The conventional X-band waveguide WR-90 is loaded at the top with a 3 mm thick dielectric slab having a relative permittivity of 2.5. The phase and attenuation constants calculated using our program are compared with the data obtained by Ansoft's HFSS, as shown in Fig. 3.4. Excellent agreement between them is observed. It is also observed that the second propagation mode rises at 11.4 GHz, while the standard X-band waveguide have only one propagation mode in the frequency range considered. This shows that the operating frequency of the dielectric slab loaded waveguide will shift lower compared to the air-filled waveguide. In other words, the waveguide dimensions can be reduced by the dielectric loading. An E-plane dielectric slab loaded waveguide is also considered, as shown in Fig. 3.5. Again an excellent agreement is observed between our data and those obtained by HFSS, which validates that this method is also suitable for analyzing E-plane dielectric-loaded rectangular waveguides.

3.3.2 Three-port T-junction

Rectangular waveguide T-junction is a basic passive microwave component and has wide applications in waveguide directional couplers, power dividers/combiners, filters and multiplexers for modern radar and satellite communication systems. Compared to the conventional air-filled waveguide T-junction, dielectric-loaded configuration can obviously reduce the size of the T-junction and hence reduce the

3.3 Formulation

(a)

(b)

Figure 3.4: (a) Normalized phase constants β/k_0 , and (b) Attenuation constants α in an H-plane dielectric slab loaded waveguide ($a = 22.86 \text{ mm}$, $b = 10.16 \text{ mm}$, $t = 7.16 \text{ mm}$, $\epsilon_{r1} = 1.0$ and $\epsilon_{r2} = 2.5$).

3.3 Formulation

(a)

(b)

Figure 3.5: (a) Normalized phase constants β/k_0 , and (b) Attenuation constants α in an E-plane dielectric slab loaded waveguide ($a = 10.16 \text{ mm}$, $b = 22.86 \text{ mm}$, $t = 10 \text{ mm}$, $\epsilon_{r1} = 2.5$ and $\epsilon_{r2} = 1.0$).

3.3 Formulation

Figure 3.6: Waveguide T-junction loaded with a dielectric slab.

dimensions of various waveguide passive components. Dielectric-loaded waveguide T-junction can also provide some other useful characteristics compared to the conventional air-filled waveguide T-junction. For example, the power transmitting along a dielectric-loaded waveguide is basically confined to the interior of the dielectric slab. This can not only reduce the radiation loss but also offer a tight coupling between the main waveguide to the side arm of a waveguide T-junction when a dielectric slab is inserted into the part close to the side arm.

Figure 3.6 depicts the waveguide T-junction loaded with a dielectric slab and having a rectangular waveguide side arm. The cross-section of the side arm is the same as that of the slot. The dielectric-loaded waveguide T-junction, a three-port network, has been formulated in [61] by the mode-matching method.

3.3 Formulation

It is seen that the whole T-junction can be divided into four regions: 1, 2, 3 and 4, as shown in Fig. 3.6. Regions 1 and 2 are two semi-infinite dielectric-loaded rectangular waveguides, whose modal functions are obtained in Sec. 3.3.1. Region 3 is a semi-infinite homogeneous waveguide of rectangular cross-section with the modal function of the following form.

$$\begin{cases} \Psi_h = \cos \frac{m\pi(x - x_c)}{w} \cos \frac{n\pi(z - z_c)}{l}, & TE \\ \Psi_e = \sin \frac{m\pi(x - x_c)}{w} \sin \frac{n\pi(z - z_c)}{l}, & TM \end{cases} \quad (3.7)$$

where m and n are mode indices, w and l are width and height of the side arm waveguide.

Once all the electromagnetic field components in Regions 1, 2 and 3 are obtained, the resonant method [71] is employed to Region 4 and its electromagnetic fields can be expressed as vectorial summations of three standing wave solutions:

$$\begin{aligned} \mathbf{E}_4 &= \mathbf{E}_{41} + \mathbf{E}_{42} + \mathbf{E}_{43} \\ \mathbf{H}_4 &= \mathbf{H}_{41} + \mathbf{H}_{42} + \mathbf{H}_{43} \end{aligned} \quad (3.8)$$

where solutions $\mathbf{E}_{41}$ and $\mathbf{H}_{41}$ are obtained when the boundary planes S_2 and S_3 are short-circuited and S_1 is left open; solutions $\mathbf{E}_{42}$, $\mathbf{H}_{42}$ and $\mathbf{E}_{43}$, $\mathbf{H}_{43}$ are obtained in a similar way. It is seen that $\mathbf{E}_{41,42}$ and $\mathbf{H}_{41,42}$ can be derived from the modal functions of Regions 1 and 2. $\mathbf{E}_{43}$ and $\mathbf{H}_{43}$ can be derived from the following modal functions:

3.3 Formulation

$$\begin{cases} \Psi_h = \cos \frac{m\pi x}{a} \cos \frac{n\pi z}{c}, & TE \\ \Psi_e = \sin \frac{m\pi x}{a} \sin \frac{n\pi z}{c}, & TM \end{cases} \quad (3.9)$$

After the electromagnetic fields in all the four regions are expanded into the summation of a series of modal functions, we can then match the tangential electric and magnetic fields along the three regional interfaces S_1 , S_2 and S_3 . The relationship between the modal amplitude coefficients of the three waveguides (1, 2, 3) can be derived to obtain the generalized scattering matrix of the waveguide T-junction.

$$[\mathbf{S}_T] = \begin{bmatrix} \mathbf{S}_{11} & \mathbf{S}_{12} & \mathbf{S}_{13} \\ \mathbf{S}_{21} & \mathbf{S}_{22} & \mathbf{S}_{23} \\ \mathbf{S}_{31} & \mathbf{S}_{32} & \mathbf{S}_{33} \end{bmatrix} = \begin{bmatrix} \mathbf{A} & \mathbf{B} & \mathbf{C} \\ \mathbf{B} & \mathbf{A} & \mathbf{D} \\ \mathbf{E} & \mathbf{F} & \mathbf{G} \end{bmatrix}^{-1} - \mathbf{I}, \quad (3.10)$$

where the sub-matrices are expressed as [61]:

$$\mathbf{C} = \frac{1}{2} \mathbf{H}_1 \mathbf{M}_A \quad (3.11)$$

$$\mathbf{D} = -\frac{1}{2} \mathbf{H}_2 \mathbf{M}_A \quad (3.12)$$

$$\mathbf{E} = -\frac{1}{2} (\mathbf{W}_A - \mathbf{W}_B \mathbf{L}_1^2) (\mathbf{I} - \mathbf{L}_1^2)^{-1} \quad (3.13)$$

$$\mathbf{F} = \frac{1}{2} (\mathbf{W}_A - \mathbf{W}_B) \mathbf{L}_1 (\mathbf{I} - \mathbf{L}_1^2)^{-1} \quad (3.14)$$

$$\mathbf{G} = \frac{1}{2} (\mathbf{I} - \mathbf{Y}_3^{-1} \mathbf{M}_A^T \mathbf{Y}_{4,\text{II}} \mathbf{G}_P \mathbf{M}_A) \quad (3.15)$$

3.3 Formulation

with

$$A_{i,j} = \frac{1}{1 - e^{-2\gamma_{1i}c}} \delta_{i,j} \quad (3.16)$$

$$B_{i,j} = \frac{e^{-\gamma_{1i}c}}{e^{-2\gamma_{1i}c} - 1} \delta_{i,j} \quad (3.17)$$

$$L_{1i,j} = e^{-\gamma_{1i}c} \delta_{i,j} \quad (3.18)$$

$$H_{1,ik} = \int_{S_{1,I}} \mathbf{e}_{1it,I} \times [K_1 \hat{x} + K_2 \hat{y}] \cdot ds + \int_{S_{1,II}} \mathbf{e}_{1it,II} \times [K_3 \hat{x} + K_4 \hat{y}] \cdot ds \quad (3.19)$$

$$H_{2,ik} = (-1)^{n^{(4)}} H_{1,ik} \quad (3.20)$$

$$M_{A,kj} = \int_{S_3} (\mathbf{e}_{4kt} \cdot \mathbf{e}_{3jt}) \cdot ds = \int_{S_3} (\hat{y} \times \mathbf{e}_{4kt}) \cdot (\hat{y} \times \mathbf{e}_{3jt}) \cdot ds \quad (3.21)$$

$$W_{A,ji} = \frac{1}{Y_{3j}} \int_{S_3} \hat{y} \times \mathbf{e}_{3jt} \cdot e^{-\gamma_{1i}z} [h_{z1i,II} \hat{z} + h_{x1i,II} \hat{x}] \cdot ds \quad \text{at } y = b \quad (3.22)$$

$$W_{B,ji} = \frac{1}{Y_{3j}} \int_{S_3} \hat{y} \times \mathbf{e}_{3jt} \cdot e^{\gamma_{1i}z} [h_{z1i,II} \hat{z} - h_{x1i,II} \hat{x}] \cdot ds \quad \text{at } y = b \quad (3.23)$$

$$G_{p,k} = \left[1 - \frac{e^{-2\gamma_{4k,II}(t-b)} Q_{1,k}}{P_{1,k}} \right] / \left[1 + \frac{e^{-2\gamma_{4k,II}(t-b)} Q_{1,k}}{P_{1,k}} \right] \quad (3.24)$$

and

$$K_1 = \frac{-2 \cosh(\gamma_{4k,II}y) e^{-\gamma_{4k,II}(t-b)}}{P_1 \cdot G_S} h_{x4k,I} \quad (3.25)$$

$$K_2 = \frac{2 \sinh(\gamma_{4k,II}y) e^{-\gamma_{4k,II}(t-b)}}{P_1 \cdot G_S} h_{y4k,I} \quad (3.26)$$

$$K_3 = \frac{e^{-\gamma_{4k,II}(y-b)} - \frac{e^{-2\gamma_{4k,II}(t-b)} Q_1}{P_1} e^{\gamma_{4k,II}(y-b)}}{G_S} h_{x4k,II} \quad (3.27)$$

$$K_4 = \frac{e^{-\gamma_{4k,II}(y-b)} + \frac{e^{-2\gamma_{4k,II}(t-b)} Q_1}{P_1} e^{\gamma_{4k,II}(y-b)}}{G_S} h_{y4k,II} \quad (3.28)$$

3.3 Formulation

$$P_{1,k} = \sinh(\gamma_{4k,I}t) - \cosh(\gamma_{4k,I}t) \frac{Y_{4k,I}}{Y_{4k,II}} \quad (3.29)$$

$$Q_{1,k} = \sinh(\gamma_{4k,I}t) + \cosh(\gamma_{4k,I}t) \frac{Y_{4k,I}}{Y_{4k,II}} \quad (3.30)$$

$$G_{S,k} = 1 + \frac{e^{-2\gamma_{4k,II}(t-b)} Q_{1,k}}{P_{1,k}} \quad (3.31)$$

$Y_{4(I,II)}$ are the admittance matrices of Layers I and II for Region 4, respectively. $\delta_{i,j}$ is the Kronecker delta, which is defined as $\delta_{i,j} = 0$ for $i \neq j$ and $\delta_{i,j} = 1$ for $i = j$. The matrix $\mathbf{M}_A$ is the E-field mode-matching matrix at interface S_3 and the superscript “T” denotes the transpose operation. $\mathbf{e}_{3jt}$ and $\mathbf{e}_{4jt}$ are the normalized transverse components of the electric field for the j th mode in Regions 3 and 4, respectively. $\mathbf{e}_{1it,(I,II)}$ is the normalized transverse component of the electric field for the i th mode in Region 1(I, II) and Region 2(I, II).

Matlab codes are developed to calculate the scattering parameters of waveguide T-junction structures. In order to verify the validity of our program, an air-filled rectangular waveguide with the side arm being rectangular is calculated first. The main waveguide is the standard WR-137 and the side arm is WR-90. Therefore, the dimensions shown in Fig. 3.6 are $a = 34.85mm$, $b = 15.799mm$, $w = 22.86mm$, $l = 10.16mm$. The side arm is located at the center of the broad wall of the main waveguide. By setting the relative permittivities ϵ_{r1} and ϵ_{r2} to 1 for both Layers I and II (Fig. 3.3), the dielectric-loaded waveguide T-junction reduces to the conventional air-filled waveguide T-junction. Generally, the number of modes used in the mode-matching method depend on the size of the waveguide. The larger the size of the cross section, the more modes are required to achieve a good accuracy. In our analysis, the number of modes employed in Regions 1, 2, 3 and

3.3 Formulation

4 are 50, 50, 30 and 200, respectively. The number of modes for Regions 1 and 2 are identical because they have same cross sections. The relative error of $|S_{11}|$ in the frequency range of 5 to 11 GHz is below 0.001. The scattering parameters obtained by our code are compared to those obtained by HFSS, as illustrated in Fig. 3.7. Good agreement is observed between them. It is noticed that the cutoff frequency for the first mode in the side arm is at 6.56 GHz. The performance of the T-junction below and above this frequency is different. The scattering parameters S_{11} and S_{21} are given in Fig. 3.7(a) and S_{31} and S_{33} are given in Fig. 3.7(b) for frequencies above the cutoff frequency of Port 3. It can be observed from Fig. 3.7 that there is some difference between the results when the frequency is above 10 GHz. This is because some higher order modes in the main waveguide start to propagate above 10 GHz. It should be mentioned that HFSS failed to precisely characterize the performance at these frequencies. In order to meet high accuracy criteria, different settings must be used and the computation using HFSS will require a very long time.

An example of a waveguide T-junction of the same dimensions partially filled with a dielectric slab in the main waveguide is considered. The main waveguide is loaded with an H-plane dielectric slab, as shown in Fig. 3.8. The relative permittivity for the dielectric slab used in Fig. 3.8 is 2.5 and the thickness of the dielectric slab is $(b - t) = 8.299mm$. In our study, the number of modes are still 50, 30 and 200 for Regions 1, 3 and 4, respectively. The CPU time of our method is less than 1 second for each frequency point, while the computation using HFSS needs 3 minutes (7 iterations, 3358 tetrahedra). From Figs. 3.7 and 3.8, it is observed that the behavior around the cutoff frequency of the first mode in the

3.3 Formulation

(a)

(b)

Figure 3.7: Scattering parameters of an air-filled rectangular waveguide T-junction with a side arm of reduced sized waveguide ($a = 34.85 \text{ mm}$, $b = 15.799 \text{ mm}$, $w = 22.86 \text{ mm}$, $l = 10.16 \text{ mm}$).

3.3 Formulation

(a)

(b)

Figure 3.8: Scattering parameters of a dielectric-loaded rectangular waveguide T-junction with a side arm of reduced size waveguide ($a = 34.85 \text{ mm}$, $b = 15.799 \text{ mm}$, $w = 22.86 \text{ mm}$, $l = 10.16 \text{ mm}$, $t = 7.5 \text{ mm}$, $\epsilon_{r1} = 1$, $\epsilon_{r1} = 2.5$).

3.3 Formulation

Figure 3.9: Relative error for $|S_{11}|$ of a dielectric-loaded waveguide T-junction (Dimensional parameters are the same as in Fig. 3.8).

side arm (about 6.56 GHz) is greatly altered after inserting the dielectric slab and the peak of the reflection coefficient at port 1 is shifted lower. It is also seen that the agreement between our data and those obtained by HFSS is very good, as we expected.

In order to show the numerical convergence of our mode-matching method, we have varied the mode numbers N_1 (for Region 1) and N_3 (for Region 3) used in the previous example to examine the change of scattering parameters. The maximum relative error of $|S_{11}|$ in the frequency range of 4 to 7.5 GHz is plotted in Fig. 3.9.

3.3 Formulation

$\Delta|S_{11}|$ is defined as:

$$|S_{11}| = \frac{|S_{11}| - |S_{11}^R|}{|S_{11}^R|} \quad (3.32)$$

where $|S_{11}^R|$ is obtained by setting $N_1 = 200$ and $N_3 = 100$ and considered as the reference data. The number of modes in Region 4 is fixed to 200 because it is the key section of the discontinuity and many higher-order modes exist in this region. It can be seen that $|S_{11}|$ converges rapidly when the mode number increases. If the mode numbers N_1 and N_3 are chosen as 50 and 30, the relative error is less than 0.001.

3.3.3 Open-ended Waveguide

Figure 3.10(a) shows the side-view of an open-ended rectangular waveguide radiating into the half space. Applying Schelkunoff's field equivalence principle [2], as shown in Fig. 3.10(b), an equivalent magnetic current $\mathbf{M}_s$ is placed at the open end surface with the waveguide short-circuited end. The equivalent magnetic current is $\mathbf{M}_s = -\mathbf{n} \times \mathbf{E}_{3t}$ with $\mathbf{n}$ being the normal unit vector.

It can be seen that $\mathbf{E}_{3t}$ in Fig. 3.10 is actually the electric field in the side-arm of the T-junction in Fig. 3.6. Therefore, the magnetic current $\mathbf{M}_s$ can be written as:

$$\mathbf{M}_s = - \sum_{j=1}^M [A_{3j}^- e^{-\gamma_{3j}s} + A_{3j}^+ e^{+\gamma_{3j}s}] \mathbf{T}_j, \quad (3.33)$$

where the superscript M is the number of the modes used in the side-arm wave-

3.3 Formulation

Figure 3.10: The field equivalent principle applied for the open-ended waveguide; (a) open-ended waveguide, and (b) equivalent problem.

guide and $\mathbf{T}_j$ has the following form:

$$\mathbf{T}_j = \begin{cases} -\nabla_t N_j \cos \frac{m\pi(x-x_c)}{w} \cos \frac{n\pi(z-z_c)}{l}, & TE \\ \hat{y} \times \nabla_t N_j \sin \frac{m\pi(x-x_c)}{w} \sin \frac{n\pi(z-z_c)}{l}, & TM \end{cases} \quad (3.34)$$

where ∇_t is the 2-D differential operator having the form $\nabla_t = \hat{x} \frac{\partial}{\partial x} + \hat{z} \frac{\partial}{\partial z}$; N_j is the normalization factor and j is the mode index used in the field expansion and stands for the combination of sub indices m and n ; x_c , z_c , w and l are dimensional parameters shown in Fig. 3.6.

3.3 Formulation

In the half free-space, the magnetic field due to the equivalent magnetic current $\mathbf{M}_s$ is [72]:

$$\mathbf{H}_{\text{out}} = -j\omega\mathbf{F} - \nabla\phi_m, \quad (3.35)$$

where $\mathbf{F}$ and ϕ_m are the electric vector potential and the magnetic scalar potential, respectively.

$$\mathbf{F} = \varepsilon_0 \iint_{S_w} \mathbf{M}_s(\mathbf{r}') G(\mathbf{r}|\mathbf{r}') dS', \quad (3.36)$$

$$\phi_m = \frac{j}{\omega\mu_0\varepsilon_0} \nabla \cdot \mathbf{F}, \quad (3.37)$$

where S_w is the end surface of the waveguide and $G(\mathbf{r}|\mathbf{r}')$ is the scalar Green's function in the half-space [70].

$$G(\mathbf{r}|\mathbf{r}') = \frac{e^{-jk_0R}}{2\pi R}, \quad (3.38)$$

where $R = |\mathbf{r} - \mathbf{r}'|$ is the distance between the field point $\mathbf{r}$ and the source point $\mathbf{r}'$ that is located on S_w . After some manipulations, we can have the following expression for the magnetic field $\mathbf{H}_{\text{out}}$:

$$\begin{aligned} \mathbf{H}_{\text{out}} = & \frac{j}{2\pi\omega\mu_0} \sum_{j=1}^M [A_{3j}^- e^{-\gamma_{3j}s} + A_{3j}^+ e^{+\gamma_{3j}s}] \\ & \times \left\{ k_0^2 \iint_{S_w} \mathbf{T}_j(\mathbf{r}') \frac{e^{-jk_0R}}{R} dS' \right. \\ & \left. + \nabla \iint_{S_w} [\nabla' \cdot \mathbf{T}_j(\mathbf{r}')] \frac{e^{-jk_0R}}{R} dS' \right\}, \end{aligned} \quad (3.39)$$

Enforcing the boundary condition for the tangential magnetic field components

3.3 Formulation

at surface S_w yields

$$\mathbf{H}_{3t} = \mathbf{H}_{\text{out}}, \quad (3.40)$$

where $\mathbf{H}_{3t}$ is the magnetic field in the side-arm of the T-junction and $\mathbf{H}_{3t} = \sum_{j=1}^M [A_{3j}^- e^{-\gamma_{3j}s} - A_{3j}^+ e^{+\gamma_{3j}s}] \mathbf{T}_j$. Using $\mathbf{T}_i$ as the weighting function and performing the testing procedure, i.e. dot multiplying $\mathbf{T}_i(\mathbf{r})$ at both sides of Eq.(3.40) and integrating the equation on the source surface S_w , we obtain:

$$\begin{aligned} A_{3i}^- e^{-\gamma_{3i}s} - A_{3i}^+ e^{+\gamma_{3i}s} &= \iint_{S_w} \mathbf{H}_{\text{out}} \cdot \mathbf{T}_i dS; \\ i &= 1, 2 \cdots M. \end{aligned} \quad (3.41)$$

After some transformations, the relation between $[\mathbf{A}_3^-]$ and $[\mathbf{A}_3^+]$ can be written as:

$$[\mathbf{u}] [\mathbf{A}_3^-] = [\mathbf{v}] [\mathbf{A}_3^+], \quad (3.42)$$

where

$$\begin{cases} u_{ij} = e^{-\gamma_{3i}s} \delta_{ij} - \frac{j e^{-\gamma_{3j}s}}{2\pi\omega\mu_0} [k_0^2 B_{ij} - C_{ij}] \\ v_{ij} = e^{+\gamma_{3i}s} \delta_{ij} + \frac{j e^{+\gamma_{3j}s}}{2\pi\omega\mu_0} [k_0^2 B_{ij} - C_{ij}] \end{cases} \quad (3.43)$$

with

$$B_{ij} = \iint_{S_w} [\mathbf{T}_i(\mathbf{r}) \cdot \iint_{S_w} \mathbf{T}_j(\mathbf{r}') \frac{e^{-jk_0 R}}{R} dS'] dS \quad (3.44)$$

$$C_{ij} = \iint_{S_w} [\nabla \cdot \mathbf{T}_i(\mathbf{r})] \left[\iint_{S_w} [\nabla' \cdot \mathbf{T}_j(\mathbf{r}')] \frac{e^{-jk_0 R}}{R} dS' \right] dS \quad (3.45)$$

From Eq.(3.42) we obtain $[\mathbf{A}_3^+] = [\mathbf{R}] [\mathbf{A}_3^-]$, where $[\mathbf{R}] = [\mathbf{v}]^{-1} [\mathbf{u}]$ is the reflection coefficient matrix of the open-ended waveguide.

3.3 Formulation

In order to obtain the numerical results for the reflection coefficient matrix, the 4-dimensional (4-D) integrals B_{ij} and C_{ij} must be calculated. From Eq.(3.34), we have the following expression for the divergence of $\mathbf{T}_i$:

$$\nabla \cdot \mathbf{T}_i = \begin{cases} N_i k_c \cos \frac{m\pi(x-x_c)}{w} \cos \frac{n\pi(z-z_c)}{l}, & TE \\ 0, & TM \end{cases} \quad (3.46)$$

where $k_c = \left(\frac{m\pi}{w}\right)^2 + \left(\frac{n\pi}{l}\right)^2$.

Substituting Eqs.(3.34) and (3.46) into Eqs.(3.44) and (3.45), B_{ij} and C_{ij} can be rewritten as:

$$B_{ij} = N_i N_j \begin{cases} \frac{m\pi}{w} \frac{m'\pi}{w} I_A + \frac{n\pi}{l} \frac{n'\pi}{l} I_B, & TE - TE \\ \frac{m\pi}{w} \frac{n'\pi}{l} I_A - \frac{n\pi}{l} \frac{m'\pi}{w} I_B, & TE - TM \\ \frac{n\pi}{l} \frac{m'\pi}{w} I_A - \frac{m\pi}{w} \frac{n'\pi}{l} I_B, & TM - TE \\ \frac{n\pi}{l} \frac{n'\pi}{l} I_A + \frac{m\pi}{w} \frac{m'\pi}{w} I_B, & TM - TM \end{cases} \quad (3.47)$$

and

$$C_{ij} = \begin{cases} N_i N_j k_c k'_c I_C, & \begin{cases} i(m, n) - TE \\ j(m', n') - TE \end{cases} \\ 0, & otherwise \end{cases} \quad (3.48)$$

3.3 Formulation

where I_A , I_B and I_C are 4-D integrals having the following forms:

$$\begin{aligned}
 I_A &= \int_0^w \int_0^w \int_0^l \int_0^l dx \, dx' \, dz \, dz' \frac{e^{-jk_0 R}}{R} \\
 I_B &= \int_0^w \int_0^w \int_0^l \int_0^l dx \, dx' \, dz \, dz' \frac{e^{-jk_0 R}}{R} \\
 I_C &= \int_0^w \int_0^w \int_0^l \int_0^l dx \, dx' \, dz \, dz' \frac{e^{-jk_0 R}}{R} \\
 &\quad \cdot \frac{\sin\left(\frac{m\pi x}{w}\right)}{\cos\left(\frac{m\pi x}{w}\right)} \cdot \frac{\sin\left(\frac{m'\pi x'}{w}\right)}{\cos\left(\frac{m'\pi x'}{w}\right)} \cdot \frac{\cos\left(\frac{n\pi z}{l}\right)}{\sin\left(\frac{n\pi z}{l}\right)} \cdot \frac{\cos\left(\frac{n'\pi z'}{l}\right)}{\sin\left(\frac{n'\pi z'}{l}\right)} \quad (3.49)
 \end{aligned}$$

with $R = \sqrt{(x - x')^2 + (z - z')^2}$ being the distance between the source point and the field point.

It is noticed that the 4-D integrals I_A , I_B and I_C contain singular points in the range of the integration. The general method to overcome this problem is to subtract the singularities from the integrand and numerically compute the remaining parts of the integral. However, the calculation of the integral will be computationally expensive due to its slow convergence. In our formulation, a new approach is proposed to reduce the burden in the numerical computation. It is obvious that the 4-D integrals contain only a first order pole in the range of the integral. Therefore, it is possible to evaluate the integrals analytically or semi-analytically.

Taking I_A as an example, the following variable transformations are made:

$$\begin{cases} x = \frac{u_1 + u_2}{2} \\ x' = \frac{u_1 - u_2}{2} \end{cases}, \quad \begin{cases} z = \frac{v_1 + v_2}{2} \\ z' = \frac{v_1 - v_2}{2} \end{cases}, \quad (3.50)$$

3.3 Formulation

Figure 3.11: Mapping of the range of the 4-D integral.

and the range of the integral is therefore mapped as shown in Fig. 3.11. Accordingly, the integrand in I_A then becomes:

$$\begin{aligned} & \frac{1}{4} \frac{e^{-jk_0 \sqrt{u_2^2 + v_2^2}}}{\sqrt{u_2^2 + v_2^2}} \sin \frac{m\pi (u_1 + u_2)}{2w} \sin \frac{m'\pi (u_1 - u_2)}{2w} \\ & \cdot \cos \frac{n\pi (v_1 + v_2)}{2l} \cos \frac{n'\pi (v_1 - v_2)}{2l}. \end{aligned} \quad (3.51)$$

3.3 Formulation

The integrand can be further expressed as:

$$\begin{aligned}
 & -\frac{1}{16} \frac{e^{-jk_0 \sqrt{u_2^2 + v_2^2}}}{\sqrt{u_2^2 + v_2^2}} \\
 & \cdot \left[\cos(m^+ u_1 + m^- u_2) \cos(n^+ v_1 + n^- v_2) \right. \\
 & + \cos(m^+ u_1 + m^- u_2) \cos(n^- v_1 + n^+ v_2) \\
 & - \cos(m^- u_1 + m^+ u_2) \cos(n^+ v_1 + n^- v_2) \\
 & \left. - \cos(m^- u_1 + m^+ u_2) \cos(n^- v_1 + n^+ v_2) \right], \quad (3.52)
 \end{aligned}$$

where the new coefficients are $m^+ = \frac{(m + m') \pi}{2w}$, $m^- = \frac{(m - m') \pi}{2w}$, $n^+ = \frac{(n + n') \pi}{2l}$ and $n^- = \frac{(n - n') \pi}{2l}$, respectively

It is found that in Eq.(3.52) only one general integral needs to be computed:

$$\begin{aligned}
 I_G = & \int_{-w}^w du_2 \int_{|u_2|}^{2w-|u_2|} du_1 \int_{-l}^l dv_2 \int_{|v_2|}^{2l-|v_2|} dv_1 \\
 & \cdot \frac{e^{-jk_0 \sqrt{u_2^2 + v_2^2}}}{\sqrt{u_2^2 + v_2^2}} \cos(p_1 u_1 + p_2 u_2) \cos(q_1 v_1 + q_2 v_2), \quad (3.53)
 \end{aligned}$$

where p_1 , p_2 , q_1 and q_2 stand for known coefficients. Since the integrand in Eq.(3.53) is smooth with respect to u_1 and v_1 , we can analytically integrate the 2-D integral about u_1 and v_1 .

$$\begin{aligned}
 I_G = & \frac{1}{p_1 q_1} \int_{-w}^w du_2 \int_{-l}^l dv_2 \frac{e^{-jk_0 \sqrt{u_2^2 + v_2^2}}}{\sqrt{u_2^2 + v_2^2}} \\
 & \cdot [\sin(p_2 u_2 - p_1 |u_2| + 2wp_1) - \sin(p_1 |u_2| + p_2 u_2)] \\
 & \cdot [\sin(q_2 v_2 - q_1 |v_2| + 2lq_1) - \sin(q_1 |v_2| + q_2 v_2)]. \quad (3.54)
 \end{aligned}$$

3.3 Formulation

By making the following variable substitutions,

$$\begin{cases} u_2 = \rho \sin \varphi \\ v_2 = \rho \cos \varphi \end{cases} \quad (3.55)$$

the remaining 2-D integral with respect to u_2 and v_2 is then transformed into polar coordinates with the following form:

$$\begin{aligned} I_G = & \frac{1}{p_1 q_1} \iint d\rho d\varphi e^{-jk_0 \rho} \\ & \cdot [\sin(p_2 \rho \sin \varphi - p_1 |\rho \sin \varphi| + 2wp_1) - \sin(p_1 |\rho \sin \varphi| + p_2 \rho \sin \varphi)] \\ & \cdot [\sin(q_2 \rho \cos \varphi - q_1 |\rho \cos \varphi| + 2lq_1) - \sin(q_1 |\rho \cos \varphi| + q_2 \rho \cos \varphi)]. \end{aligned} \quad (3.56)$$

It is noted that the first order pole in Eq.(3.54) is then eliminated in Eq.(3.56). The integral range of I_G in the polar coordinate can also be readily obtained. The integral on ρ in Eq.(3.56) can be analytically worked out and I_G can therefore be obtained by calculating the remaining 1-D integral with respect to φ using Gaussian quadrature [68]. Integrals I_B and I_C are then computed in the same manner as I_A . It is seen that this approach only needs numerical calculation of 1-D integrals, which contain no singular points and can therefore be very efficiently computed. Similar attempt was reported and utilized implicitly [73]–[75], but it was not known to the author before our work.

The example of an open-ended rectangular waveguide (WR-90: 22.86 mm × 10.16 mm) is considered here for comparison. The magnitude and phase of the

3.3 Formulation

(a)

(b)

Figure 3.12: Reflection coefficient of an open-ended rectangular waveguide (WR-90: $22.86\text{mm} \times 10.16\text{mm}$); (a) magnitude, and (b) phase.

3.3 Formulation

reflection coefficient are shown in Fig. 3.12(a) and Fig. 3.12(b), respectively. The reflection coefficients obtained by our method are compared to theoretical results given in [67]. Excellent agreement between them is observed. As described in the previous section, the calculation of the reflection coefficient depends on the computation of 1-D integrals using Gaussian quadrature. Numerical tests show that only ten points in Gaussian quadrature are required in computing the 1-D integrals, which makes our method very efficient. It can also be seen in Fig. 3.12 that six modes [$M = 6$ in Eq.(3.33); they are TE_{10} , TE_{20} , TE_{01} , TE_{11} , TM_{11} , TE_{30} modes] are enough to achieve accurate results for both magnitude and phase of the reflection coefficient for an open-ended rectangular waveguide, since there is only one propagating mode in the frequency range considered.

3.3.4 Cascading Technique

After the waveguide T-junction and the open-ended waveguide are formulated, the slot problem can be solved by cascading these two blocks together. As described in Sec.3.3.3, relation $[\mathbf{A}_3^+] = [\mathbf{R}] [\mathbf{A}_3^-]$ is obtained for $\mathbf{A}_3^+$ and $\mathbf{A}_3^-$ in Fig. 3.6, where $[\mathbf{R}]$ is the reflection coefficient matrix of the open-ended waveguide. Substituting this relation into the equation:

$$\begin{bmatrix} \mathbf{A}_1^- \\ \mathbf{A}_2^- \\ \mathbf{A}_3^- \end{bmatrix} = [\mathbf{S}_T] [\mathbf{A}^+] = \begin{bmatrix} \mathbf{S}_{11} & \mathbf{S}_{12} & \mathbf{S}_{13} \\ \mathbf{S}_{21} & \mathbf{S}_{22} & \mathbf{S}_{23} \\ \mathbf{S}_{31} & \mathbf{S}_{32} & \mathbf{S}_{33} \end{bmatrix} \begin{bmatrix} \mathbf{A}_1^+ \\ \mathbf{A}_2^+ \\ \mathbf{A}_3^+ \end{bmatrix} \quad (3.57)$$

where the generalized scattering matrix $[\mathbf{S}_T]$ is derived in Sec.3.3.2, the scattering

3.4 Numerical Results

matrix of the waveguide slot $[\mathbf{S}_{\text{slot}}]$ is then obtained.

$$\begin{bmatrix} \mathbf{A}_1^- \\ \mathbf{A}_2^- \end{bmatrix} = [\mathbf{S}_{\text{slot}}] \begin{bmatrix} \mathbf{A}_1^+ \\ \mathbf{A}_2^+ \end{bmatrix} = \begin{bmatrix} \mathbf{S}_{11}^s & \mathbf{S}_{12}^s \\ \mathbf{S}_{21}^s & \mathbf{S}_{22}^s \end{bmatrix} \begin{bmatrix} \mathbf{A}_1^+ \\ \mathbf{A}_2^+ \end{bmatrix} \quad (3.58)$$

where sub-matrices of $[\mathbf{S}_{\text{slot}}]$ can be expressed as:

$$\begin{cases} \mathbf{S}_{11}^s = \mathbf{S}_{11} + \mathbf{S}_{13} \mathbf{R}_s \mathbf{S}_{31} \\ \mathbf{S}_{12}^s = \mathbf{S}_{12} + \mathbf{S}_{13} \mathbf{R}_s \mathbf{S}_{32} \\ \mathbf{S}_{21}^s = \mathbf{S}_{21} + \mathbf{S}_{23} \mathbf{R}_s \mathbf{S}_{31} \\ \mathbf{S}_{22}^s = \mathbf{S}_{22} + \mathbf{S}_{23} \mathbf{R}_s \mathbf{S}_{32} \end{cases} \quad (3.59)$$

with $\mathbf{R}_s = (\mathbf{R}^{-1} - \mathbf{S}_{33})^{-1}$.

3.4 Numerical Results

Based on the formulation described in the previous section, a computer program has been developed to calculate the scattering parameters of various waveguide slots. In order to verify our hybrid technique for analyzing waveguide slots, longitudinal and transverse slots cut in the broad wall of an air-filled waveguide are analyzed and compared to available simulated or measured data in the literature. Longitudinal and transverse slots in a dielectric-loaded waveguide are also studied and the results obtained by our method are then compared with measured data and those obtained by Ansoft's HFSS. Good agreement is achieved in all the cases considered. After having verified the validity of our hybrid technique, parametrical studies of the dielectric-loaded slot problem are performed to gain insight into its

3.4 Numerical Results

Figure 3.13: Frequency dependence of normalized series resistance and reactance for a centered transverse slot in an air-filled waveguide ($a = 22.86 \text{ mm}$, $b = 10.16 \text{ mm}$, $w = 14 \text{ mm}$, $l = 3 \text{ mm}$, $s = 1 \text{ mm}$).

radiation performance.

3.4.1 Air-filled Waveguide Slots

Transverse slot cut in the broad wall of an air-filled waveguide is studied and the normalized series impedance is plotted in Fig. 3.13. In our program, by setting the relative permittivities ϵ_{r1} and ϵ_{r2} (in Fig. 3.6) to 1.0 for both Layer I and Layer II, the dielectric-loaded T-junction will reduce to the conventional air-filled waveguide T-junction. Numerical results for the normalized series resistance and reactance obtained by our method are compared to those obtained by HFSS.

3.4 Numerical Results

Excellent agreement between them is observed. In our method, the numbers of the modes employed in Regions 1, 3 and 4 of Fig. 3.6 are 50, 30 and 200, respectively. Because the slot is located at the center of the waveguide wall and the whole structure is symmetric, the field distribution along the slot is dominated by TE_{10} mode. Therefore, only seven modes (TE_{10} , TE_{30} , TE_{50} , TE_{70} , TE_{90} , TE_{12} and TM_{12}) out of the 30 modes are used in computing the scattering of the dominant mode in the open-ended waveguide. Generally, the number of modes required to achieve a good accuracy depends on the size and shape of the slot. For a rectangular slot, the larger the slot's cross-section, the more modes are needed. It is also noted that good results may be obtained using only a few modes if the aspect ratio of the slot is big.

Another example of a longitudinal slot in an air-filled waveguide is considered in Fig. 3.14. The slot has a width of $1.5875mm$ and an off-centered displacement of $1.8796mm$. The waveguide is of dimensions $22.86\text{ mm} \times 10.16\text{ mm}$. The thickness of the wall is $1mm$. The mode numbers used in this longitudinal slot are the same as those in the previous example. Numerical results obtained by our method are compared with Oliner's theoretical data [27] together with Stegen's measured data [24]. Simulated data by HFSS are also presented. The agreement between our results and HFSS data is excellent. It is noticed that there is a small discrepancy between the experimental data and our results. This is because the measurements were taken on slots with rounded ends [24]. The slight difference between our results and those by Oliner is due to the fact that only the dominant mode in the waveguide and in the slot was included in Oliner's variational method.

3.4 Numerical Results

(a)

(b)

Figure 3.14: Normalized shunt (a) conductance and (b) susceptance versus the length of the slot for a longitudinal slot in an air-filled waveguide ($a = 22.86\text{mm}$, $b = 10.16\text{mm}$, $w = 1.5875\text{mm}$, $d = 1.8796\text{mm}$, $s = 1.27\text{mm}$, frequency is 9.375GHz).

3.4 Numerical Results

Figure 3.15: Scattering parameters of a transverse slot in a rectangular waveguide partially filled with a dielectric slab ($a = 15mm$, $b = 6mm$, $t = 3mm$, $w = 15mm$, $l = 6mm$, $s = 1mm$, $\epsilon_{r1} = 6$, $\epsilon_{r2} = 1$).

3.4.2 Dielectric-loaded Waveguide Slots

In this subsection, both transverse and longitudinal slots in a dielectric-loaded waveguide are analyzed. Two different slots, including one transverse slot and one longitudinal slot, are fabricated and measured in order to verify our hybrid technique. The test waveguide section consists of two parts: one is a metallic cover and the other is a rectangular waveguide partially filled with a dielectric slab. The transverse and longitudinal slots are separately cut on the metallic covers; then the covers are mounted on the dielectric-loaded waveguide. The thickness of the metallic covers is $1mm$. The inner dimensions of the rectangular waveguide are

3.4 Numerical Results

Figure 3.16: Normalized series resistance and reactance for a centered transverse slot in a rectangular waveguide partially filled with a dielectric slab (Parameters as in Fig. 3.15).

$15\text{mm} \times 6\text{mm}$. The slab has a thickness of 3mm and a relative permittivity of 6. It is clear that, by inserting a dielectric slab, the dimensions of the waveguide are greatly reduced in the same frequency range. The measurement is then conducted by fixing different covers on the waveguide. Computed results for the scattering parameters of the dielectric-filled waveguide with a transverse slot of $15\text{mm} \times 6\text{mm}$ are presented in Fig. 3.15 and compared to experimental data. Good agreement between them is observed.

The circuit parameters of this transverse slot are also extracted from the scat-

3.4 Numerical Results

Figure 3.17: Scattering parameters of a longitudinal slot in a rectangular waveguide partially filled with a dielectric slab ($w = 3mm$, $l = 14mm$, $d = 3.75mm$; Other parameters the same as in Fig. 3.15).

tering parameters and compared to those obtained by HFSS, as shown in Fig. 3.16. It is seen that our results agree very well with those by HFSS even at the resonant point of the slot. The mode numbers used in the analysis of the dielectric-loaded T-junction shown in Fig. 3.6 are 50, 30 and 200 for Regions 1, 3 and 4, respectively. Nine modes out of the 30 modes of the slot are used in the calculation of the open-ended waveguide. The computation time of our hybrid method for the dielectric-loaded structure is 11 seconds for one frequency point, while the CPU time for HFSS is nearly 5 minutes on the same computer (10 iterations, 5273 tetrahedra). It is clear that our hybrid method is very efficient compared to the

3.4 Numerical Results

Figure 3.18: Normalized shunt admittance for a longitudinal slot in a rectangular waveguide partially filled with a dielectric slab ($w = 3mm$, $l = 14mm$, $d = 3.75mm$; Other parameters the same as in Fig. 3.15).

finite-element method used in HFSS.

The longitudinal slot cut on the broad wall of a rectangular waveguide partially filled with a dielectric slab has dimensions of $3mm \times 14mm$. The offset of the slot from the center of the waveguide is $3.75mm$. Calculated results of the scattering parameters are compared to the experimental data, as illustrated in Fig. 3.17. The agreement between them is good. The normalized shunt conductance G and susceptance B are extracted from our calculated S-parameters and plotted in Fig. 3.18. Simulated data by HFSS are also presented for comparison. Again, a good

3.4 Numerical Results

agreement between them is observed. A slight difference in susceptance B is noticed when the frequency is above 11 GHz. This is due to fact that, at high frequencies, the imaginary part of the transmission coefficient is very close to 0 while its real part is very close to 1. Since the susceptance is sensitive to the imaginary part of the transmission coefficient, it will take much more time for HFSS to obtain convergent results for susceptance B .

3.4.3 Parametrical Study

In order to design a waveguide slot, the waveguide dimension should be determined first to ensure that only the dominant mode can propagate in the frequency range considered. Next, the slot dimension is then decided so that the slot is designed at its resonant point. Generally, the slot length determines the resonant frequency, while the slot width and the offset from the waveguide center have an effect on its normalized resonant impedance/admittance. In addition, the thickness of the waveguide wall is also an important parameter, which has a great effect on the resonant impedance/admittance.

Compared to the conventional air-filled waveguide slot, a partially dielectric-loaded waveguide slot can offer two additional structural parameters, the thickness t of the dielectric slab and its relative permittivity ε_{r1} . The effect of these two parameters on the impedance/admittance of a transverse slot and a longitudinal slot is examined here.

A transverse slot cut on the broad-wall of a rectangular waveguide partially filled with a dielectric slab has the same dimension as the waveguide of $15mm \times$

3.4 Numerical Results

Figure 3.19: Normalized series resistance and reactance for a partially dielectric-loaded waveguide transverse slot with different permittivities of the dielectric slab ($a = 15mm$, $b = 6mm$, $t = 3mm$, $w = 15mm$, $l = 6mm$, $s = 1mm$, $\epsilon_{r2} = 1$).

$6mm$ and is located at the center of the waveguide wall. The normalized resistance and reactance are illustrated for different values of ϵ_{r1} in Fig. 3.19, when the dielectric slab's thickness is fixed to be $3mm$. It is observed that the normalized resonant series resistance R can be adjusted from 1.5 to 0.5 when the dielectric constant changes from 3 to 9. The resonant frequency also decreases when ϵ_{r1} increases. Next, the effect of the dielectric slab's thickness on the same transverse slot is studied while the relative permittivity ϵ_{r1} is fixed at 6, as shown in Fig. 3.20. It can be seen that the normalized resonant series resistance R changes from 1.3 to 0.9 when the slab's thickness increases from $2.2mm$ to $3mm$.

3.4 Numerical Results

Figure 3.20: Normalized series resistance and reactance for a partially dielectric-loaded waveguide transverse slot with different thicknesses of the dielectric slab ($a = 15\text{mm}$, $b = 6\text{mm}$, $w = 15\text{mm}$, $l = 6\text{mm}$, $s = 1\text{mm}$, $\varepsilon_{r1} = 6$, $\varepsilon_{r2} = 1$).

A longitudinal slot cut on the broad-wall of a rectangular waveguide partially filled with a dielectric slab has a dimension of $3\text{mm} \times 14\text{mm}$ and an off-centered displacement of 3.75mm . The thickness of the dielectric slab is initially fixed to 3mm to study the effect of the slab's relative permittivity on the normalized shunt conductance and susceptance, as shown in Fig. 3.21. It is observed that the normalized resonant shunt conductance G can be adjusted from 0.65 to 0.1 when the dielectric constant changes from 3 to 9. The resonant frequency also decreases when ε_{r1} increases, as expected. Comparing Fig. 3.21 with Fig. 3.19, it is seen that the slab's dielectric constant has similar effect on both transverse

3.4 Numerical Results

and longitudinal slots. It is also observed that the shift in the resonant frequency of the longitudinal slot is larger than that of the transverse slot. The effect of the dielectric slab's thickness on the same longitudinal slot is examined in Fig. 3.22 when the relative permittivity ϵ_{r1} is fixed at 6. It is seen that the normalized resonant shunt conductance G can be adjusted from 0.45 to 0.25 when the slab's thickness increases from 2.2mm to 3mm.

From Figs. 3.19-3.22, it can be concluded that both the slab's thickness and dielectric constant have significant effect on the normalized impedance/admittance

Figure 3.21: Normalized shunt conductance and susceptance for a partially dielectric-loaded waveguide longitudinal slot with different permittivities of the dielectric slab ($a = 15\text{mm}$, $b = 6\text{mm}$, $t = 3\text{mm}$, $w = 3\text{mm}$, $l = 14\text{mm}$, $d = 3.75\text{mm}$, $s = 1\text{mm}$, $\epsilon_{r2} = 1$).

3.4 Numerical Results

of a partially dielectric-loaded waveguide slot. Desired resistance/conductance and reactance/susceptance values may be achieved by adjusting these two parameters along with slot length and slot width.

In addition to the slab's thickness and dielectric constant, the location of the dielectric layer can be adjusted. The dielectric slab can be inserted either at the bottom or at the top in the rectangular waveguide. A transverse slot is considered here. The slot has the same dimensions as the waveguide of $15\text{mm} \times 6\text{mm}$. The slab's thickness and dielectric constant are 3mm and 6mm , respectively. The

Figure 3.22: Normalized shunt conductance and susceptance for a partially dielectric-loaded waveguide longitudinal slot with different thicknesses of the dielectric slab ($a = 15\text{mm}$, $b = 6\text{mm}$, $w = 3\text{mm}$, $l = 14\text{mm}$, $d = 3.75\text{mm}$, $s = 1\text{mm}$, $\epsilon_{r1} = 6$, $\epsilon_{r2} = 1$).

3.4 Numerical Results

(a)

(b)

Figure 3.23: Normalized series (a) resistance and (b) reactance for a partially dielectric-loaded waveguide transverse slot when the dielectric slab is at different layers of the waveguide ($a = 15\text{mm}$, $b = 6\text{mm}$, $w = 15\text{mm}$, $l = 6\text{mm}$, $s = 1\text{mm}$).

3.4 Numerical Results

Figure 3.24: Normalized series resistance and reactance for a transverse slot cut on an air-filled rectangular waveguide ($a = 22.86\text{mm}$, $b = 10.16\text{mm}$, $w = 14.1\text{mm}$, $l = 5\text{mm}$, $s = 1\text{mm}$).

normalized resistance and reactance are shown in Fig. 3.23(a) and Fig. 3.23(b), respectively. It is seen that the difference for their resistances is significant. It is also observed that the magnitude of the reactance is very big when the slab is at the top of the waveguide. This is due to the fact that the electromagnetic fields near the slot are different in the two configurations and this in turn results in different coupling coefficients from the waveguide to the slot.

In order to appreciate the advantage using a partially dielectric-loaded waveguide slot antenna, two transverse slots are designed at X -band with $R = 1$ and

3.4 Numerical Results

Figure 3.25: Normalized series resistance and reactance for a transverse slot cut on a partially dielectric-loaded rectangular waveguide ($a = 15mm$, $b = 6mm$, $t = 3mm$, $w = 14.5mm$, $l = 5mm$, $s = 1mm$, $\epsilon_{r1} = 6$, $\epsilon_{r2} = 1$).

$X = 0$. The air-filled waveguide slot is of dimension $14.1mm \times 5mm$ and the off-centered displacement $d = 3.15mm$. The normalized resistance and reactance are plotted in Fig. 3.24. In the case of a partially dielectric-loaded waveguide slot, the waveguide dimensions are $15mm \times 6mm$ and a $3mm$ thick slab with $\epsilon_{r1} = 6$ is placed at the bottom. The slot has dimensions of $14.5mm \times 5mm$ and is located at the center of the waveguide. R and X of the partially dielectric-loaded waveguide slot are illustrated in Fig. 3.25. The guide wavelength is changed from $1.3\lambda_0$ to $0.8\lambda_0$ after inserting the dielectric slab into the waveguide, where λ_0 is the wavelength in free space. It is seen that the partially dielectric-loaded structure

3.4 Numerical Results

greatly reduces the waveguide size, while the behaviors of R and X near the resonant frequency are similar. It is known that the normalized resonant resistance of a transverse slot cut in a waveguide fully filled with a dielectric material is too high and its bandwidth is very narrow. Similar behavior is also observed for longitudinal slots in a rectangular waveguide partially filled with a dielectric slab.

Many researchers had pointed out that the thickness of the waveguide wall has a great effect on the characteristics of the slot cut on the waveguide. We also present an example to show such effect. The slot element designed in the previous paragraph is considered here. The normalized resistance and reactance of the slot cut on the rectangular waveguide with different wall thicknesses are illustrated in Fig. 3.26. It is observed that changing the thickness of the waveguide wall can notably alter the radiation performance of the slot. When the wall thickness is increased from $0mm$ (which is a general assumption for earlier researchers) to $4mm$, the normalized resonating resistance is decreased from 1.1 to 0.85. The resonating frequency is also shifted from 9.8 GHz to 10.2 GHz, which is about 4 percent of the center frequency. From Fig. 3.26 it is demonstrated that the thickness of the waveguide wall should be taken into account with care when a high accuracy is required in waveguide slot design. However, zero-thickness can still be a good assumption for the waveguide wall of a thickness less than $1mm$ when the working frequency is around 10 GHz.

It should be mentioned that, the mode numbers should change to keep a constant accuracy when the permittivity and thickness of the two layers in the waveguide vary. However, we fix the mode numbers to 50, 30 and 200 for Regions 1, 3

3.4 Numerical Results

(a)

(b)

Figure 3.26: Normalized series resistance and reactance of a dielectric-loaded waveguide slot with different thicknesses of the waveguide wall (Parameters are the same as in 3.25 except the thickness s of the waveguide wall).

3.4 Numerical Results

Figure 3.27: Relative error for $|S_{11}|$ of a dielectric-loaded waveguide slot (Parameters are the same as in 3.25 except the thickness s of the waveguide wall; Mode numbers for Regions 1 and 4 in Fig. 3.6 are 50 and 200, respectively).

and 4, respectively. After a careful convergence investigation, it is found that the relative error of the scattering parameters is below 0.01 for all the numerical examples considered in this section. In fact, the required number of modes in Region 3 in Fig. 3.6 is also dependent on the thickness s of the waveguide wall. Figure 3.27 illustrates the relative error for $|S_{11}|$ of a transverse slot cut on a partially dielectric loaded rectangular waveguide. The mode numbers in Regions 1 and 4 are fixed to 50 and 200, respectively. It is seen that the relative error decreases with an increasing wall thickness s . This is due to the fact that higher-order modes in Region 3 are attenuated very rapidly. Therefore, a better accuracy can be ob-

3.5 Conclusion

tained with a larger s when the mode number in Region 3 is a constant. It can also be observed in Fig. 3.27 that the relative error is lower than 0.005 with 30 modes in Region 3 even when the waveguide has a zero wall thickness.

3.5 Conclusion

In this chapter we presented a rigorous full-wave analysis of longitudinal and transverse slots cut in the wall of a rectangular waveguide partially filled with a dielectric slab. The slot structure is firstly decomposed into two parts: a dielectric-loaded T-junction and an open-ended waveguide, which are formulated by the mode-matching method and moment method, respectively. Cascading of the two blocks yields the scattering parameters of the waveguide slot structure. This new method has a great advantage over the conventional moment method since the complexity of deriving the Green's function of a partially dielectric-loaded waveguide is avoided. A new approach has been proposed for computing the integrals in the analysis of an open-ended waveguide, which can greatly reduce the computation time in the moment method. This hybrid mode-matching/moment method can be easily extended to analyze and design more complicated slot antennas and arrays.

Numerical results for the scattering parameters and circuit parameters have been presented and verified by comparing with available measured data or those obtained by Ansoft's HFSS. Excellent agreement has been observed for all the cases considered. The accuracy and efficiency of the proposed hybrid method for modeling dielectric-loaded waveguide slot structures have been demonstrated. The effect of the dielectric slab's thickness and relative permittivity on the slot antenna's im-

3.5 Conclusion

pedance/admittance has been examined. Calculated results have shown that the normalized impedance/admittance can be effectively adjusted by changing the loaded slab's thickness and dielectric constant, in addition to the slot dimensions.

Chapter 4

Radiation Patterns of Dielectric-loaded Waveguide Slots

4.1 Introduction

Rectangular waveguide slot antenna has found wide applications in modern radar and communication systems due to their attractive features such as low profile, high power, high efficiency and ease of being flush-mounted. Since World War II, waveguide slots have been extensively investigated. The internal Green's function for a rectangular waveguide was established first and the resistance/conductance of longitudinal slot, inclined slot and transverse slot were determined [15]. After that, a variational formulation was developed [26] to calculate the slot impedance/admittance taking the effect of the finite wall thickness into account. The more complicated compound slot problem was solved later [28]. Recently, a transverse slot array on the broad wall of a rectangular waveguide partially filled with a dielectric slab was presented [42]. Compared to conventional air-filled

4.1 Introduction

waveguide slot antennas, dielectric-loaded waveguide slot antennas can realize a compact antenna array and provide extra parameters to optimize the antenna performance. A hybrid mode-matching method was also proposed [76] to analyze partially dielectric-loaded waveguide slot antennas. In almost all the previous studies, the basic idea of analyzing the waveguide slot structure is to separate the problem into two regions: outside and inside of the slot, and then expand the field in the slot in terms of the summation of basis functions. The H-field can then be found for both regions by some full wave methods. The expansion coefficients are therefore determined by matching the two H-field expressions along the slot. Once the field in the slot is obtained, the radiation pattern can be calculated as well as the slot impedance/admittance.

Although much work has been carried out for waveguide slot antennas over the years, little was focused on the finite ground plane effect on their radiation performance. When the radiation problem is considered, it is normally assumed that the outer region of the waveguide slot is a half free space bounded by an infinite ground plane [53, 77]. However, ground planes of infinite extent are not realizable in practice, though they can be approximated closely by a very large size. The radiation patterns of waveguide slot antennas on a finite ground plane can be modified considerably by the effects of the edges, especially in regions of very low field intensity such as the backward radiation. Presently, there are two kinds of methods that can be used conveniently to account for the edge effects. One is the full wave methods, such as the method of moment (MOM) [78] and the cylindrical-wave-spectrum method [79]. The other one is the geometrical theory of diffraction (GTD) [80]. However, the full wave methods generally need a lot of numerical

4.1 Introduction

computations in modeling different ground planes and waveguide structures and will become inefficient for objects that are electrically large. This limitation can be overcome by using the GTD. The GTD is an extension of classical geometrical optics (GO) by introducing a diffraction mechanism. It is simple for numerical computation and can be used to solve complicated structures that do not have exact solutions [38, 81]. GTD has been successfully applied to model various antennas such as microstrip antennas [82] and microstrip-fed cavity-backed slot antennas [83]. It has been shown that the GTD can yield accurate results, which compare well with other methods.

In this chapter, a hybrid GTD method is presented to calculate the radiated fields and to examine the finite ground plane effect for both air-filled and partially dielectric-loaded rectangular waveguide slot antennas. A recently presented mode-matching/moment method [76] is used to calculate the electric field distribution along the slot. This method is computationally efficient by taking advantage of the mode-matching method's high efficiency in analyzing the waveguide T-junction and the moment method's simplicity for the open-ended waveguide. After the field distribution in the waveguide slot is obtained, the GO field can be easily calculated and the GTD is employed to account for the diffracted fields from all the waveguide wall edges. Numerical results for various slots are validated by comparing with measured and simulated data by Ansoft's HFSS. Good agreement is observed for all the cases considered. It is shown that a finite ground plane has a small effect on both E-plane and H-plane patterns of rectangular waveguide slot antennas in the forward direction, while its effect on the backward radiation is very significant and must be taken into consideration in the design of slot antenna arrays of low

4.2 Field Distribution along the Waveguide Slot

side-lobe level.

4.2 Field Distribution along the Waveguide Slot

Let us consider a slot cut in the wall of a dielectric-loaded rectangular waveguide, as shown in Fig. 3.1. The rectangular waveguide is of width a and height b . There are two layers of dielectric materials in the waveguide with the relative permittivities being ε_{r1} and ε_{r2} , respectively. The thickness of the lower layer is t . The slot cut in the wall of the waveguide is of width w and length l . The thickness of the waveguide wall is s . The displacement of the slot from the center of the waveguide is d . The side view of the slot structure is given in Fig. 3.2(a). Following the procedure in [76], the waveguide slot antenna is separated into two parts: a three-port waveguide T-junction and an open-ended waveguide, as illustrated in Fig. 3.2(b). The length of the rectangular side-arm of the T-junction is set to s , which is the thickness of the waveguide wall. The cross-sectional dimensions of the side-arm of the T-junction and the open-ended waveguide are equal to those of the original slot.

The electric field distribution along the waveguide slot can then be expressed as:

$$\mathbf{E}_t = \sum_{j=1}^{M_3} [A_{3j}^- e^{-\gamma_{3j}s} + A_{3j}^+ e^{+\gamma_{3j}s}] \mathbf{e}_j \quad (4.1)$$

where M_3 is the mode number used in the slot; A_{3j}^+ and A_{3j}^- are the modal amplitude coefficients for incident and reflected waves, respectively. $\mathbf{e}_j$ is the normalized

4.2 Field Distribution along the Waveguide Slot

transverse field component given in the coordinates of Fig. 4.1 by:

$$\mathbf{e}_j = \begin{cases} \hat{y} \times \nabla_t N_j \cos \frac{m\pi x}{w} \cos \frac{n\pi z}{l} & TE \text{ modes} \\ \nabla_t N_j \sin \frac{m\pi x}{w} \sin \frac{n\pi z}{l} & TM \text{ modes} \end{cases} \quad (4.2)$$

where ∇_t is the 2-D differential operator having the form $\nabla_t = \hat{x} \frac{\partial}{\partial x} + \hat{z} \frac{\partial}{\partial z}$; N_j is the normalization factor and j is the mode index used in the field expansion and stands for the combination of sub indices m and n .

The dielectric-loaded three-port waveguide T-junction has been formulated by the mode-matching method [61] and the generalized scattering matrix $[\mathbf{S}_{T-junction}]$ is obtained.

$$\begin{bmatrix} \mathbf{A}_1^- \\ \mathbf{A}_2^- \\ \mathbf{A}_3^- \end{bmatrix} = [\mathbf{S}_{T-junction}] [\mathbf{A}^+] = \begin{bmatrix} \mathbf{S}_{11} & \mathbf{S}_{12} & \mathbf{S}_{13} \\ \mathbf{S}_{21} & \mathbf{S}_{22} & \mathbf{S}_{23} \\ \mathbf{S}_{31} & \mathbf{S}_{32} & \mathbf{S}_{33} \end{bmatrix} \begin{bmatrix} \mathbf{A}_1^+ \\ \mathbf{A}_2^+ \\ \mathbf{A}_3^+ \end{bmatrix} \quad (4.3)$$

where $[\mathbf{A}_i^+]$ and $[\mathbf{A}_i^-]$ are the modal amplitude coefficients of incident and reflected waves at port i ($i = 1, 2, 3$), respectively. The open-ended waveguide has been characterized [76] and the reflection coefficient matrix $[\mathbf{R}]$ is derived as $[\mathbf{A}_3^+] = [\mathbf{R}] [\mathbf{A}_3^-]$. The two parts are then cascaded together to give the scattering parameters of the waveguide slot $[\mathbf{S}_{slot}]$.

$$[\mathbf{S}_{slot}] = \begin{bmatrix} \mathbf{S}_{11}^S & \mathbf{S}_{12}^S \\ \mathbf{S}_{21}^S & \mathbf{S}_{22}^S \end{bmatrix} \quad (4.4)$$

4.3 Calculation of Radiation Patterns

If the excitation at the waveguide ports 1 and 2 are given, $[\mathbf{A}_3^+]$ and $[\mathbf{A}_3^-]$ can be calculated from:

$$\begin{cases} \mathbf{A}_3^- = (\mathbf{I} - \mathbf{S}_{33}\mathbf{R})^{-1} (\mathbf{S}_{31}\mathbf{A}_1^+ + \mathbf{S}_{32}\mathbf{A}_2^+) \\ \mathbf{A}_3^+ = \mathbf{R}\mathbf{A}_3^- \end{cases} \quad (4.5)$$

and the field distribution along the slot can be readily obtained from Eq.(4.1).

4.3 Calculation of Radiation Patterns

In this section, we first derive the magnetic field in the half free space outside the waveguide slot when the E-field distribution along the slot is known. This calculated field is referred to as the geometrical optics (GO) field. Then we show how to include the effect of a finite ground plane in the calculation of radiation patterns using GTD.

4.3.1 GO Fields

When a waveguide slot with a field distribution $\mathbf{E}_{slot}$ is mounted on an infinite ground plane, an equivalent magnetic current $\mathbf{M}_s$ can be used to replace the E-field with the slot short-circuited. The equivalent magnetic current is $\mathbf{M}_s = -\mathbf{n} \times \mathbf{E}_{slot}$ with $\mathbf{n}$ being the normal unit vector of the ground plane. In the half free space, the magnetic field due to the equivalent magnetic current $\mathbf{M}_s$ is given by [2]:

$$\mathbf{H}_{GO} = -j\omega\mathbf{F} - \nabla\phi_m \quad (4.6)$$

where $\mathbf{F}$ and ϕ_m are the electric vector potential and the magnetic scalar potential,

4.3 Calculation of Radiation Patterns

respectively.

$$\mathbf{F} = \varepsilon_0 \iint_{slot} \mathbf{M}_s(\mathbf{r}') G(\mathbf{r}|\mathbf{r}') dS' \quad (4.7)$$

$$\phi_m = \frac{j}{\omega\mu_0\varepsilon_0} \nabla \cdot \mathbf{F} \quad (4.8)$$

with $G(\mathbf{r}|\mathbf{r}')$ being the scalar Green's function in the half-space.

$$G(\mathbf{r}|\mathbf{r}') = \frac{e^{-jk_0 R}}{2\pi R} \quad (4.9)$$

where $R = |\mathbf{r} - \mathbf{r}'|$ is the distance between the field point $\mathbf{r}$ and the source point $\mathbf{r}'$. In this way, the GO field can be calculated for any observation point in the half space.

4.3.2 GTD Formulation

For a waveguide slot antenna mounted on a finite ground plane, as illustrated in Fig. 4.1, the actual field in the outer space will be different compared with that for an infinite ground plane. Following the procedure in [38], we utilize GTD to calculate the diffracted fields from the ground plane edges. The E-plane and H-plane radiation patterns are calculated separately.

Figure 4.2 depicts the radiation and diffraction mechanisms of the E-plane pattern for a waveguide slot antenna. The total field can be calculated by summing the GO field radiated from the slot and the diffracted fields from the edges. The GO electric field $\mathbf{E}_{GO}$ can be easily obtained from $\mathbf{H}_{GO}$, which is calculated in

4.3 Calculation of Radiation Patterns

Figure 4.1: Waveguide slot antenna on a finite ground plane.

Eq.(4.6). The singly diffracted fields from the edges 1 and 2 can be expressed as:

$$\mathbf{E}_d(i) = \hat{\mu}_i E_{inc} D_h A_d(S_i) e^{-jk_0 S_i} \quad (4.10)$$

where E_{inc} is the incident electric field at the edge, and $A_d(S_i)$ is the amplitude attenuation coefficient

$$A_d(S) = \begin{cases} \frac{1}{\sqrt{S}} & \text{for plane, cylindrical, and} \\ & \text{conical wave incidence} \\ \sqrt{\frac{S'}{S(S' + S)}} & \text{for spherical wave incidence} \end{cases} \quad (4.11)$$

4.3 Calculation of Radiation Patterns

(a) Radiation and singly diffraction

(b) Multiple diffraction

Figure 4.2: E-plane radiation and diffraction mechanisms.

4.3 Calculation of Radiation Patterns

and D_h is the hard-boundary diffraction coefficient given by [38]:

$$\begin{aligned}
 D_h(\phi, \phi'; \beta_0) = & \frac{-e^{-j\pi/4}}{2n\sqrt{2\pi k_0} \sin \beta_0} \\
 & \times \left\{ \cot \left(\frac{\pi + (\phi - \phi')}{2n} \right) F[k_0 L a^+(\phi - \phi')] \right. \\
 & + \cot \left(\frac{\pi - (\phi - \phi')}{2n} \right) F[k_0 L a^-(\phi - \phi')] \\
 & + \cot \left(\frac{\pi + (\phi + \phi')}{2n} \right) F[k_0 L a^+(\phi + \phi')] \\
 & \left. + \cot \left(\frac{\pi - (\phi + \phi')}{2n} \right) F[k_0 L a^-(\phi + \phi')] \right\} \quad (4.12)
 \end{aligned}$$

where $F(\phi, \phi')$ is the transition function, β_0 is the angle between the incident wave and the edge. n represents the conducting wedge with an angle of $(2 - n)\pi$. If the diffraction is from the edge of a ground plane with a zero thickness, the angle of the wedge will be zero and $n = 2$. However, in practice, the size of the ground plane on which the waveguide slot antenna is mounted may be just the same as that of the rectangular waveguide. The angle of the diffraction wedge is $\pi/2$ and $n = 1.5$. In this case, the diffractions from the other two edges must be included in the E-plane pattern calculation, as illustrated in Fig. 4.2(b). The diffracted fields from the edges 3 and 4 can be calculated in the similar way as Eq.(4.10), while the incident waves are due to the diffraction from the edges 1 and 2, respectively.

The H-plane pattern in the forward region can also be calculated by summing the GO field from the slot and the diffracted fields from the edges, as shown in Fig. 4.3. The first-order diffracted field from edges 1' and 2' is zero because the electric

4.3 Calculation of Radiation Patterns

Figure 4.3: H-plane radiation and diffraction mechanisms.

field on the surface of a conductor wedge vanishes for the parallel polarized grazing incident wave (soft boundary condition). However, a second-order diffracted field derived from Maxwell's equations is nonzero and known as slope diffraction [84], which is due to the rapid change of GO field near the edge. Therefore, the electric field is given by [81]:

$$\mathbf{E}_{slope}(i) = \hat{z} \frac{1}{jk_0} \frac{1}{HC_i} \left. \frac{\partial E_{GO}}{\partial \mu} \right|_{\mu=0 \text{ or } \pi} D_{slope} A_d(S_i) e^{-jk_0 S_i} \quad (4.13)$$

4.3 Calculation of Radiation Patterns

where D_{slope} is the slope diffraction coefficient given by [81]

$$\begin{aligned}
 D_{slope}(\phi, \phi'; \beta_0) = & \frac{-e^{-j\pi/4}}{4n^2\sqrt{2\pi k_0} \sin \beta_0} \\
 & \times \left\{ \csc^2 \left(\frac{\pi + (\phi - \phi')}{2n} \right) F_s[k_0 L a^+(\phi - \phi')] \right. \\
 & + \csc^2 \left(\frac{\pi - (\phi - \phi')}{2n} \right) F_s[k_0 L a^-(\phi - \phi')] \\
 & + \csc^2 \left(\frac{\pi + (\phi + \phi')}{2n} \right) F_s[k_0 L a^+(\phi + \phi')] \\
 & \left. + \csc^2 \left(\frac{\pi - (\phi + \phi')}{2n} \right) F_s[k_0 L a^-(\phi + \phi')] \right\} \quad (4.14)
 \end{aligned}$$

Unlike the E-plane diffraction, there is no diffraction from the edges 3' and 4' because the incident waves on edges 3' and 4' are zero.

When we are calculating the back lobe region of the H-plane pattern, the contributions from the E-plane edge diffractions need to be included because the E-plane edge diffraction has a much larger magnitude than that of the H-plane edge slope diffraction. This E-plane contribution has been characterized by an equivalent current method [85]. The equivalent current model is illustrated in Fig. 4.4 and can be expressed as [82]

$$\mathbf{E}_{eq} = \frac{jk_0}{4\pi} \int_{-HC2}^{HC1} \hat{S} \times \mathbf{I}_m(x') \frac{e^{-jk_0 S}}{S} dx' \quad (4.15)$$

with S being the distance between the diffraction point x' and the observation

4.3 Calculation of Radiation Patterns

(a)

(b)

Figure 4.4: Equivalent magnetic current for the H-plane radiation pattern.

4.4 Numerical Results

point. $I_m(x')$ is the equivalent magnetic line current given in [81] as:

$$I_m(x') = -\hat{x} \cdot \frac{\mathbf{H}_{inc}(x')}{Y_0} D_h \sqrt{\frac{8\pi}{k_0}} e^{-j\pi/4} \quad (4.16)$$

where $\mathbf{H}_{inc}(x')$ is the incident magnetic field at x' , Y_0 is the free space admittance, and D_h is the hard boundary diffraction coefficient. So it can be concluded that the total H-plane field is the vector superposition of the GO field ($\mathbf{E}_{GO}$), the slope diffracted field ($\mathbf{E}_{slope}$) and the equivalent current field ($\mathbf{E}_{eq}$).

4.4 Numerical Results

In this section, we present E-plane and H-plane radiation patterns obtained by our hybrid GTD method for both air-filled and dielectric-loaded rectangular waveguide slot antennas. Longitudinal and transverse slot antennas on an air-filled waveguide are examined first with both finite and infinite ground planes. The effect of a finite ground plane on the radiation patterns is then studied. The results of the slot antenna with a finite ground plane are compared to those obtained by Ansoft's HFSS for the purpose of validation. After that, longitudinal and transverse slot antennas on a partially dielectric-loaded waveguide are also studied and the results obtained by our method are then compared with measured data. The effect of a finite ground plane on the dielectric-loaded waveguide slots is also shown.

4.4.1 Air-filled Rectangular Waveguide Slot Antennas

First, we consider a longitudinal slot antenna on the X-band waveguide (WR-90: $22.86mm \times 10.16mm$) operating at 10 GHz. The center offset of the slot is 4.6mm.

4.4 Numerical Results

E-plane and H-plane radiation patterns for the slot antenna with both an infinite ground plane and a finite ground plane are illustrated in Fig. 4.5.

The finite ground plane has the same size as the rectangular waveguide which has a length of four wavelengths ($120mm$), and a wall thickness of $1mm$. The results obtained by HFSS for the slot antenna with the same finite ground plane are also given. Good agreement between our method and HFSS is observed. The computation time of our method is 20 seconds for calculating the scattering parameters and generating the radiation patterns, while the CPU time for HFSS is 30 minutes on the same computer. It is clear that our method is very efficient compared to the finite-element method used in HFSS. From Fig. 4.5, it can be seen that both E-plane and H-plane patterns of the slot antenna with an infinite ground plane are symmetric, as we expected. This is because the outer half space is symmetric and the E-field in the resonating slot follows a TE_{10} distribution. Generally speaking, both E-plane and H-plane patterns of the slot antenna with the finite ground plane do not change much compared to that with an infinite ground plane. The E-plane pattern for the slot antenna in a finite ground plane exhibits a squint of 20 degrees. This is due to the fact that the two E-plane edges are asymmetric and very close to the radiating slot. Therefore, the diffraction from one edge has a much larger magnitude than that of the other, which noticeably changes the radiating direction. The H-plane pattern of the slot antenna in a finite ground plane remains symmetric. This is obvious because the two H-plane edges are symmetric with respect to the slot.

Next, a transverse slot in the same X-band waveguide is considered. The center

4.4 Numerical Results

(a) E-plane

(b) H-plane

Figure 4.5: Radiation patterns of a longitudinal slot antenna in an air-filled rectangular waveguide ($a = 22.86mm$, $b = 10.16mm$, $w = 3mm$, $l = 14.3mm$, $d = 4.6mm$, $s = 1mm$, $HD1 = HD2 = 60mm$, $HC1 = 17.03mm$, $HC2 = 7.83mm$, $f = 10$ GHz).

4.4 Numerical Results

(a) E-plane

(b) H-plane

Figure 4.6: Radiation patterns of a transverse slot antenna in an air-filled rectangular waveguide ($a = 22.86\text{mm}$, $b = 10.16\text{mm}$, $w = 14.1\text{mm}$, $l = 5\text{mm}$, $d = 3.15\text{mm}$, $s = 1\text{mm}$, $\text{HD1} = \text{HD2} = 60\text{mm}$, $\text{HC1} = 9.28\text{mm}$, $\text{HC2} = 15.58\text{mm}$, $f = 10\text{ GHz}$).

4.4 Numerical Results

offset is 3.15mm . The radiation patterns calculated by our method for the slot antenna with both infinite and finite ground planes are plotted in Fig. 4.6. It is obvious that the E-plane pattern is symmetric, since the E-plane edges are symmetric ($\text{HD1} = \text{HD2} = 60\text{mm}$). However, the vectorial combination of the GO field and the diffracted fields from the edges results in the small ripples in the forward region. The H-plane pattern in the forward region is almost symmetric but this is not the case for the backward region. This is due to the fact that the magnitude of the H-plane slope diffraction is very small compared to the GO field. Therefore, the edge diffracted field has only slight effect on the H-plane in the forward region, though the slope diffracted fields from the two H-plane edges are different for this off-centered transverse slot. When $|\theta|$ is greater than $\frac{2\pi}{3}$, the magnitude of GO field decreases rapidly and the effect of slope diffraction becomes apparent. In the back-side lobe of H-plane pattern, the superposition of the slope diffraction from H-plane edges and the equivalent current diffraction from E-plane edges creates an asymmetric pattern.

4.4.2 Partially Dielectric-loaded Rectangular Waveguide Slot Antennas

Longitudinal and transverse slot antennas in a partially dielectric-loaded rectangular waveguide are also examined. A longitudinal slot antenna is fabricated and measured in order to verify our hybrid GTD method. The slot antenna consists of two parts, a base as the waveguide and a metal cover on which the slot is opened. The waveguide has a inner dimension of $15\text{mm} \times 6\text{mm}$. A dielectric slab with a relative permittivity of 6 and a thickness of 3mm is inserted in the bottom layer

4.4 Numerical Results

of the waveguide. The metal cover is 1mm thick and the slot is $3\text{mm} \times 14\text{mm}$. After the assembling of the two parts, the outer dimension of the cross section of the whole structure is $53\text{mm} \times 13\text{mm}$. Computed results for the radiation patterns of the partially dielectric-loaded waveguide slot antenna with both finite and infinite ground planes are presented in Fig. 4.7. Measured data for E-plane pattern is also given. Good agreement is observed between calculated data and experimental ones. It is seen that the E-plane pattern of the slot antenna with a finite ground plane has a squint of 15 degrees compared to that with an infinite ground plane. Similar to the results of the air-filled waveguide slot antenna, which is illustrated in Fig. 4.5(a), the squint is due to the center offset and the different magnitudes of the diffraction from the two E-plane edges. The measured H-plane pattern is not given because it is dependent on the complicated feeding structures at the two ends. As the two H-plane edges are symmetric with respect to the slot, the H-plane pattern for the slot with a finite ground plane is symmetric, as we expected. However, a finite ground plane introduces backward radiation as well as the small ripples in the forward region.

Next, a transverse slot antenna in the same rectangular waveguide partially loaded with dielectric is studied. The slot is located at the center of the waveguide. Both E-plane and H-plane edges are symmetric with respect to the slot, as illustrated in Fig. 4.8. It is observed for the E-plane pattern that a finite ground plane generates backward radiation and small ripples in the forward region, compared to the data with an infinite ground plane. As for the H-plane pattern, the contribution of the slope diffraction from the H-plane edges and the equivalent current diffraction from the E-plane edges eliminates the nulls of GO field at $\theta = \pm \frac{\pi}{2}$

4.4 Numerical Results

(a) E-plane

(b) H-plane

Figure 4.7: Radiation patterns of a longitudinal slot antenna in a partially dielectric-loaded rectangular waveguide ($a = 15mm$, $b = 6mm$, $t = 3mm$, $w = 3mm$, $l = 14mm$, $d = 3.75mm$, $s = 1mm$, $\epsilon_{r1} = 6$, $\epsilon_{r2} = 1$, $HD1 = HD2 = 60mm$, $HC1 = 22.75mm$, $HC2 = 30.25mm$, $f = 10$ GHz).

4.4 Numerical Results

(a) E-plane

(b) H-plane

Figure 4.8: Radiation patterns of a centered transverse slot antenna in a partially dielectric-loaded rectangular waveguide ($a = 15\text{mm}$, $b = 6\text{mm}$, $t = 3\text{mm}$, $w = 15\text{mm}$, $l = 6\text{mm}$, $s = 1\text{mm}$, $\epsilon_{r1} = 6$, $\epsilon_{r2} = 1$, $\text{HD1} = \text{HD2} = 60\text{mm}$, $\text{HC1} = \text{HC2} = 26.5\text{mm}$, $f = 10\text{ GHz}$).

4.4 Numerical Results

Figure 4.9: E-plane radiation pattern of a centered transverse slot in a partially dielectric-loaded rectangular waveguide (Parameters are the same as in Fig. 4.8 except the frequency f).

and forms a continuous pattern in both forward and backward regions.

The effect of different frequencies on the radiation patterns of a waveguide slot is also examined. The dimensional parameters of the slot and the ground plane are the same as those in the previous example. The E-plane radiation pattern of the slot is calculated at three frequencies and illustrated in Fig. 4.9. It can be seen that the radiation pattern changes only a little when the frequency changes from 9 GHz to 11 GHz. In fact, the change of the operating frequency can be considered as the change of electrical size of the ground plane, which will be studied in detail in the next section.

4.4 Numerical Results

Table 4.1: Front-to-back ratio of slot antennas on the broad wall of a partially dielectric-loaded waveguide with different wall sizes.

Size of the Waveguide Broad Wall	Front-to-Back Ratio (dB)	
	Longitudinal Slot	Transverse Slot
$120mm \times 120mm$ ($4\lambda \times 4\lambda$)	21.1	17.8
$240mm \times 240mm$ ($8\lambda \times 8\lambda$)	24.4	21.1
$480mm \times 480mm$ ($16\lambda \times 16\lambda$)	27.9	24.6

4.4.3 Effect of Different Wall Sizes on Waveguide Slots' Radiation Patterns

Further studies are carried out to examine the effect of different waveguide wall sizes on the radiation patterns of waveguide slot antennas. The partially dielectric-loaded waveguide slot antennas studied in the previous subsection are taken as examples. Figure 4.10 shows the radiation patterns of the longitudinal slot with different sizes of the broad wall. It can be seen that both E-plane and H-plane radiation patterns in the forward region do not change significantly when the ground plane size increases, though a larger ground plane introduces more ripples. However, the backward radiation is closely dependent upon the ground plane size. A large ground plane will result in more backward radiation lobes. Similar behavior is also observed for the transverse slot shown in Fig. 4.11.

The front-to-back ratio of both longitudinal and transverse slot antennas with different sizes of waveguide broad wall is listed in Table 4.1. It is seen that the front-to-back ratio becomes higher when the size of the ground plane increases, as we expected.

4.4 Numerical Results

(a) E-plane

(b) H-plane

Figure 4.10: Radiation patterns of a longitudinal slot antenna in a partially dielectric-loaded rectangular waveguide with different sizes of ground plane ($a = 15\text{mm}$, $b = 6\text{mm}$, $t = 3\text{mm}$, $w = 3\text{mm}$, $l = 14\text{mm}$, $d = 3.75\text{mm}$, $s = 1\text{mm}$, $\epsilon_{r1} = 6$, $\epsilon_{r2} = 1$, $f = 10\text{GHz}$).

4.4 Numerical Results

(a) E-plane

(b) H-plane

Figure 4.11: Radiation patterns of a centered transverse slot antenna in a partially dielectric-loaded rectangular waveguide with different sizes of ground plane ($a = 15\text{mm}$, $b = 6\text{mm}$, $t = 3\text{mm}$, $w = 15\text{mm}$, $l = 6\text{mm}$, $s = 1\text{mm}$, $\epsilon_{r1} = 6$, $\epsilon_{r2} = 1$, $f = 10\text{ GHz}$).

4.5 Conclusion

4.5 Conclusion

In this chapter, the effect of a finite ground plane on the radiation patterns of rectangular waveguide slot antennas has been studied by a hybrid GTD method. Both longitudinal and transverse slots have been considered. Calculated radiation patterns have been presented and compared with measured data or those obtained by HFSS. Excellent agreement has been observed. The hybrid GTD method has been demonstrated to be accurate and efficient for solving waveguide slot antennas with a finite ground plane. Comparing the radiation patterns of waveguide slot antennas with a finite ground plane to those with an infinite ground plane, it has been demonstrated that the finite ground plane has a small effect on E-plane and H-plane patterns of slot antennas on both air-filled and partially dielectric-loaded waveguides in the forward direction. It has also been illustrated that the ground plane size can noticeably change the backward radiation level as well as the front-to-back ratio of the slot antennas. Numerical simulations have shown that if the edges of the waveguide are asymmetric with respect to the slot, the E-plane pattern will exhibit a small squint while the main lobe of the H-plane pattern has only a slight change. This property can then be taken into consideration in designing compact waveguide slot antenna arrays.

Chapter 5

Mutual Coupling between Dielectric-loaded Waveguide Slots

5.1 Introduction

Waveguide slot antenna has been widely used in modern communication systems and airborne applications because it can be conveniently fabricated and flush-mounted on the surfaces of different systems. When broad wall slots are used in array applications, mutual coupling effects between slot elements must be considered to achieve the scattering and radiation characteristics accurately. Elliott presented a procedure for designing dielectric-loaded small shunt slot array [54] with the external mutual coupling effects taken into consideration. Shan and Shen [42] designed a slot array on the broad wall of a partially dielectric-loaded waveguide with the mutual coupling extracted from HFSS. They also presented a mode-matching analysis [43] for extracting the equivalent circuit parameters of transverse slots in the broad wall of a rectangular waveguide partially loaded with

5.2 Analysis Model

an H-plane slab. However, the extracted mutual impedance data in [43] have not been validated and the scattering parameters of the waveguide slots structure have not been studied in detail.

This chapter applies the hybrid full-wave method to analyze two slots cut in the broad wall of a rectangular waveguide partially loaded with a dielectric slab. Following the similar procedure in Chapter 3, the slot element can be considered as two separate parts of an open-ended waveguide and a waveguide T-junction. The two T-junctions are then cascaded to construct a four-port network. The problem consisting of two open-ended waveguides has been studied by many researchers [86]–[88]. In our approach, we use the same method as [88] to calculate the external mutual coupling of the two open-ended waveguides. Combining the results of the four-port network and the open-ended waveguides yields the scattering parameters of the whole waveguide slot structure.

5.2 Analysis Model

Figure 5.1 illustrates the slot structure considered in our study. The waveguide has a width a and a height b . There are two layers of dielectric materials in the waveguide with the relative permittivities being ε_{r1} and ε_{r2} , respectively. The thickness of the lower layer is t . Two rectangular slots are cut in the broad wall of the partially dielectric-loaded waveguide and their dimensions are (w_1, l_1) and (w_2, l_2) , respectively. Only rectangular slots are considered here for example, though our method can be easily extended to compound slot and other slots. The thickness of the waveguide wall is s . The displacements of the two slots from the center of

5.2 Analysis Model

Figure 5.1: Dielectric-loaded rectangular waveguide slots.

the waveguide are d_1 and d_2 , respectively. The separation between the two slots is D .

The equivalent model of the slot structure is shown in Fig. 5.2. It is seen that the whole structure can be separated into two parts: one is a two-port network having two open-ended waveguides and the other is a four-port network consisting of two waveguide T-junctions and a uniform waveguide section. The scattering matrix of the first network with two open-ended waveguides radiating into the half free-space can be written as [88]:

$$\mathbf{S}_{\text{open}} = 2(\mathbf{I} + \mathbf{Y})^{-1} - \mathbf{I}, \quad (5.1)$$

where $\mathbf{Y}$ is the admittance matrix and the details of its elements for self and mutual admittances can be found in Chapter 3 or [88], respectively. The scattering matrix

5.2 Analysis Model

Figure 5.2: Equivalent model of the dielectric-loaded waveguide slots.

$[\mathbf{S}_{\text{open}}]$ can also be expressed in a form of sub-matrices:

$$\mathbf{S}_{\text{open}} = \begin{bmatrix} \mathbf{S}_{11}^{\circ} & \mathbf{S}_{12}^{\circ} \\ \mathbf{S}_{21}^{\circ} & \mathbf{S}_{22}^{\circ} \end{bmatrix} \quad (5.2)$$

where $\mathbf{S}_{21}^{\circ}$ is the matrix of mutual coupling coefficients between the two open-ended waveguides.

Next, we characterize the four-port network. From the analysis in Chapter 3, the scattering matrices of the two dielectric-loaded waveguide T-junctions can be obtained as $[\mathbf{S}_{\mathbf{T}_1}]$ and $[\mathbf{S}_{\mathbf{T}_2}]$, respectively. Shifting the reference planes of Port 2'

5.2 Analysis Model

and Port 1'' to the center of the uniform waveguide section, two new scattering matrices are obtained:

$$\mathbf{S}'_{\mathbf{T}_1} = \mathbf{F}_1 \mathbf{S}_{\mathbf{T}_1} \mathbf{F}_1 \quad (5.3)$$

$$\mathbf{S}'_{\mathbf{T}_2} = \mathbf{F}_2 \mathbf{S}_{\mathbf{T}_2} \mathbf{F}_2 \quad (5.4)$$

where

$$\mathbf{F}_1 = \begin{bmatrix} \mathbf{I} & & \\ & \mathbf{e}^{-\gamma D/2} & \\ & & \mathbf{I} \end{bmatrix} \quad (5.5)$$

$$\mathbf{F}_2 = \begin{bmatrix} \mathbf{e}^{-\gamma D/2} & & \\ & \mathbf{I} & \\ & & \mathbf{I} \end{bmatrix} \quad (5.6)$$

where $\mathbf{I}$ is the identity matrix, γ is the propagation constant in the dielectric-loaded rectangular waveguide and D is the length of the waveguide section between the two T-junctions. Expressing $[\mathbf{S}'_{\mathbf{T}_1}]$ and $[\mathbf{S}'_{\mathbf{T}_2}]$ as:

$$\mathbf{S}'_{\mathbf{T}_1} = \begin{bmatrix} S_{11}^1 & S_{12}^1 & S_{13}^1 \\ S_{21}^1 & S_{22}^1 & S_{23}^1 \\ S_{31}^1 & S_{32}^1 & S_{33}^1 \end{bmatrix} \quad (5.7)$$

$$\mathbf{S}'_{\mathbf{T}_2} = \begin{bmatrix} S_{11}^2 & S_{12}^2 & S_{13}^2 \\ S_{21}^2 & S_{22}^2 & S_{23}^2 \\ S_{31}^2 & S_{32}^2 & S_{33}^2 \end{bmatrix}, \quad (5.8)$$

5.2 Analysis Model

the scattering matrix of the four-port network $[\mathbf{S}_{4\text{-port}}]$ is derived:

$$\begin{aligned}
 & \mathbf{S}_{4\text{-port}} \\
 &= \begin{bmatrix} \mathbf{S}_{11} & \mathbf{S}_{12} & \mathbf{S}_{13} & \mathbf{S}_{14} \\ \mathbf{S}_{21} & \mathbf{S}_{22} & \mathbf{S}_{23} & \mathbf{S}_{24} \\ \mathbf{S}_{31} & \mathbf{S}_{32} & \mathbf{S}_{33} & \mathbf{S}_{34} \\ \mathbf{S}_{41} & \mathbf{S}_{42} & \mathbf{S}_{43} & \mathbf{S}_{44} \end{bmatrix} \\
 &= \begin{bmatrix} \mathbf{S}_{11}^1 + \mathbf{S}_{12}^1 \mathbf{T}_1 \mathbf{S}_{11}^2 \mathbf{S}_{21}^1 & \mathbf{S}_{12}^1 \mathbf{T}_1 \mathbf{S}_{12}^2 & \mathbf{S}_{13}^1 + \mathbf{S}_{12}^1 \mathbf{T}_1 \mathbf{S}_{11}^2 \mathbf{S}_{23}^1 & \mathbf{S}_{12}^1 \mathbf{T}_1 \mathbf{S}_{13}^2 \\ \mathbf{S}_{21}^2 \mathbf{T}_2 \mathbf{S}_{21}^1 & \mathbf{S}_{22}^2 + \mathbf{S}_{21}^2 \mathbf{T}_2 \mathbf{S}_{22}^1 \mathbf{S}_{12}^2 & \mathbf{S}_{21}^2 \mathbf{T}_2 \mathbf{S}_{23}^1 & \mathbf{S}_{23}^2 + \mathbf{S}_{21}^2 \mathbf{T}_2 \mathbf{S}_{22}^1 \mathbf{S}_{13}^2 \\ \mathbf{S}_{31}^1 + \mathbf{S}_{32}^1 \mathbf{T}_1 \mathbf{S}_{11}^2 \mathbf{S}_{21}^1 & \mathbf{S}_{32}^1 \mathbf{T}_1 \mathbf{S}_{12}^2 & \mathbf{S}_{33}^1 + \mathbf{S}_{32}^1 \mathbf{T}_1 \mathbf{S}_{11}^2 \mathbf{S}_{23}^1 & \mathbf{S}_{32}^1 \mathbf{T}_1 \mathbf{S}_{13}^2 \\ \mathbf{S}_{31}^2 \mathbf{T}_2 \mathbf{S}_{21}^1 & \mathbf{S}_{32}^2 + \mathbf{S}_{31}^2 \mathbf{T}_2 \mathbf{S}_{22}^1 \mathbf{S}_{12}^2 & \mathbf{S}_{31}^2 \mathbf{T}_2 \mathbf{S}_{23}^1 & \mathbf{S}_{33}^2 + \mathbf{S}_{31}^2 \mathbf{T}_2 \mathbf{S}_{22}^1 \mathbf{S}_{13}^2 \end{bmatrix} \quad (5.9)
 \end{aligned}$$

where $\mathbf{T}_1 = (\mathbf{I} - \mathbf{S}_{11}^2 \mathbf{S}_{22}^1)^{-1}$ and $\mathbf{T}_2 = (\mathbf{I} - \mathbf{S}_{22}^2 \mathbf{S}_{11}^1)^{-1}$. In this approach, the effect of the finite thickness of the waveguide wall as well as the internal mutual coupling due to higher order modes is automatically included.

After $[\mathbf{S}_{\text{open}}]$ and $[\mathbf{S}_{4\text{-port}}]$ are obtained, the scattering parameters of the whole waveguide slot structure can then be written in the form:

$$\mathbf{S}_{2\text{P}} = \begin{bmatrix} \mathbf{S}_{11} + \mathbf{S}_{13} \mathbf{W}_{11} + \mathbf{S}_{14} \mathbf{W}_{21} & \mathbf{S}_{12} + \mathbf{S}_{13} \mathbf{W}_{12} + \mathbf{S}_{14} \mathbf{W}_{22} \\ \mathbf{S}_{21} + \mathbf{S}_{23} \mathbf{W}_{11} + \mathbf{S}_{24} \mathbf{W}_{21} & \mathbf{S}_{22} + \mathbf{S}_{23} \mathbf{W}_{12} + \mathbf{S}_{24} \mathbf{W}_{22} \end{bmatrix} \quad (5.10)$$

5.3 Numerical Results

where

$$\begin{aligned} \mathbf{W} = & \mathbf{S}_{\text{open}} \begin{bmatrix} \mathbf{I} - \mathbf{S}_{33}\mathbf{S}_{11}^{\circ} - \mathbf{S}_{34}\mathbf{S}_{21}^{\circ} & -(\mathbf{S}_{33}\mathbf{S}_{12}^{\circ} + \mathbf{S}_{34}\mathbf{S}_{22}^{\circ}) \\ -(\mathbf{S}_{43}\mathbf{S}_{11}^{\circ} + \mathbf{S}_{44}\mathbf{S}_{21}^{\circ}) & \mathbf{I} - \mathbf{S}_{43}\mathbf{S}_{12}^{\circ} - \mathbf{S}_{44}\mathbf{S}_{22}^{\circ} \end{bmatrix}^{-1} \\ & \times \begin{bmatrix} \mathbf{S}_{31} & \mathbf{S}_{32} \\ \mathbf{S}_{41} & \mathbf{S}_{42} \end{bmatrix} \end{aligned} \quad (5.11)$$

The same technique can be easily extended to model the structure with more than two open-ended waveguides. Therefore, the slot structure consisting of more than two slots can also be analyzed in the similar procedure.

5.3 Numerical Results

A computer program has been developed to calculate the scattering parameters of the dielectric-loaded waveguide slots based on the approach described in the previous section. In order to verify the validity of our hybrid technique, the results obtained by our method are compared to the simulated data by HFSS. Good agreement between them are observed for all the cases considered.

5.3.1 Mutual Coupling between Two Open-ended Waveguides

Two rectangular waveguides radiating into the half space are investigated firstly. The geometry of the two rectangular waveguides is illustrated in Fig. 5.3. An infinite ground plane is assumed in our analysis. The waveguides have the dimen-

5.3 Numerical Results

Figure 5.3: Geometry of two coupled rectangular waveguides.

sions of (a_1, b_1) and (a_2, b_2) , respectively. The distances between them for the two orientations are D_x and D_y , respectively.

First, two identical WR-90 waveguides placed in the collinear location are studied. The dimensions of the two waveguides are $a_1 = a_2 = 22.86mm$ and $b_1 = b_2 = 10.16mm$. The distances between them are $D_x = 25.4mm$ and $D_y = 0mm$, respectively. In our calculation, 12 modes are employed for representing the field distribution along the waveguide openings. The mutual coupling coefficient [86, 88], which is $\mathbf{S}_{21}^o$ in the Eq. 5.2, is calculated by our method and plotted in Fig. 5.4. Results obtained by HFSS are also presented for comparison. Excellent agreement is observed for both the magnitude and the phase of the coupling coefficient. To achieve the similar accuracy, the CPU times of our method and HFSS for a single frequency are 8 seconds and 20 minutes, respectively. It can be seen that the integration method used in our formulation is much more efficient

5.3 Numerical Results

Figure 5.4: Mutual coupling between two rectangular waveguides ($a_1 = a_2 = 22.86$ mm, $b_1 = b_2 = 10.16$ mm, $D_x = 25.4$ mm, $D_y = 0$ mm).

5.3 Numerical Results

(a) Magnitude

(b) Phase

Figure 5.5: Mutual coupling between two rectangular waveguides ($a_1 = a_2 = 10.16$ mm, $b_1 = b_2 = 22.86$ mm, $D_x = 12.7$ mm, $D_y = 0$ mm).

5.3 Numerical Results

than the FEM used in HFSS.

The two rectangular waveguides are then rotated by 90 degrees and the distance between them is also reduced. They remain in the collinear configuration. This is because for the noncollinear situation an extremely long computation time is required by HFSS to obtain a good accuracy. The mutual coupling coefficients are calculated and plotted in Fig. 5.5. Again, an excellent agreement between our method and HFSS is observed. It is seen that the magnitude of the coupling coefficient is much larger in this example compared to the previous one. This is due to the fact that, for rectangular waveguides, the broad side coupling is stronger than the narrow side coupling.

5.3.2 Mutual Coupling between Three or More Open-ended Waveguides

The structure with more than two open-ended rectangular waveguides mounted on the same infinite ground plane is studied next. For the purpose of convenience, WR-90 waveguides are employed in all our calculations.

Three WR-90 waveguides are placed in the collinear location, as depicted in Fig. 5.6(a). The magnitude of the coupling coefficients between them is shown in Fig. 5.6(b). It is observed that the mutual coupling between waveguides #1 and #3 is much smaller than that between #1 and #2, as we expected. Coupling coefficients obtained by HFSS are also presented. It can be observed that our results agree very well with HFSS. The phase of the coupling coefficients is not compared because it

5.3 Numerical Results

Figure 5.6: (a) Geometry and (b) coupling coefficients of three collinear rectangular waveguides with narrow side coupling ($a = 22.86mm$, $b = 10.16mm$ and $D = 25.4mm$).

is extremely time consuming for HFSS to obtain a good convergence.

Next, we rotate the three waveguides by 90 degrees with respect to their centers and shorten the distance between them. The three waveguides are still in collinear with each other, as illustrated in Fig. 5.7(a). Magnitude of coupling coefficients are shown in Fig. 5.7(b). HFSS data are also provided for comparison. Again

5.3 Numerical Results

Figure 5.7: (a) Geometry and (b) coupling coefficients of three collinear rectangular waveguides with broad side coupling ($a = 10.16mm$, $b = 22.86mm$ and $D = 12.7mm$).

an excellent agreement between them is observed. In this example, the distance between waveguides #1 and #3 is $25.4mm$, which is the same as the distance between #1 and #2 in the previous example. Comparing S_{13} in Fig. 5.7(b) with S_{12} in Fig. 5.6(b), it is seen obviously that broad side coupling is stronger than narrow side coupling between collinear rectangular waveguides.

5.3 Numerical Results

Figure 5.8: (a) Geometry and (b) coupling coefficients of four rectangular waveguides ($a = 22.86\text{mm}$, $b = 10.16\text{mm}$, $D_1 = 25.4\text{mm}$ and $D_2 = 12.7\text{mm}$).

An array consisting of four open-ended rectangular waveguides is also considered, as depicted in Fig. 5.8(a). Due to the symmetry of the structure, only three mutual coupling coefficients S_{12} , S_{13} and S_{14} are given in Fig. 5.8(b). The agreement between our data and those obtained by HFSS is very good. It is

5.3 Numerical Results

observed that S_{13} has the smallest magnitude among these three coupling coefficients. Therefore, it can be concluded that the shorter the distance between the open-ended waveguides, the smaller the magnitude of the coupling coefficient.

5.3.3 Two Slots Cut on Partially Dielectric-loaded Rectangular Waveguide

Two identical longitudinal slots cut in the broad wall of a partially dielectric-loaded rectangular waveguide are studied. The waveguide has an inner dimension of $15\text{mm} \times 6\text{mm}$. A dielectric slab with a thickness of 3mm is placed at the bottom of the waveguide. The relative permittivity of the slab is 6. Each slot has a dimension of $3\text{mm} \times 14\text{mm}$. In our full-wave analysis, 10 modes are used to represent the field distribution along the slot. The scattering parameters calculated by our method are compared to those obtained by HFSS, as shown in Fig. 5.9. Excellent agreement is observed. The computation time of our hybrid method is 9 second for one frequency point, while the CPU time for HFSS is nearly 20 minutes on the same computer (10 iterations). It is clear that our hybrid technique is very efficient compared to HFSS that is based on the finite-element method. From Fig. 5.9 it can also be seen that the distance between the two slot elements can greatly change the results of the scattering parameters. This is due to the fact that the mutual coupling effect is sensitive to the distance between the two elements.

Two identical transverse slots on the same dielectric-loaded waveguide are also studied. The slots have a dimension of $15\text{mm} \times 6\text{mm}$. The results obtained by our method are plotted in Fig. 5.10 and compared with HFSS. Again a good

5.3 Numerical Results

(a)

(b)

Figure 5.9: Scattering parameters of two longitudinal slots in a rectangular waveguide partially loaded with a dielectric slab ($a = 15mm$, $b = 6mm$, $t = 3mm$, $s = 1mm$, $\epsilon_{r1} = 6$, $\epsilon_{r2} = 1$, $w_1 = w_2 = 3mm$, $l_1 = l_2 = 14mm$, $d_1 = d_2 = 3.75mm$); (a) $D = 5mm$, and (b) $D = 10mm$.

5.3 Numerical Results

(a)

(b)

Figure 5.10: Scattering parameters of two transverse slots in a rectangular waveguide partially loaded with a dielectric slab ($w_1 = w_2 = 15mm$, $l_1 = l_2 = 6mm$, $d_1 = d_2 = 0mm$; other parameters are the same as in Fig. 5.9); (a) $D = 5mm$, and (b) $D = 10mm$.

5.3 Numerical Results

Figure 5.11: Block diagram of two slot antennas in the broad wall of a partially dielectric-loaded rectangular waveguide.

agreement between them is observed. It is also observed from Fig. 5.10(a) and (b) that the mutual coupling has a great effect on the slots' scattering parameters, as we expected.

5.3.4 Circuit Parameters of Two Slots Structure

In order to study in more details the partially dielectric-loaded waveguide slot antennas, we examine the mutual coupling effect between slot antennas by extracting the circuit parameters from the obtained results.

From Fig. 5.2, we can see that the whole waveguide slots structure can be represented by the block diagram shown in Fig. 5.11. The scattering matrix $\mathbf{S}_{\text{open}}$ of the two port network is actually the mutual coupling between two open-ended waveguide, which has been investigated in previous sections. The $\mathbf{S}_{\text{T1}}$, $\mathbf{S}_{\text{T2}}$ and $\mathbf{S}_{\text{L}}$ represent the scattering matrices of the two T-junctions and the transmission line between the two slots, respectively. All these blocks have been solved in the

5.3 Numerical Results

Figure 5.12: Circuit model of two waveguide slot antennas.

previous section.

To extract the mutual impedance between the two slots, we introduce a circuit model shown in Fig. 5.12. In this model Z_1' and Z_2' are the impedances of the slots, and they already take the mutual coupling into account.

$$\begin{aligned} Z_1' &= Z_1 + \frac{I_2}{I_1} Z_{12} \\ Z_2' &= Z_2 + \frac{I_1}{I_2} Z_{22} \end{aligned} \quad (5.12)$$

where I_1/I_2 can be derived as

$$\frac{I_1}{I_2} = -Z_2' \sinh(\beta l) + \cosh(\beta l) \quad (5.13)$$

Impedances Z_1' and Z_2' can be obtained from the solved structure shown in Fig. 5.11. The self-impedance of the two slots Z_1 and Z_2 can be solved separately. Therefore, the mutual impedance Z_{21} (Z_{12}) can be readily calculated using the

5.3 Numerical Results

Figure 5.13: Frequency dependency of the mutual impedance between two transverse slots in the broad wall of a partially dielectric-loaded waveguide (Distance between the two slots is 20 mm).

following equations:

$$\begin{aligned}
 Z_{12} &= (Z'_1 - Z_1) (\cosh(\beta l) - Z'_2 \sinh(\beta l)) \\
 Z_{21} &= \frac{Z'_2 - Z_2}{\cosh(\beta l) - Z'_2 \sinh(\beta l)}
 \end{aligned} \tag{5.14}$$

The obtained mutual impedances Z_{12} and Z_{21} should be identical since the structure is reciprocal.

Figure 5.13 illustrates the frequency dependency of the mutual resistance and

5.3 Numerical Results

Figure 5.14: Frequency dependency of the mutual impedance between two transverse slots in the broad wall of a partially dielectric-loaded waveguide (Distance between the two slots is 60 mm).

reactance between two identical transverse radiating slots. The slots are cut on the wall of the same partially-loaded waveguide, as illustrated in Fig 5.1. The dimensions of the waveguide are as follows: $a = 15\text{mm}$, $b = 6\text{mm}$, $s = 1\text{mm}$. The dielectric slab has a relative permittivity of 6 and a thickness of 3mm. The transverse slots have dimensions of $w_1 = w_2 = 15\text{mm}$ and $l_1 = l_2 = 6\text{mm}$, respectively. It is observed that the mutual resistance reaches its peak at 9.9 GHz and the mutual reactance is quite small at that frequency point.

Another example is presented in Fig. 5.14 showing the normalized mutual

5.3 Numerical Results

impedance when the distance between the same two slots is increased from $20mm$ to $60mm$. It can be seen that the mutual resistance reaches its peak at 10.2 GHz . It can also be observed that the peak value of mutual resistance in Fig. 5.14 is less than 0.1 , which is much lower than that in Fig. 5.13. This shows that a large distance between the slots may decrease the mutual coupling between them. It can also be concluded that changing the distance between the slots will shift the peak of the mutual resistance.

Figure 5.15 illustrates the variation of the mutual impedance with respect to the distance between the two slots. The geometrical parameters are the same as those used in Figs. 5.13 and 5.14. It can be seen that the magnitude of the mutual impedance between the two slots fluctuates periodically and becomes weaker when the distance increases, as we expected.

5.3.5 An Array of Three Slots

After having studied the waveguide section consisting of two slots, we also conduct the analysis for an array of three slots cut on a same rectangular waveguide partially loaded with a dielectric slab. For the purpose of convenience, we choose the same waveguide as discussed in most previous examples in this chapter. The dimensional parameters are given as: $a = 15mm$, $b = 6mm$, $s = 1mm$, $t = 3mm$, $\epsilon_{r1} = 6$, $\epsilon_{r2} = 1$. To compare with the case of two slots cut on the wall of the waveguide, which has been presented in Fig. 5.10, the dimensions of the slots are chosen to be $w_1 = w_2 = w_3 = 15mm$ and $l_1 = l_2 = l_3 = 6mm$, as illustrated in Fig. 5.16. Usually, the distances D_1 and D_2 are set to be identical to form a uniform array.

5.3 Numerical Results

(a) Magnitude

(b) Phase

Figure 5.15: Magnitude and phase of the mutual impedance versus slot separation (Structural parameters are $a = 15\text{mm}$, $b = 6\text{mm}$, $s = 1\text{mm}$, $t = 3\text{mm}$, $\epsilon_{r1} = 6$, $\epsilon_{r2} = 1$, $w_1 = w_2 = 15\text{mm}$, $l_1 = l_2 = 6\text{mm}$).

5.3 Numerical Results

Figure 5.16: Geometry of a linear array consisting of three slots.

First we set $D_1 = D_2 = 11mm$, where the inter-element spacing is the same as that used in Fig. 5.10(a). The scattering parameters of the whole waveguide slot structure are calculated by our method, as shown in Fig. 5.17. The simulated data by HFSS are also presented. Excellent agreement between them is observed. Comparing Fig. 5.17 with Fig. 5.10(a), it can be seen that the scattering parameters change a little after one more slot is cut on the waveguide wall. This is due to the combining of internal and external mutual coupling between the slots. The internal mutual coupling is introduced by the higher modes interaction in the waveguide sections between the slots. The external mutual coupling is caused by the radiation from the slots to the outer space. It is also noticed that in both cases, the return loss and the insertion loss is very high. Assuming all the materials in the structure are lossless, the normalized radiated power can be calculated according to the following equation:

$$P_{rad} = 1 - (|S_{11}|^2 + |S_{21}|^2). \quad (5.15)$$

5.3 Numerical Results

Figure 5.17: Scattering parameters of three transverse slots cut on a partially dielectric-loaded rectangular waveguide ($D_1 = D_2 = 11mm$).

From Fig. 5.10(a) and Fig. 5.17, it is seen that only about 50 percent of the input power entering the waveguide port can be radiated by the slot antenna.

To reduce the return loss and the insertion loss and in turn enhance the power radiating efficiency, the inter-element spacing should be changed. In our calculation, it is found that only after we adjusted the spacings to $D_1 = D_2 = 19mm$, good results can be achieved. This is because the internal mutual coupling varies when the inter-element spacing changes. For certain values of the spacings, the power fed into the waveguide can be efficiently delivered to the slot radiators and therefore the return and insertion loss are improved. The computed scattering

5.3 Numerical Results

Figure 5.18: Scattering parameters of three transverse slots cut on a partially dielectric-loaded rectangular waveguide ($D_1 = D_2 = 19mm$).

parameters are shown in Fig. 5.18. It can be calculated from Eq. 5.15 that about 90 percent input power can be radiated into the free space by the slots. However, when the spacings are set to $19mm$, the E-plane radiation pattern has a squint for the main beam. Figure 5.19 illustrates the effect of the inter-element spacing of this three-slot array on the E-plane radiation patterns. It is observed that the E-plane pattern changes significantly when the spacing varies from $11mm$ to $19mm$. Although a spacing of $19mm$ optimizes the radiated power, the direction of the main beam has a squint of 20° with respect to the broadside. This is due to the fact that this spacing is not close to one guide wavelength λ_g in the waveguide and the three slots are excited with different phases. When the inter-element spacing is

5.4 Conclusions

Figure 5.19: E-plane radiation patterns of three transverse slots cut on a partially dielectric-loaded rectangular waveguide ($f = 10$ GHz).

set to 24.8mm , which equals λ_g at 10 GHz, the main beam of the E-plane radiation pattern points to the broadside, as we expected.

5.4 Conclusions

This chapter has proposed a hybrid full-wave technique for characterizing the mutual coupling between rectangular slots cut in the wall of a rectangular waveguide partially loaded with a dielectric slab. The internal and external mutual coupling has been theoretically taken into consideration in this approach. Numerical results for the mutual coupling coefficients between two or more rectangular waveguides have been presented. It has been shown that the larger the distance between the waveguide openings, the weaker the mutual coupling between them. Broad side coupling between rectangular waveguides has been demonstrated to be stronger

5.4 Conclusions

than narrow side one. The scattering parameters of a partially dielectric-loaded rectangular waveguide with two slots cut on its broad wall have been calculated. Circuit model for such structure has been introduced and the mutual impedance between the slots has been extracted. It has been shown that the magnitude of the mutual impedance becomes smaller when the distance between the two slots increases. A linear array consisting of three transverse slots cut on the wall of the same partially dielectric-loaded rectangular waveguide has also been analyzed. It has been illustrated that the scattering parameters vary when the distance between the slots changes. When the element spacing is carefully adjusted, both the return loss and the insertion loss can be very small and most of the power fed into the waveguide is radiated to the free space.

In our analysis, simulated data obtained by HFSS have also been provided for comparison. Good agreement has been observed for all the cases considered. The hybrid method proposed in Chapter 3 has been verified to be very accurate and efficient again. It can be readily extended to analyze and design slot array antennas.

Chapter 6

Dielectric-Loaded Waveguide Slot Antenna Arrays

6.1 Introduction

As early as in 1938, Hansen and Woodyard [89] proposed the phase shift requirements for an infinitely long antenna with continuous distribution. These requirements also yield good approximations for very long uniform arrays with closely spaced elements. They were later known as the Hansen-Woodyard conditions for end-fire antenna arrays. The directivity of the Hansen-Woodyard array can be enhanced without destroying any of the other characteristics compared to arrays with uniform amplitude and spacing. This was probably the earliest innovation made on the conventional uniform antenna arrays. However, controlling the phase of the excitations accurately is a hard task, especially for arrays operating at high frequencies.

6.1 Introduction

The amplitude distribution had also been considered as an important factor to control the radiation performances of antenna arrays. Non-uniform excitations for the array with uniform spacing had been considered. Most of these discussions were focused on binomial and Dolph-Tschebyscheff arrays. Stone [90] suggested to use the coefficients of binomial expansion as the excitation amplitude. Later, Ricardi [91] derived the approximate closed-form expressions for the half-power beamwidth (HPBW) and maximum directivity for the $d = \lambda/2$ spacing. Although binomial arrays generally have very small side lobes, or even have no side lobes in the particular cases when element spacing equal to or less than $\lambda/2$, they always have larger HPBW compared to uniform arrays. Therefore, one must compromise between the side lobe level and the beamwidth. Another array, Dolph-Tschebyscheff, have more practical applications than the binomial array. It is originally introduced by Dolph [92] and then studied by many other researchers [93]–[95]. The excitation coefficients are related to Tschebyscheff polynomials. Dolph-Tschebyscheff array is a compromise between uniform and binomial arrays, where a Dolph-Tschebyscheff array with no side lobes becomes to the binomial one. Both the binomial and the Dolph-Tschebyscheff array have been employed in forming scanning arrays. However, the realization of excitations with non-uniform amplitude is rather complicated and the gain of the array is lower compared to uniform ones.

Another means of controlling the radiation performance is adjusting the inter-element spacing of the array. When the array elements are placed non-uniformly, the required radiation characteristics such as scanning the main beam, suppressing the side lobes and inserting additional nulls can be easily achieved by tuning the inter-element spacings. This measure also has another attractive advantage over

6.2 Design Considerations

others in that its physical realization is very simple. A few researchers [96]–[99] have investigated the spacing synthesis based on the mean square value or similar algorithms. However, like some other research works [100, 101] in which the amplitude and the phase of the excitation were also included, all the previous studies were limited to theoretical analysis. In fact, the obtained radiation patterns were the array factor patterns only, not including the element’s radiation characteristics. When we examine practical array applications, the results could be quite different due to the mutual coupling between elements. In addition, many desired patterns may not be achieved due to physical constraints, e.g. an inter-element spacing of 0.1λ is often impossible in practice. Therefore, the design of array depends not only on the array factor synthesis but also a full-wave analysis of the whole structure. Physical constraints should also be taken into consideration.

In this chapter, we aim at designing an array with low side lobe level by adjusting the inter-element spacing between the slot element while keeping the elements excited uniformly. Based on the studies in previous chapters, slot cut on the wall of a rectangular waveguide partially filled with a dielectric slab is chosen as the element of the array. Full-wave analysis is performed to account for the mutual coupling effect.

6.2 Design Considerations

6.2.1 Choosing the Slot Element

The frequency performance of slot antenna array cut on a rectangular waveguide has been examined by several researchers [102]–[104]. The slot arrays can be

6.2 Design Considerations

Figure 6.1: (a) Standing-wave and (b) traveling-wave longitudinal slot arrays.

classified into two types: standing-wave slot arrays and traveling-wave slot arrays [105].

For a standing-wave slot array, the slots are located at the peaks of the standing wave. A typical distribution of longitudinal slots cut on air-filled rectangular waveguide is depicted in Fig. 6.1(a). Because the guided wavelength λ_g in an air-filled rectangular waveguide is larger than that in free space, grating lobes will rise if all the longitudinal slots are placed at one side of the broad wall with an inter-element spacing of λ_g . This problem is overcome by interlacing the slot elements along the waveguide wall and setting the distance between the adjacent

6.2 Design Considerations

slots to $\lambda_g/2$. All the slots then have the same excitation amplitude and phase under this arrangement and the grating lobes are avoided. This technique has been prevalently used for designing waveguide slot arrays. When a number of slots are placed in series on the waveguide wall, the return loss bandwidth is often quite small due to the rapid change of the inter-element spacing with frequency. Generally, the greater the number of the slots, the smaller is the bandwidth [102]. The bandwidth can be improved when the number of slots in one waveguide is reduced and the subarrays fed by split waveguide branches are employed [106].

For a traveling-wave slot array, the slots are cut on the wall of a waveguide transmitting a traveling wave. Figure 6.1(b) illustrates a traveling wave longitudinal slot array. In this case, the return loss bandwidth is usually much wider than the standing-wave slot array. However, the main beam of radiation pattern is shifting when the operating frequency changes. This is also due to the change of the inter-element spacing with frequency. Although a center feeding technique [107, 108] can be used to keep the squint angle invariant within a frequency range for the main beam, the deterioration of the radiation pattern and the degradation of antenna gain are inevitable when the operating frequency changes.

Transverse slots cut on the broad wall of a rectangular waveguide are not commonly used in array applications, although they have symmetric radiation patterns with main radiation beam pointing in the broadside stably. In traveling wave arrays, which usually have an off-broadside radiation, non-resonant transverse slots have found some applications [16, 20, 109]. However, resonant transverse slots in the wall of an air-filled rectangular waveguide are not suitable candidates for

6.2 Design Considerations

broadside radiation. This is due to the fact that the technique of interlacing the slots along the waveguide cannot be applied to transverse slots. Therefore, the inter-element spacing, which is the guided wavelength in the waveguide, is larger than one free-space wavelength. That will result in grating lobes and increase the array size. Using grating lobe filter [110] or partially loading the waveguide with dielectric materials [39, 39] can effectively prevent the occurring of grating lobes, but the bandwidth of a transverse slot array is usually very narrow. Recently, a transverse slot array in a partially dielectric-loaded rectangular waveguide has been presented [42]. The researchers developed a novel feeding structure and designed a 2×4 transverse slot array based on the simulation study using HFSS. The slot array has a compact size as well as a wide bandwidth. However, the design procedure is mainly dependent on the commercial software, which is inefficient and very time-consuming.

Based on the above investigations, we decided to use transverse slot as array element because of its attractive feature of symmetric and stable radiation pattern. Rectangular waveguide partially loaded with a dielectric slab is also employed in our design to shorten the guided wavelength and reduce the array size. In the previous discussions, it is understood that the more slots cut on one waveguide, the smaller is the return loss bandwidth. Therefore, only one element is used in each waveguide to obtain the maximum bandwidth available. To ensure that most of the power transmitted to the slot is emitted to outer space, standing-wave slot is adopted in our consideration, as shown in Fig. 6.2.

6.2 Design Considerations

Figure 6.2: Transverse slot element in the broad wall of a rectangular waveguide partially loaded with a dielectric slab.

6.2.2 Slot Element Characteristics

The dimensional parameters of the transverse slot element are illustrated in Fig. 6.2. The rectangular waveguide is of width a and height b . An H-plane dielectric slab is placed on the bottom wall of the waveguide with a thickness t and a relative permittivity ϵ_r . The thickness of the waveguide wall is s . The size of the slot cut in the broad wall of the waveguide is $w \times l$. The waveguide is short circuited at a distance d_1 away from the slot. The offset of the slot from the center of the waveguide is d_2 . Reference plane T is placed at the center of the slot for the purpose of analysis.

The equivalent circuit of the transverse slot element on a waveguide transmitting traveling wave is shown in Fig. 6.3(a). The slot is represented by a T-type network consisting of two series and one shunt impedances. The two se-

6.2 Design Considerations

Figure 6.3: Equivalent circuit of a transverse slot in the wall of a partially dielectric-loaded rectangular waveguide.

ries impedance are always the same, which is shown as $\frac{1}{2}(R_1 + jX_1)$. The shunt impedance has a form of $R_2 + jX_2$. Usually, the resistance R_2 is very small and can be approximated by zero, while the reactance X_2 is very large and considered as infinite. In many research works including those addressed in Chapter 3, the equivalent circuit of a transverse slot element cut in the waveguide wall is generally simplified and consists of only one series impedance $R_1 + jX_1$ if the slot width is electrically narrow.

6.2 Design Considerations

To obtain the required standing-wave slot antenna, the partially dielectric-loaded rectangular waveguide is shorted at one end. The equivalent circuit is illustrated in Fig. 6.3(b). The input impedance at plane T can be calculated using the following formula [42]:

$$Z_{in} = \frac{-2X_2 \tan(\beta d_1) - X_1 X_2 + jR_1 X_2}{R_1 + j[X_1 + 2 \tan(\beta d_1) + 2X_2]} + \frac{R_1 + jX_1}{2}, \quad (6.1)$$

Because R_2 is usually neglected and X_2 is very large compared to R_1 and X_1 , we choose R_1 to be nearly 1 and X_1 to be close to 0. In order to place the transverse slot at the peak of the standing wave, the distance d_1 is set to nearly half a guided wavelength. By making these configurations, a good return loss at the excitation port can be expected.

A dielectric material with a relative permittivity of 3.5 was available. Therefore, we first fix one parameter ϵ_r to 3.5. Utilizing the hybrid technique developed in Chapter 3 as the analyzing tool, the task of adjusting the other dimensional parameters is easily accomplished to achieve the desired values of R_1 , X_1 and X_2 . Figure 6.4 depicts the frequency dependence of normalized resistance and reactance with optimized geometrical parameters. To achieve a broad bandwidth, the slot dimensions w and l are chosen to be $13.6mm$ and $6mm$, respectively. The waveguide dimensions are adjusted simultaneously and obtained as follows: $a = 15mm$, $b = 7mm$, $t = 4.8mm$, $s = 1mm$. It is seen from Fig. 6.4 that the slot structure with above dimensional parameters meet well with our requirements around the center frequency of 10 GHz. It is also demonstrated that the shunt resistance R_2 and reactance X_2 can be approximated by zero and infinite, respectively. This

6.2 Design Considerations

Figure 6.4: Normalized resistance and reactance components of a transverse slot on the broad wall of a partially dielectric-loaded rectangular waveguide ($w = 13.6mm$, $l = 6mm$, $a = 15mm$, $b = 7mm$, $t = 4.8mm$, $s = 1mm$, and $\epsilon_r = 3.5$).

dielectric-loaded waveguide slot can also be represented by a π -network, as illustrated in Fig. 6.5(a). The normalized conductance and susceptance are extracted from scattering parameters and shown in Fig. 6.5(b).

When the transverse slot is mounted on the broad wall of the partially dielectric-loaded waveguide with one end shorted, the return loss at the excitation port (as shown in Fig. 6.3(b)) is presented in Fig. 6.6. The distance between the slot and the shorted end is $d_1 = 11.6mm$ and the displacement of the slot from the center of waveguide is $d_2 = 0mm$. It is observed that the 10 dB return

6.2 Design Considerations

(a) Circuit representation of the slot element

(b) Normalized conductance and susceptance components

Figure 6.5: A transverse slot represented by a π -network. (Parameters are the same as in Fig. 6.4.)

loss bandwidth, which is from 9.2 GHz to 10.8 GHz, is about 16% for this single radiating slot element. The input impedance Z_{in} is immediately extracted from the return loss data and its resistance and reactance components R_{in} and X_{in} are plotted in Fig. 6.7. A normalized resistance near 1 and a reactance close to 0 are observed around the center frequency 10 GHz, which explains the good matching

6.2 Design Considerations

Figure 6.6: Return loss of the slot element on the wall of a partially dielectric-loaded rectangular waveguide with one end shorted ($d_1 = 11.6\text{mm}$ and $d_2 = 0\text{mm}$. Other parameters are the same as in Fig. 6.4).

at the excitation port. Another method of substituting the data from Fig. 6.4 into Eq. 6.1 is also employed to calculate the input impedance Z_{in} and the results are given in Fig. 6.7 for comparison. Good agreement between the data obtained by these two methods is observed. This shows when the distance d_1 is carefully adjusted to locate the slot at the peak of the standing wave, the circuit model illustrated in Fig. 6.3(b) is very accurate in predicting the one port response for the radiating slot element. Therefore, Eq. 6.1 provide a good guideline for designing such slot elements.

6.2 Design Considerations

Figure 6.7: Input impedance of a single slot element (Parameters are the same as in Fig. 6.6).

For the above designed radiating slot element, the normalized guide wavelength (λ_g/λ_0) of the dominant mode in the partially dielectric-loaded rectangular waveguide is examined and plotted versus frequency in Fig. 6.8. It can be seen that the normalized guided wavelength is about 0.86 at the center frequency of 10 GHz. The guide wavelength λ_g can be further shortened by adjusting the inner dimensions of the waveguide and the thickness and dielectric constant of the loaded dielectric material simultaneously. When such dielectric-loaded waveguide is employed in array applications consisting of several series transverse slots, the element spacing is guaranteed to be less than λ_0 and the grating lobes can be effectively prevented.

6.2 Design Considerations

Figure 6.8: Normalized guided wavelength of the dominant mode in a partially dielectric-loaded rectangular waveguide. ($a = 15\text{mm}$, $b = 7\text{mm}$, $t = 4.8\text{mm}$, and $\epsilon_r = 3.5$).

6.2.3 Mutual Coupling between Slot Elements

It is noticed that the designed waveguide slot element has a compact size compared to conventional air-filled waveguide slots, which is due to the dielectric loading of the rectangular waveguide. The broad side of the partially dielectric-loaded waveguide working at frequencies near 10 GHz is only 15mm , while that of the standard X-band waveguide WR-90 is 22.86mm . To appreciate this attractive advantage, we decide to arrange all the waveguides side by side to form a slot array, as described in Fig. 6.9(a). Under this configuration, we have enough freedom to adjust the inter-element spacing d to achieve a desired radiation pattern

6.2 Design Considerations

Figure 6.9: Geometry and equivalent circuit of two coupling slot elements.

as well as account for the mutual coupling effect.

Figure 6.9(a) shows an example of array having two radiating slot. These two elements are identical and separated by a distance d . This structure can be represented by a simple equivalent circuit shown in Fig. 6.9(b) when we look into the reference plane S . Z_1 and Z_2 are the self impedance and Z_{12} is the mutual impedance between the two slot elements. These circuit parameters can then be extracted from the scattering parameters calculated by our hybrid method. Figure 6.10 illustrates the scattering parameters of this two port system when excited by waveguide port. The distance d between the two elements is $18mm$. It is seen that the mutual coupling between the two ports is about -20 dB near the center

6.2 Design Considerations

Figure 6.10: Scattering parameters of the structure consisting of two slot elements ($a = 15\text{mm}$, $b = 7\text{mm}$, $t = 4.8\text{mm}$, $s = 1\text{mm}$, $\epsilon_r = 3.5$, $w = 13.6\text{mm}$, $l = 6\text{mm}$, $d_1 = 11.6\text{mm}$, $d_2 = 0\text{mm}$, $d = 18\text{mm}$).

frequency 10 GHz. This level of mutual coupling does not have great effect on the return loss of the excitation port, as we can compare Fig. 6.10 with Fig. 6.6. However, it can greatly change the radiation patterns of the slot array. It is also observed that the agreement between our data and those from HFSS is very good. The extracted mutual impedance between these two slots are plotted in Fig. 6.11. It can be seen that the mutual resistance reaches its peak around 10 GHz while the mutual reactance is close to zero at that point. To include the mutual coupling effect in the design, the dimensions of each slot element can be slightly tuned. Another method is to adjust the spacings between the element. In our

6.3 Geometry of the Antenna Array

Figure 6.11: Mutual impedance between two slot elements (Parameters are the same as in Fig. 6.10).

consideration, to construct a slot array with low side lobes, only the inter-element spacings are adjusted while the dimensions of the slot element are fixed for the convenience of fabrication.

6.3 Geometry of the Antenna Array

6.3.1 Optimizing the Inter-element Spacings

To validate our idea of achieving a radiation pattern with low side lobes by adjusting the inter-element spacings, eight radiating elements are employed in designing the nonuniform slot array. Figure 6.12 shows the geometry of the 8-element slot

6.3 Geometry of the Antenna Array

Figure 6.12: Geometry of the 8-element slot array.

array antenna. The slot element is designed in the previous section with the optimized structural dimensions listed as follows: $a = 15mm$, $b = 7mm$, $t = 4.8mm$, $s = 1mm$, $\epsilon_r = 3.5$, $w = 13.6mm$, $l = 6mm$. The transverse slot is located at the centerline of the waveguide and the distance between the slot to the shorted end is $11.6mm$. All the slot elements are excited uniformly with the same amplitude and phase. It is seen that we can use this 8-element slot array to shape the H-plane radiation to meet our low side lobe requirement by adjusting the inter-element spacings, while maintaining the E-plane radiation pattern similar to that of a single slot. In order to assure the symmetry of the radiation patterns, only four different spacings d_1 , d_2 , d_3 and d_4 are used for this 8-element array, as illustrated in Fig. 6.12.

If all the array elements are considered as ideal point sources, desired radiation pattern can be easily obtained by adjusting the inter-element spacings and analyzing the array factor pattern. Figure 6.13 shows an example of the normalized array factor pattern. The inter-element spacings are adjusted to $d_1 = 0.57\lambda_0(17mm)$,

6.3 Geometry of the Antenna Array

Figure 6.13: Normalized array factor pattern of the 8-element nonuniform linear array ($d_1 = 0.57\lambda_0$, $d_2 = 0.57\lambda_0$, $d_3 = 0.7\lambda_0$ and $d_4 = 0.7\lambda_0$, $f = 10$ GHz).

$d_2 = 0.57\lambda_0(17mm)$, $d_3 = 0.7\lambda_0(21mm)$ and $d_4 = 0.7\lambda_0(21mm)$, where the operating frequency is 10 GHz. It is observed that the main-to-side lobe ratio is 16 dB. It is found in our study that the side lobes can be further suppressed if inter-element spacings can be smaller than $0.57\lambda_0$. However, when we consider the actual fabrication of the array, at least $17mm$ should be left because the partially dielectric-loaded rectangular waveguide has a broad side dimension of $15mm$ and the waveguide wall will be about $2mm$. In our study, it is also observed that array factor patterns with low side lobes can be obtained when the spacings between the elements close to the center of the array are small and those for the elements near the two ends are larger. For example, d_1 and d_2 should be small while d_3 and d_4

6.3 Geometry of the Antenna Array

Figure 6.14: Simulated H-plane radiation patterns of the 8-element nonuniform slot array at 10 GHz.

should be relatively large.

Next we apply the above designed inter-element spacings to our partially dielectric-loaded waveguide slot array, which is to be modeled using the full-wave technique. By employing the hybrid method developed in Chapter 3, a full-wave analysis is applied on the 8-element slot array. The H-plane radiation pattern is illustrated as the solid line in Fig. 6.14. It is seen that the main-to-side lobe ratio is only about 14 dB, which is 2 dB worse compared to Fig. 6.13. This is due to the fact that the mutual coupling between the slot elements changes the field distribution along the slots and in turn changes the radiation pattern of the slot

6.3 Geometry of the Antenna Array

array. Therefore, we must modify the inter-element spacings based on our full-wave calculation for the slot array to obtain a better H-plane pattern with lower side lobes. After carefully adjusting all the spacings d_1 to d_4 , a successful design is obtained. The computed H-plane radiation pattern is also plotted as the dashed line in Fig. 6.14 for comparison, where $d_1 = 17mm$, $d_2 = 17mm$, $d_3 = 24mm$ and $d_4 = 26mm$. It can be seen that the main-to-side lobe ratio is 16 dB for this case, which meets our requirement well. It is also observed that the main radiation beam only changes slightly compared to the side lobes. Further studies for the H-plane radiation patterns of the slot array show that the main beam and the first null of this kind of linear slot array with nonuniform spacing do not change much compared to those of a uniform slot array. This is an attractive advantage of this nonuniform slot array because we need not compromise the beamwidth with the side lobe level in designing such slot arrays. Actually, it can be considered that the radiated power for the side lobes is redistributed when the nonuniform inter-element spacings are introduced to the uniform slot array.

In order to further illustrate the effect of mutual coupling on the slot array's radiation patterns, we modified our computer codes, which still perform a full-wave analysis on each element of the array but don't take the mutual coupling between them into consideration. The calculated H-plane radiation patterns are shown in Fig. 6.15. Comparing Fig. 6.15 to Fig. 6.14, it is observed that the radiation pattern changes, especially for the first two side-lobes. The side-lobe level is also 2 dB lower than that with the mutual coupling.

The scattering parameters of the nonuniform slot array are also examined and

6.3 Geometry of the Antenna Array

Figure 6.15: Simulated H-plane radiation patterns of the 8-element nonuniform slot array at 10 GHz (Mutual coupling is not included).

some of them are shown in Fig. 6.16. It can be seen that the return loss of the excitation ports changes only a little compared to single slot element. And the 10 dB return loss bandwidth almost remains unchanged for each port. It is also observed that the larger the distance between two elements, the smaller is the mutual coupling, as we expected. The mutual coupling from Ports 2, 3, 4 and 8 to Port 1, as illustrated in Fig. 6.16(a), is rather small. The strongest coupling from Port 2 to Port 1, which is below -25 dB, occurs because Port 2 is the closest port to Port 1 and the distance between them is 26mm, which is nearly $0.87\lambda_0$. The other ports are located farther from Port 1 and therefore their mutual coupling is weaker. Similar performance is also observed for Port 5, as shown in Fig. 6.16(b).

6.3 Geometry of the Antenna Array

(a)

(b)

Figure 6.16: Simulated return loss and coupling coefficients of (a) Port 1 and (b) Port 5 ($d_1 = 17\text{mm}$, $d_2 = 17\text{mm}$, $d_3 = 24\text{mm}$ and $d_4 = 26\text{mm}$).

6.3 Geometry of the Antenna Array

However, the strongest mutual coupling to Port 5 is above -20 dB, which comes from Port 4. This is due to the fact that the distance between Ports 4 and 5 is only 17mm, which is much smaller than that between Ports 1 and 2. In fact, when we analyze one port, only the coupling from the nearby ports should be included in the calculation. For example, only Ports 3, 4, 6 and 7 need to be taken into consideration when we calculate scattering parameters of Port 5. Other ports having only weak coupling to Port 5 can be neglected. This approximation gives out almost the same scattering parameters as well as the radiation patterns. Furthermore, by doing so, the computation time can be greatly reduced.

6.3.2 Feeding Structure

Now the 8-element waveguide slot array is designed and the inter-element spacings are determined. However, the slot element is fed by the dielectric-loaded rectangular waveguide, which can not be directly connected to standard connectors for testing. Therefore, a transitional structure that transmits electromagnetic waves from the SMA connector to the partially dielectric-loaded rectangular waveguide is

Figure 6.17: Geometry of the SMA to waveguide transition section.

6.3 Geometry of the Antenna Array

Figure 6.18: Simulated scattering parameters of the SMA to waveguide transition (single transition, $t_1 = 5.8mm$ and $t_2 = 4.8mm$).

needed. Such a transitional section connecting a coaxial line to an air-filled rectangular waveguide was described in [70, 111]. However, the design of the transition from a coaxial line to partially dielectric-loaded waveguide has not been reported. Figure 6.17 illustrates the geometry of the SMA to the dielectric-loaded waveguide transitional section. The waveguide has the dimensions of $15mm \times 7mm$. The thickness and relative permittivity of the dielectric slab are $4.8mm$ and 3.5 , respectively. In order to excite the dominant mode in the dielectric-loaded waveguide, the SMA connector is located at the center of the waveguide. The distance between the center conductor of the SMA to the shorted end of the waveguide is t_1 . The length of the metal pin and the teflon shell in the waveguide is t_2 . t_1 and t_2

6.3 Geometry of the Antenna Array

Figure 6.19: Photograph of the fabricated one-to-eight Wilkinson power divider.

mainly determine the center frequency and the level of the return loss, respectively. With the help of HFSS, they are easily adjusted to $t_1 = 5.8mm$ and $t_2 = 4.8mm$. The scattering parameters at the SMA port are shown in Fig. 6.18. It can be seen that the return loss is very good in the whole frequency range of interest.

A one-to-eight power divider is then needed so that the excitation of each slot element has the same amplitude and phase. In order to test each element in the slot array individually, we decided to use a separate power splitting network. Wilkinson power divider [112] is chosen since it has been widely adopted in many applications at microwave frequencies for its attractive features such as wide band,

6.4 Measured Results

lossless with matched output ports and low coupling between the output ports. First, we designed a unit of microstrip-line Wilkinson power divider. Next, we arranged seven units in three stages to build a one-to-eight power divider. The input power is first split into two ways, then four ways, and finally eight ways at the output ports. The one-to-eight Wilkinson power divider is then fabricated on a printed circuit board (PCB, Roger 4003) and its photograph is shown in Fig. 6.19. Exactly the same amplitude and phase of the voltage is obtained for all the output ports due to the symmetry of the circuit.

6.4 Measured Results

Based on the dimensional parameters obtained in the previous section, an 8-element nonuniform slot array is fabricated. The top and the bottom views of the slot array are shown in Fig. 6.20. All the rectangular waveguides are fabricated on one aluminium base via traditional metal machining technology. Our dielectric material is cut into 8 slabs of a thickness $4.8mm$, which are placed at the bottom wall of the waveguide. The cover of the array is made of $1mm$ thick steel with the slots opened on it. The aluminium base and the steel cover are then fixed through metal screws. Eight ports with SMA connectors can be seen from the bottom view, which need to be excited with the same amplitude and phase.

The scattering parameters of each slot element are tested on a vector network analyzer (VNA) HP-8722ES. Full two-port calibration are conducted which enable us to measure a two port network. When two of the ports are connected to the VNA, other ports are terminated by perfectly matched terminations to avoid

6.4 Measured Results

(a)

(b)

Figure 6.20: (a) Top and (b) bottom views of the fabricated nonuniform waveguide slot array.

6.4 Measured Results

(a)

(b)

Figure 6.21: Measured scattering parameters of the slot elements.

6.4 Measured Results

Figure 6.22: Measured return loss of the 8-element slot array.

reflection. Some of the measured scattering parameters are illustrated in Fig. 6.21. It can be observed that the 10 dB return loss bandwidth for each port is from 9.1 GHz to 10.6 GHz, which is about 15 percent of the center frequency 10 GHz. However, there is a small shift for this bandwidth compared to the calculated data plotted in Fig. 6.16. The measured return loss of the excitation port also shows two peaks in the frequency range of interest. These small discrepancies are due to the fact that there were some inaccuracies during the mechanical machining. However, the measured coupling coefficients have similar shapes and levels compared to the theoretical ones shown in Fig. 6.16.

The one-to-eight power divider is then connected to the SMA ports of the slot

6.4 Measured Results

Figure 6.23: Measured radiation patterns of the 8-element nonuniform slot antenna array at 10 GHz.

array through eight RG400 cables, which have a maximum phase difference of 5 degrees between each two of them in the frequency range considered. The return loss of the final array is measured on the same VNA and presented in Fig. 6.22. It can be seen that the 10 dB return loss bandwidth of this slot array is from 9 GHz to 10.7 GHz, which is 17% of the center frequency 10 GHz.

The radiation patterns of the nonuniform slot array are obtained experimentally in a microwave anechoic chamber. The H-plane and E-plane radiation patterns of the antenna array at 10 GHz are shown in Fig. 6.23. It can be seen that the side lobe level of the H-plane radiation pattern is 16 dB lower than the main beam. The

6.4 Measured Results

Table 6.1: Comparison between calculated and measured gain and main-to-side lobe ratio of the 8-element nonuniform slot array.

Frequency (GHz)	Gain (dB)		Main-to-side lobe ratio (dB)	
	Calculated	Measured	Calculated	Measured
9.5	13.6	13.5	15.8	15.7
10	13.9	13.1	16	16.1
10.5	14	13.8	16.1	15.6

H-plane pattern is a little asymmetric at angles far away from the main beam. It is also observed that the E-plane radiation pattern is similar to that of a single slot element, except that the gain is much higher for the slot array. It is noticed that the E-plane pattern has small ripples and reaches the maximum value at about 60 degrees. This is due to the finite ground plane effect, which has been studied in Chapter 4.

The H-plane radiation patterns at 9.5 GHz and 10.5 GHz are presented in Fig. 6.24(a) and (b), respectively. In both cases, the main-to-side lobe ratio is also better than 15 dB. From the Figs. 6.23 and 6.24, it can be observed that the main radiation beam is very stably pointing to the broad side and there is no shift when the operating frequency changes. It is also noted that the main radiation beam of the slot array becomes narrower when the frequency is higher, as we expected.

The theoretical and experimental results for the gain and main-to-side lobe ratio of the 8-element nonuniform slot array are given in Table 6.1. It can be seen from Table 6.1 that the gain is around 13.5 dB and the gain variation is less than 1 dB in the frequency range from 9.5 GHz to 10.5 GHz. The measured gain is

6.4 Measured Results

(a)

(b)

Figure 6.24: Measured H-plane radiation patterns of the 8-element nonuniform slot antenna array at (a) 9.5 GHz and (b) 10.5 GHz.

6.5 Conclusions

a little lower than that calculated by our full-wave method. This is due to the fact that the material loss is not included in our calculation. The main-to-side lobe ratio is better than 15 dB in the frequency range of 1 GHz, which meets our requirement well.

6.5 Conclusions

In this chapter, an 8-element nonuniform slot antenna array in the broad wall of a rectangular waveguides partially loaded with an H-plane dielectric slab operating at X-band has been designed, fabricated and measured. Based on the detailed review on previous research works on antenna arrays, the element of the antenna array has been chosen to be standing-wave transverse slot. Employing the hybrid method proposed in Chapter 3 as the analyzing tool, the structural parameters of the slot element have been determined. Eight elements have been used with nonuniform inter-element spacings to construct a linear slot array. Utilizing the good feature of a compact size for partially dielectric-loaded waveguide, the inter-element spacings of the slot antenna array have been adjusted to obtain low side lobes for its H-plane radiation pattern. It has been demonstrated that full-wave analysis of the slot array including the mutual coupling between the radiating elements is necessary for predicting the radiation pattern accurately.

A feeding network has been designed and fabricated, which provides excitations of equal amplitude and phase for all the eight ports. The measured data has shown that the 10 dB return loss bandwidth is about 17% for the final antenna array. The main-to-side lobe ratio is higher than 15 dB in the frequency range of 1

6.5 Conclusions

GHz. The radiation pattern is very stable and the gain has only 1 dB variation in this frequency range. The described analysis and design procedure can be readily extended to large slot antenna arrays with low side lobes.

Chapter 7

Conclusion and Recommendations

7.1 Conclusion

In the past three and a half years, a thorough study of dielectric-loaded waveguide slot antennas has been conducted. Firstly in Chapter 3, we developed a rigorous full-wave method for analyzing both longitudinal and transverse slots cut in the wall of a rectangular waveguide partially loaded with an H-plane dielectric slab. The slot structure is decomposed into two parts: a dielectric-loaded waveguide T-junction and an open-ended waveguide radiating into half space, which are modeled by mode-matching method and moment method, respectively. Cascading of the two blocks gives the scattering parameters of the waveguide slot structure. The complexity of deriving the Green's functions of a partially dielectric-loaded waveguide is avoided. A new approach has been proposed for computing the integrals in the analysis of an open-ended waveguide, which greatly reduces the computa-

7.1 Conclusion

tion time in the moment method. A large amount of numerical results for the scattering parameters and circuit parameters have been presented and verified by comparing with available measured data or those obtained by HFSS. Excellent agreement has been observed for all the cases considered. The accuracy and efficiency of the proposed hybrid method for modeling dielectric-loaded waveguide slot structures have been demonstrated.

Parametrical studies have been performed and the effect of the dielectric slab's thickness and relative permittivity on the slot antenna's impedance/admittance has been examined. Calculated results have shown that the normalized impedance/admittance can be effectively adjusted by changing the loaded slab's thickness and dielectric constant, in addition to the slot dimensions. It has been illustrated that placing the dielectric slab at the bottom of the waveguide is a better configuration than at the top of the waveguide or fully loading the waveguide with dielectric material. It is also seen that the thickness of the waveguide broad wall can greatly change the normalized impedance of the slot.

The radiation patterns of the partially dielectric-loaded waveguide slot antennas have been investigated in Chapter 4. Combination of the proposed hybrid method and GTD is applied for the slot antennas to account for the finite waveguide wall. Both longitudinal and transverse slots have been analyzed in Chapter 4. Calculated radiation patterns have been presented and compared with measured data or those obtained by HFSS. Excellent agreement has been observed. The hybrid GTD method has been demonstrated to be accurate and efficient for solving waveguide slot antennas with a finite ground plane. Comparing the radia-

7.1 Conclusion

tion patterns of waveguide slot antennas with a finite ground plane to those with an infinite ground plane, it has been demonstrated that the finite ground plane has a small effect on E-plane and H-plane patterns of slot antennas on both air-filled and partially dielectric-loaded waveguides in the forward direction, except small ripples are introduced. It has also been illustrated that the ground plane size can noticeably change the backward radiation level as well as the front-to-back ratio of the slot antennas. Numerical examples have shown that if the edges of the waveguide are asymmetric with respect to the slot, the E-plane pattern will exhibit a small squint while the main lobe of the H-plane pattern has only a slight change.

The proposed hybrid full-wave technique is then employed for characterizing the mutual coupling between rectangular slots cut in the wall of a rectangular waveguide partially loaded with a dielectric slab. In Chapter 5, the mutual coupling between open-ended rectangular waveguides are examined first. It has been shown that the larger the distance between the waveguide openings, the weaker the mutual coupling. Broadside coupling between rectangular waveguides has been demonstrated to be stronger than narrow side one. Coupling between two slots cut on the same partially dielectric-loaded waveguide has also been analyzed. The internal and external mutual coupling has been theoretically taken into consideration in our approach. Circuit model for two slots cut on the wall of a dielectric-loaded waveguide has been introduced and the mutual impedance between the two slots has been extracted. It has been shown that the magnitude of the mutual impedance becomes smaller when the distance between the two slots increases, as we expected. A linear array consisting of three transverse slots cut on the wall of a same partially dielectric-loaded rectangular waveguide has also been studied. The

7.1 Conclusion

calculated scattering parameters have been compared to those obtained by HFSS. Good agreement has been observed for all the cases considered. It has been illustrated that the scattering parameters vary when the distance between the slots changes. When the element spacing is carefully adjusted, both the return loss and the insertion loss can be very small and most of the power fed into the waveguide is radiated to the free space through the slots.

In order to appreciate the advantage of the compact size of dielectric-loaded waveguide slot antenna, an 8-element nonuniform slot array has been designed, fabricated and tested, as described in Chapter 6. Standing-wave transverse slots have been adopted as radiating elements. The structural parameters of the slot element has been determined to obtain a wide return loss bandwidth. After the mutual coupling between slot elements is examined, eight elements have been then used with nonuniform inter-element spacings to construct a slot antenna array. The inter-element spacings has been designed to obtain an H-plane radiation pattern with low side lobes. It has been demonstrated that the full-wave analysis of the slot array including the mutual coupling between radiating elements is necessary for predicting the radiation pattern accurately. Feeding network has also been designed and fabricated, which provide excitations of equal amplitude and phase for all the eight ports. It has been seen from the measured data that the 10 dB return loss bandwidth is about 17% for the finally fabricated antenna array at the operating frequency of 10 GHz. It has also been observed that the main-to-side lobe ratio is higher than 15 dB in the frequency range of 1 GHz. The radiation pattern is very stable and the gain has only 1 dB variation in this frequency range.

7.2 Recommendations for Further Research

In conclusion, dielectric-loaded waveguide slot antennas and arrays are extensively studied in this thesis. With the property of compact size and light weight compared to conventional air-filled waveguide slot antennas, dielectric-loaded waveguide slot antennas are suitable candidates for various applications such as phased arrays and scanning arrays in modern wireless communications systems. A rigorous full-wave method is also presented, which is very accurate and efficient compared to commercial software. The proposed method will be very useful in analyzing and designing dielectric-loaded waveguide slot antennas and arrays.

7.2 Recommendations for Further Research

In order to improve the performance of the nonuniform partially dielectric-loaded slot arrays, waveguide T-junctions can be employed to accomplish the power distribution from one port to eight ports or more. Air-filled waveguide T-junction has been designed [113] by introducing a full-height cylindrical post, where the reflection is greatly improved. Dielectric-loaded waveguide T-junction can be designed in a similar way. By integrating the feeding network into the waveguide slot array, the sizes of the array can be further reduced. A large planar array could also be designed in the future to demonstrate the advantages of compact size and lower side lobe level of nonuniform slot arrays cut on the wall of partially dielectric-loaded waveguides.

Another area where progress can be made is the employing of substrate integrated waveguide (SIW) in the slot array application. In recent years, slot antennas

7.2 Recommendations for Further Research

and arrays on SIW has attracted the attentions of many researchers [114]–[116]. This kind of waveguide slot antenna allows the integration of the feeding network and the slot array on a single substrate. As a result, the size and weight of the waveguide are reduced. And the fabrication of SIW slot antenna arrays can be very easy using the modern PCB manufacturing technology. However, it is noticed that there still lacks detailed theoretical analysis for SIW slot antennas. Therefore, full-wave analysis can be conducted to study SIW slot antennas in more detail. In addition, the impedance/admittance of the transverse/longitudinal slots on SIW is very high due to the fact that the waveguide is fully loaded with dielectric material. Partially dielectric loading can be an attempt to lower the impedance/admittance and therefore gain more freedom in designing SIW slot antenna arrays.

Author's Publications

Journal Papers

1. B. Zheng and Z. Shen, "Effect of a finite ground plane on microstrip-fed cavity-backed slot antennas," *IEEE Trans. Antennas Propagat.*, vol. 53, no. 2, pp. 862–865, Feb. 2005.
2. B. Zheng and Z. Shen, "Analysis of dielectric-loaded waveguide slot antennas by the hybrid mode-matching/moment method," *IEICE Trans. Commun.*, vol. E88-B, no. 8, pp. 3416–3427, Aug. 2005.
3. B. Zheng and Z. Shen, "Effect of a finite ground plane on rectangular waveguide slot antennas' radiation patterns," *Electromagn.*, vol. 26, no. 6, pp. 485–502, 2006.

Submitted and under Review

4. B. Zheng and Z. Shen, "Design of a nonuniform slot array on dielectric-loaded rectangular waveguides," submitted to *Proc. IEE*

Conference Papers

1. B. Zheng and Z. Shen, "Effect of a finite ground plane on circularly polarized microstrip antennas," in *Proc. IEEE Antennas and Propagation Society Int. Symposium*, vol. 2A, pp. 238–241, Washington, July 2005.

7.2 Recommendations for Further Research

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