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A DUAL POLARIZED X-BAND ARRAY

by

Eric Johnson, B.Eng.

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

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

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A DUAL POLARIZED X BAND ARRAY

submitted by

Eric Johnson, B. Eng.

in partial fulfilment of the requirements for
the degree of Master of Engineering

Thesis Supervisor

Chair, Department of Electronics

Carleton University
January, 1992

ABSTRACT

This thesis covers two areas related to the development of a dual polarized X-Band antenna. The first thrust looks at series fed microstrip arrays. Specifically, a 17 element series fed patch array is synthesized using phase only techniques. The calculated and measured patterns are presented and show good correlation.

The second area of research in this work concerns the integration of a second orthogonal feed to the array designed in the first thrust. To this end, a relatively new feed type, aperture coupling to feedlines, was investigated. Again, both measured and calculated data are presented.

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LIST OF SYMBOLS

- W - Width of the microstrip line.
- w_e - Equivalent magnetic wave guide width.
- h - Thickness of the dielectric substrate.
- ϵ_r - Relative dielectric constant.
- $\tan\delta$ - Loss tangent of the dielectric substrate.
- $\tan\delta_e$ - Effective loss tangent.
- ϵ_e - Effective dielectric constant.
- λ_0 - Free space wave length.
- λ_g - Wavelength in a microstrip line.
- c - Speed of light in a vacuum.
- f - Frequency.
- β_m - Propagation constant in microstrip.
- Z - Microstrip characteristic impedance.
- α_c - Conductor loss.
- α_c' - Conductor loss correction for surface roughness.
- α_d - Dielectric loss of microstrip.
- Δ_{rms} - RMS surface roughness of the conducting strip in microstrip.
- δ_s - Skin depth.
- δ_n - Length extension of a feedline into a microstrip patch due to excess currents at patch feed junction.
- G_r - Radiation conductance.
- G_s - Surface wave conductance.

B - Equivalent edge capacitance of open circuit microstrip.
 Δl - Equivalent length extension of open circuit microstrip.
 k - Wavenumber.
 k_x, k_y, k_z - Spectral domain coordinates.
 x, y, z - Cartesian coordinates.
 x_0, y_0, z_0 - Source coordinates.
 θ, ϕ, R - Spherical coordinates.
 E - Electric field.
 H - Magnetic field.
 A - Magnetic vector potential.
 V - Voltage.
 J - Current intensity.
 f_s - Sinusoidal current mode in the space domain.
 F_s - Sinusoidal current mode in the spectral domain.
 f_u - Uniform current in the space domain.
 F_u - Uniform current in the spectral domain.
 f_p - Piecewise sinusoidal current mode in the space domain.
 F_p - Piecewise sinusoidal current mode in the spectral domain.
 G_{EJxx} - Green's function relating an x directed electric current element to an x directed electric field.
 G_{EMxy} - Green's function relating an y directed magnetic current element to an x directed electric field.

G_{HJyx} - Green's function relating an x directed electric current element to an y directed magnetic field.

G_{HMyx} - Green's function relating an y directed magnetic current element to an y directed magnetic field.

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CHAPTER 1

INTRODUCTION

1.0 INTRODUCTION

An electromagnetic wave propagating in free space can be polarized in a circular, elliptical, or linear sense. In all cases the actual polarization can be resolved into a linear combination of two orthogonal polarizations. These two independent polarization channels are available to any radio or radar system and can be advantageously utilized. For example, it is possible to transmit two completely different signals in a radio system on the same carrier without interfering with one another since each of the polarization channels are independent. It is also possible to send the same signal on both polarizations to provide an additional source of system diversity since obstacles and multipath sources have differing responses to different polarizations.

Radars can take advantage of the variation in the returns with polarization to improve the sub-clutter visibility of the return signal since hard targets have a particular polarization signature. Polarimetric Synthetic Aperture Radars

achieve this by alternately transmitting two orthogonal linear polarizations from a single dual polarized antenna. While the transmitted pulses are ideally singly polarized waves, the returns generally have components in both orthogonal channels. By suitably processing both polarizations, the radar can, in principle, be used to synthesize any arbitrary combination of transmit and receive polarizations.

Synthetic Aperture Radars require some sort of moving platform that usually takes the form of an aircraft or orbiting satellite. In the case of an airborne installation, the antenna needs to fulfil environmental as well as system requirements. Antennas mounted external to the fuselage need to have a small aerodynamic cross-section and low weight. The system requires that the antenna have high gain, and a shaped beam in elevation.

Microstrip patch array antennas fulfil these requirements quite well, since they are essentially two dimensional, have only a relatively small mass and can be used to form arrays with particular pattern shapes. In the past, batch to batch variability of the materials and high losses would have precluded the use of microstrip technology for this application. In recent years, new materials, and better manufacturing techniques have alleviated many of these problems. Microstrip patch arrays are also highly repeatable

electrically since the manufacturing of the radiating skins is accomplished using standard print and etch techniques.

One of the most stringent requirements on an antenna when used in a polarimetric radar is on its cross-polarization characteristics. The antenna is required to generate two orthogonal polarizations with as little cross-polarized radiation as possible. In microstrip patch arrays cross-polarized energy can arise from several areas. While direct radiation from feeds and the cross-polarization properties of the patches themselves are problematical, an important limitation of the patches is due to inadequate isolation of the two orthogonal feed channels. Any radiation that couples from one feed channel to the other can potentially be radiated at full strength in a parasitic cross-polarized pattern.

One other limitation is placed on patches when they are used to generate two polarizations. The patches are required to be resonant in both dimensions and as a result, standard series feed techniques which rely on modulation of the patch widths along the array in order to achieve a specific far field pattern are not permissible.

1.1 THESIS OBJECTIVES

The objective of this thesis is to discuss, analyze, and design a dual polarized microstrip patch array. The problem is broken into two main thrust areas.

The first thrust covers the development of a series fed microstrip array. This array will have a shaped far field radiation pattern consistent with its use in a polarimetric airborne imaging radar. In the course of this effort, this work will explore a phase only synthesis technique. The phase only technique is required due to the fact that the microstrip patches are resonant in both dimensions. This reduces the degrees of freedom in the series fed design in the sense that the ability to modify the magnitude of the array excitation is essentially lost. This thesis will also examine a technique for the extraction of the series fed patch circuit parameters.

The second thrust of this thesis is the development of a feed, which can be used in conjunction with the series fed array designed in the first thrust. The purpose of the second feed is to excite the orthogonal polarization on the same elements as the series fed design. This will include a 'full wave' analysis using numerical methods. Specifically, this work required the development of a general purpose moment method analysis for analyzing microstrip patch radiators.

1.2 THESIS OUTLINE

Chapter 2 is a discussion of all of the relevant background in the design of microstrip and rectangular microstrip patch radiators. The transmission line parameters of microstrip lines are defined and derived. The microstrip open circuited line is explored in an effort to understand the radiation mechanism of microstrip patches. With the above material in hand the chapter then moves to a discussion of aperture coupled patches and the design of series fed arrays. The chapter closes with an outline of the phase only optical synthesis technique for line sources.

Chapter 3 discusses the design and test philosophy used in the creation of a series fed shaped beam antenna. The required beam shape is derived and the results of the synthesis are incorporated into this chapter. As a part of the work in this chapter the characterization results for the series fed elements are included. The chapter finishes with measured far field radiation patterns of the array.

Chapter 4 outlines the design process which was followed in the selection of a suitable feed for providing dual polarized operation using the patch geometry from Chapter 3. This chapter will detail the criteria which restrict the design choices for the orthogonal polarization feed. In

particular the sources of cross-polarized radiation in microstrip arrays and some of the difficulties encountered with a standard approach are detailed. Again this chapter finishes with various measurements from the design.

Chapter 5 is a summary of the work in this thesis. As well, it discusses several possible areas of further work.

CHAPTER 2

MICROSTRIP TRANSMISSION LINES AND PATCH RADIATORS

2.0 INTRODUCTION

The purpose of this chapter is to outline the current theoretical techniques used in the synthesis and analysis of microstrip feed lines and patch radiators. Due to the high degree of mathematical complexity involved in the complete full wave analysis of any microstrip structure, both lines and patches, most of the following description will examine the various approximate expressions and analysis techniques available to the designer of microstrip components. In addition, the Method of Moments, one of several powerful numerical techniques available for analyzing microstrip structures has been detailed in Appendix A of this text, with a particular emphasis on its use for extracting the performance parameters of rectangular microstrip patches.

2.1 MICROSTRIP TRANSMISSION LINES

Microstrip is one of a class of planar transmission structures. It is a two conductor transmission system

Grounded Dielectric Slab

Figure 2.1-1 Illustration of a microstrip line.

a) Rectangular Waveguide

b) Stripline in dielectric.

c) Coaxial Line

Figure 2.1-2 Uniform transmission line structures.

consisting of a narrow conducting strip printed on a grounded dielectric slab as shown in Figure 2.1-1. The abrupt dielectric boundary at the air-substrate interface makes microstrip an inhomogeneous transmission media. Unlike waveguide or stripline, shown in Figures 2.1-2 a) and b), which are homogeneous structures microstrip is incapable of propagating energy in a single well defined mode except in the degenerate case where the substrate is air. Gupta (as cited in [1]) used Maxwell's equations to show that all six cartesian components of the fields were required to fully describe the propagating mode. This result makes any complete analytical study of microstrip lines virtually impossible. Figure 2.1-3 is an approximate representation of the fields.

All is not lost, however, due to one important feature of the dominant mode. The bulk of the transmitted energy, as one might expect, is contained in the region between the printed conductor and the ground plane. The field structure in this region, as shown in Figure 2.1-3, approximates the field structure that would exist in a homogenous media which is capable of transmitting energy with a TEM field structure. As a result microstrip is often referred to as a "quasi-TEM" transmission structure.

The implications of the TEM label are very important. TEM lines have a unique characteristic impedance, a dispersionless

a) Approximate field structure of Stripline

b) Approximate fields of a microstrip line.

Figure 2.1-3 Illustrations of the approximate fields of a) Triplate or Stripline and b) Microstrip.

relationship between propagation constant and frequency, and are generally well behaved in an electromagnetic sense. Like any approximation the quasi-TEM model does have some limitations. In the case of microstrip, the model works very well up to several GHz in frequency, at which point some appreciable dispersion does begin to occur. The model can easily be extended to account for this dispersion and, having done so the parameters are again reasonably accurate over a smaller bandwidth.

The remainder of this section will outline the computation of the propagation parameters for microstrip. In most cases the expressions are empirical in nature since, as noted previously, any full analysis is highly difficult. It is felt that this is representative of the technology to date and that deeper analysis is outside of the scope of this thesis.

2.1.1 TRANSMISSION PARAMETERS

Using the quasi-TEM model we can define the propagation parameters of the line.

2.1.1.1 EFFECTIVE DIELECTRIC CONSTANT

The concept of the effective dielectric constant is the most critical parameter of a microstrip line in the quasi-TEM

model. It represents the relative amounts of power in the air and the dielectric substrate. It is a variable dependent on the substrate thickness, the width of the conducting strip, and the actual relative dielectric constant of the substrate material. While there is no exact analytic expression relating the effective dielectric constant to the geometry of a microstrip line this relationship has been investigated experimentally and through the use of approximate models. Wheeler [2] derived a set of design data by conformally mapping the geometry of the line into a parallel plate configuration. Depending on the geometry of the line the gap between the plates was filled to a lesser or greater extent. From this he derived an equivalent filling factor, q , which then related the effective dielectric constant, ϵ_e , to the relative dielectric constant, ϵ_r , according to [2]

$$\epsilon_e = 1 + q(\epsilon_r + 1) \quad 0.5 < q < 1.0 \quad (2.1)$$

When the line is very narrow q approaches 0.5 and the effective value is equivalent to the arithmetic mean of the air and substrate relative dielectric constants. As the line grows in width the value of q approaches unity and the effective constant approaches the relative dielectric constant of the substrate indicating that the bulk of the energy is now contained in the region between the strip and the substrate. An empirical expression for ϵ_e was derived by Schneider [3];

$$\epsilon_e = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left(1 + 12 \frac{h}{w} \right)^{-0.5} \quad (2.2)$$

2.1.1.2 GUIDE WAVELENGTH AND PROPAGATION CONSTANT

Once the effective dielectric constant has been established the guide wavelength and the propagation constant are then simply defined by

$$\lambda_g = \frac{c}{f \sqrt{\epsilon_e}} \quad (2.3)$$

$$\beta = \frac{2\pi}{\lambda_g} \quad (2.4)$$

2.1.1.3 MICROSTRIP CHARACTERISTIC IMPEDANCE

The remaining parameter that defines a TEM transmission line is the characteristic impedance of the microstrip line. In its simplest form the microstrip impedance Z_m is defined by the phase velocity and the capacitance of a similar line where the substrate has been replaced by air. Several standard techniques such as conformal mapping can be applied to the geometry to calculate this parameter. Using a conformal mapping approach, Wheeler [2] gave

for $W/h > 1$

$$Z_n = \frac{377}{\sqrt{\epsilon_e} \left[\frac{W}{h} + 1.393 + 0.667 \ln \left(\frac{W}{h} + 1.444 \right) \right]} \Omega \quad (2.5)$$

for $W/h < 1$

$$Z_n = \frac{60}{\sqrt{\epsilon_e}} \ln \left(8 \frac{h}{W} + 0.25 \frac{W}{h} \right) \Omega \quad (2.6)$$

2.1.1.4 DISPERSION IN MICROSTRIP

While the above expressions are generally applicable over most of the useful frequency range of microstrip it is necessary to account for the dispersion characteristics of the microstrip lines. As the frequency is increased the transverse currents on the microstrip line increase with the result that the fields of the dominant mode become increasingly confined below the strip. This effect causes a change in the observed effective dielectric constant of the line. This has been examined and the expression given by Getsinger [4] for this effect is

$$\epsilon_e(f) = \epsilon_r - \frac{\epsilon_r - \epsilon_e(0)}{1 + G(f/f_p)^2} \quad (2.7)$$

where

$$f_p = \frac{Z_n}{2\mu_0 h} \quad (2.8)$$

and G is a purely empirical parameter. The expression

given by Getsinger [4] for alumina substrates was

$$G = \frac{\pi^2}{12} \frac{(\epsilon_r - 1)}{\epsilon_e(0)} \sqrt{\frac{2\pi Z_0}{\eta_0}} \quad (2.9)$$

2.1.1.5 OTHER NON-IDEALITIES

There are two other factors which have an effect on microstrip lines. These are finite cover height in the case of shielded microstrip and the finite thickness of the conductor metallization of the printed strip. In both cases these effects add capacitance which in turn increases the effective dielectric constant of the line. By minimizing the thickness and maximizing the cover height these effects can be reduced to negligible levels.

2.1.2 LOSSES IN MICROSTRIP

Losses in microstrip are due to several phenomenon. These include conductor and dielectric losses, power lost to surface wave generation, and radiation losses. Conductor, dielectric, and radiation losses occur at every point in a microstrip circuit. On a uniform line, radiation is generally negligible, while dielectric and conductor losses are not.

2.1.2.1 CONDUCTOR LOSSES

Hammerstad and Bekkadal (as cited in [5]) gave an expression for the conductor loss which accounts for the skin effect and finite conductivity for a copper line.

$$\alpha_c = 0.072 \frac{\sqrt{f}}{w Z_0} \lambda_g \quad \text{dB/(guide wavelength)} \quad (2.10)$$

In addition to this component of the conductor loss an additional multiplicative factor which accounts for the surface roughness of the metallization must be included. The empirical expression derived by Hammerstad and Bekkadal is

$$\alpha_c' = \alpha_c \left[1 + \frac{2}{\pi} \arctan \left\{ 1.4 \left(\frac{\Delta_{rms}}{\delta_s} \right)^2 \right\} \right] \quad (2.11)$$

2.1.2.2 DIELECTRIC LOSSES

Dielectric losses are due to the non-zero conductivity of the substrate material. An expression for this part of the loss was also derived by Hammerstad and Bekkadal (in [5])

$$\alpha_d = 27.3 \frac{\epsilon_r (\epsilon_e - 1)}{\epsilon_e (\epsilon_r - 1)} \tan \delta_e \quad \text{dB/(guide wavelength)} \quad (2.12)$$

Note that the effective dielectric loss tangent has been

used here. This reflects the fact that not all of the fields are confined to the substrate. The effective loss tangent is given by (in [5])

$$\tan\delta_e = \left[\frac{1-\epsilon_r}{1-\epsilon_r} \right] \tan\delta \quad (2.13)$$

2.2 MICROSTRIP OPEN CIRCUIT

The microstrip open end is the simplest configuration which radiates energy in a manner similar to a patch radiator. As an introduction to microstrip patches, the radiation mechanism and equivalent circuit of an open ended microstrip line will be discussed.

In the case of an ideal open circuit, power incident on the plane of the open end would be fully reflected. It is very difficult to construct an ideal open circuit in any high frequency wave guiding structure and the nature of the open end is actually quite complex. The difficulty lies in the fact that an open end is just that, open. Open structures always have associated fringing fields which in turn lead to energy being radiated from the open end. For microstrip this is further complicated by the presence of an abrupt dielectric discontinuity in the terminal region with the result that some of the incident power is lost to the excitation of dielectric surface wave modes. The final contribution of the fringing

fields at the microstrip open end is that the effective location of the terminal plane, as determined by the phase of the returned signal, is further away from the source than the physical terminal plane is.

Fig 2.2-1 illustrates the model of the open end which is extensively used in the literature. In addition to a section of uniform transmission line the end effects are incorporated in the form of three lumped circuit elements. G_r is the radiation conductance, G_s is the surface wave conductance, and B is the equivalent end capacitance to account for the additional phase of the reflected circuit. While this model mimics reality fairly well the limitation has been in finding adequate expressions for determining the circuit parameters.

2.2.1 EQUIVALENT END CAPACITANCE

The equivalent end capacitance of an open ended microstrip is due to the presence of fringing fields at the end of the line. Expressions for the excess capacitance at the open end are available from several sources [6] [7] [8]. Experimentally, the expressions are derived from measurements of the return phase. The equivalent capacitance of the end is then determined by

a) Physical Open Circuit

b) Equivalent quasi-static transmission line model.

Figure 2.2-1 Microstrip physical open circuit and its electrical equivalent.

a) Model Including End Capacitance

b) Open end replaced by short length of line Δl

Figure 2.2.1-1 Effective length extension of microstrip open circuit.

$$B = \tan(\beta \Delta \ell) \quad (2.14)$$

Hammerstad and Bekkadal (as cited in [5]) derived an empirical expression for the equivalent extension of the open end which is given as

$$\Delta \ell = 0.412h \left(\frac{\epsilon_r + 0.300}{\epsilon_r - 0.258} \right) \cdot \left(\frac{w/h + 0.262}{w/h + 0.813} \right) \quad (2.15)$$

2.2.2 RADIATION CONDUCTANCE

There are two general approaches for determining the radiation conductance in the literature. The radiation conductance is related to the amount of power radiated by an open circuit which has a unity voltage wave incident upon it. Lewin used a model based on the approximate currents in the region of the open end. Several other approaches approximate the fields at the open end using an equivalent aperture.

2.2.2.1 LEWIN'S CURRENT MODEL

One of the early analyses of the open end microstrip line was that of Lewin [9]. In that analysis the current of the microstrip line was approximated to be that of a quasi-TEM wave confined below the strip, and it was assumed that the wave was completely reflected at the open end. It was further assumed that the conduction current in the strip flows only

along the length of the strip.

The analysis follows by breaking the current into two components, the physical conduction current and a fictitious current which is equivalent to the dielectric polarization current of the substrate. Using the above assumptions and Figure 2.2.2.1-1, the currents can be expressed vectorially as [9]

Conduction Current

$$J_z = e^{-j\beta z} - e^{+j\beta z} \quad (2.16)$$

Dielectric Displacement Current due to E_y

$$J_{py} = j(\epsilon_e - 1)\omega\epsilon_0 E_y \quad (2.17)$$

The vector sum of these currents is then related to an effective Hertzian dipole, π , by [9]

$$\pi = \frac{-j30}{k_0} \iiint_V \frac{e^{-jkr}}{r} J \, dv \quad (2.18)$$

where V is the volume under the conducting strip. The resulting radiation conductance is [9]

$$G_r = 60 \left(\frac{\pi h}{Z_0 \lambda} \right)^2 F_1(\epsilon_e) \quad (2.19)$$

where

$$F_1 = \left[\frac{\epsilon_e + 1}{\epsilon_e} - \frac{(\epsilon_e + 1)^2}{2\epsilon_e \sqrt{\epsilon_e}} \ln \left(\frac{\sqrt{\epsilon_e} + 1}{\sqrt{\epsilon_e} - 1} \right) \right] \quad (2.20)$$

a) Polarization Current as per Lewin's Model

b) Conduction Current as per Lewin's Model

Figure 2.2.2.1-1 Currents of Lewin's model.

2.2.2.2 APERTURE MODELS

An alternative approach for analyzing an open end microstrip is the aperture approach. In this case the aperture is assumed to be between the feed line and its image in the ground plane, coincident with the end of the line. This is illustrated in a) of Figure 2.2.2.2-1. From the work of Sobol [10], an expression for the radiation conductance using this approach was given as

$$G_r = \frac{\sqrt{\epsilon_e}}{240\pi^2} F_2 \left(\sqrt{\epsilon_e} \frac{2\pi w}{\lambda_0} \right) \quad (2.21)$$

where

$$F_2(x) = xS_i(x) - 2\sin^2\left(\frac{x}{2}\right) - 1 + \frac{\sin x}{x} \quad (2.22)$$

Another approach is to assume that an aperture exists on the air-substrate interface as shown in b) of Figure 2.2.2.2-1. This was the approach taken by James, Hall, and Wood [5] and the resulting expression is given as

$$G_r = \frac{1}{120\pi^2} F_2 \left(\frac{2\pi w}{\lambda_0} \right), \quad (2.23)$$

where F_2 is the same expression given with Sobol's result. One of the difficulties with using the above expressions was that they were difficult to verify experimentally. This was addressed by James, Wood, and Hall. [5] Their measured results are shown in Figure 2.2.2.2-2 along

a) Sobol's choice of aperture location.
Between the strip and the groundplane.

b) Aperture location chosen by James et al.
In the same plane as the conducting strip.

Figure 2.2.2.2-1 a) Sobol's Aperture Model

b) Aperture location in model of James et al.

Comparison of G_r values obtained by various estimations

$w = 2.82$ mm, $h = 1.58$ mm, $\epsilon_r = 2.32$

a Lewin (1960)

b Sobol (1971)

c eqn. 3.10 using w_e instead of w

d Surface field analysis eqn. 3.12 using w_e instead of w

e Kompa (1976)

Experimental $\bullet$ Wood *et al.* (1978)

Figure 2.2.2.2-2 Comparison of various models of the microstrip open circuit to measurements. Equations are referred to James *et al* [5]. Eqn. (3.10) is Sobol's Model.

with the results of various models. Note that by using w_e to account for the fringing fields in the aperture methods the agreement is quite good. w_e is the equivalent magnetic wall waveguide, see Figure 2.2.2.2-3 with dielectric ϵ_e with the same propagation constant and characteristic impedance [5]:

$$w_e = \frac{120\pi h}{Z_m(\epsilon_e)^{1/2}} \quad (2.24)$$

2.2.3 SURFACE WAVE CONDUCTANCE

The surface wave conductance is a measure of the percentage of the incident energy at the open end of the microstrip line that is lost due to the excitation of a substrate surface wave. For the most part this component is small for thin substrates and low dielectric constants. As a general rule the surface wave component may be ignored where [11]

$$h < 0.09\lambda_0 \quad (2.25)$$

2.3 MICROSTRIP PATCH RADIATOR

By making the length of a wide line slightly less than one half of a substrate wavelength the result is a rectangular microstrip patch resonator as shown in Figure 2.3-1.

Similar to the open ended strip a standard rectangular

a) Approximate fields of a microstrip line of width w .

b) Equivalent parallel plate magnetic wall waveguide.

Figure 2.2.2.2-3 Transformation of a microstrip line to an equivalent parallel plate waveguide.

Figure 2.3-1 Microstrip patch and approximate fields.

microstrip patch can be thought of as radiating at its boundaries due the presence of fringing fields at the end of the metallization. Under most of the patch the fields are tightly confined between the ground plane and the resonant length conductor. At the edges the fields form a fringe region where time varying electromagnetic fields exist thus leading to radiation of energy into space. The other mechanism at work at the patch edge is surface wave excitation. For thin patches (i.e $h \ll \lambda$) the dominant loss is due to radiation since the lowest order surface wave mode, TM_0 , is only loosely coupled.

2.3.1 EDGE ADMITTANCE

As in the case of the open circuit microstrip line the edge admittance of a microstrip patch consists of three main components. These are the edge capacitance due to the presence of fringe fields at the edge of the patch, the radiation conductance which represents the amount of power lost to free space and, the surface wave conductance.

To evaluate the radiation conductance experimental evidence [5] suggests that the expression of Sobol gives good agreement with measured results where the width w is the effective width of the equivalent parallel plate guide w_e . Thus G_r is given by

Figure 2.3.1-1 Physical patch with equivalent length extensions and Sobol's apertures. Also shown is the equivalent electrical circuit.

$$G_r = \frac{\sqrt{\epsilon_e}}{240\pi^2} F_2 \left(\sqrt{\epsilon_e} \frac{2\pi w_e}{\lambda_0} \right) \quad (2.26)$$

where

$$F_2(x) = x S_1(x) - 2 \sin^2 \left(\frac{x}{2} \right) - 1 + \frac{\sin x}{x} \quad (2.27)$$

The surface wave conductance is a representation of the amount of power which is lost to the excitation of substrate surface waves. For thin patches this is not generally a problem and it will not be discussed further here.

The edge capacitance of the patch can be represented by an equivalent length extension of the patch edge. Again, this was given by Hammerstad and Bekkadal (as cited in [5])

$$\Delta l = 0.412h \left(\frac{\epsilon_e + 0.300}{\epsilon_e - 0.258} \right) \cdot \left(\frac{w/h + 0.262}{w/h + 0.813} \right) \quad (2.28)$$

With all of these parameters calculated the equivalent network of Figure 2.3.1-1 can be constructed. The input impedance at the edge of the patch is then the combination of the two edge conductances connected by a resonant transmission line.

$$Z_{in}(x) = \frac{1}{2 G_r} \quad (2.29)$$

2.3.2 RESONANT FREQUENCY

A cavity in isolation will become resonant when one of

its dimensions becomes an integer number of half wavelengths long. In the case of the microstrip patch the first resonance occurs when frequency is such that the longest physical dimension of the patch is slightly less than one half wavelength in the substrate. While physically the patch is smaller than resonant size the electrical length of the patch is correct due to the fringe fields at the edges of the patch. Thus resonance is determined by

$$f_r = \frac{c}{2(L+2\Delta\ell)} \quad (2.30)$$

2.3.3 RADIATED FIELDS

An approximate analysis of the radiated fields from a thin microstrip patch ($h \ll \lambda_0$) has been discussed in [12]. In that analysis the patch is assumed to be resonant with the field structure of a magnetic wall waveguide. (i.e. the dual of the transverse electric mode of rectangular waveguide.). The fields can be written [12]:

$$E_z = \frac{V_0}{h} \cos\left(\frac{m\pi x}{a}\right) \cos\left(\frac{n\pi y}{b}\right) \quad (2.31)$$

$$H_x = -\frac{j\omega\epsilon_0}{k_0^2} \frac{n\pi}{b} \frac{V_0}{h} \cos\left(\frac{m\pi x}{a}\right) \sin\left(\frac{n\pi y}{b}\right) \quad (2.32)$$

$$H_y = \frac{j\omega\epsilon_0}{k_0^2} \frac{m\pi}{a} \frac{V_0}{h} \sin\left(\frac{m\pi x}{a}\right) \cos\left(\frac{n\pi y}{b}\right) \quad , \quad (2.33)$$

where V_0 is the peak voltage.

Figure 2.3.3-1 Coordinate system of a rectangular patch [12].

Patch is in the x - y plane.

The lowest order resonance of the patch is determined when the length of the resonant cavity is equal to one half wavelength. At resonance, the voltages at the patch edges are then [12], (see Figure 2.3-1)

$$V(y) = V_0 \cos\left(\frac{n\pi y}{b}\right) \quad x=0; 0 \leq y \leq b \quad (2.34)$$

$$V(y) = V_0 \cos\left(\frac{n\pi y}{b}\right) \cos(m\pi) \quad x=a; 0 \leq y \leq b \quad (2.35)$$

$$V(x) = V_0 \cos\left(\frac{m\pi x}{a}\right) \quad y=0; 0 \leq x \leq a \quad (2.36)$$

$$V(x) = V_0 \cos\left(\frac{m\pi x}{a}\right) \cos(n\pi) \quad y=b; 0 \leq x \leq a \quad (2.37)$$

The electric fields at the edge of the patch can then be related to an equivalent magnetic current around the periphery of the patch using the equivalence principle. Using this assumption and the above voltage definitions yields, in spherical coordinates, the radiated field components of a patch excited in its lowest order mode [12].

$$E_\theta = -j \frac{\exp(-jk_0 R)}{2\lambda_0 R} 8V_0 \exp\left[j\frac{k_0 a}{2} \cos\phi \sin\theta\right] \exp\left[j\frac{k_0 b}{2} \sin\phi \sin\theta\right] \\ \sin\left[\frac{k_0 a}{2} \cos\phi \sin\theta\right] \cos\left[\frac{k_0 b}{2} \sin\phi \sin\theta\right] k_0 \sin\phi \cos\phi \sin\theta \\ \left\{ \frac{1}{k_0^2 \cos^2\phi \sin^2\theta} - \frac{1}{k_0^2 \cos^2\phi \sin^2\theta - \pi^2/b^2} \right\} \quad (2.38)$$

E-Plane ($\phi = 90$) Power Pattern

Figure 2.3.3-2 E plane power pattern of a square microstrip patch.

Figure 2.3.3-3 H plane power pattern of a square microstrip patch.

$$\begin{aligned}
E_{\phi} = & -j \frac{\exp(-jk_0 R)}{2\lambda_0 R} 8V_c \exp\left(j \frac{k_0 a}{2} \cos\phi \sin\theta\right) \exp\left(j \frac{k_0 b}{2} \sin\phi \sin\theta\right) \\
& \sin\left(\frac{k_0 a}{2} \cos\phi \sin\theta\right) \cos\left(\frac{k_0 b}{2} \sin\phi \sin\theta\right) k_0 \sin\theta \cos\theta \\
& \left\{ \frac{1}{k_0^2 \sin^2\theta} - \frac{\sin^2\phi}{k_0^2 \sin^2\phi \sin^2\theta - \pi^2/b^2} \right\}
\end{aligned}
\tag{2.39}$$

Figures 2.3.3-2 and 2.3.3-3 show plots of (2.38) and (2.39) in the principal planes of the radiation (i.e. $\phi=0^\circ$ and $\phi=90^\circ$). It can be easily shown that one or the other of the two components is zero on the principal axes. This is an important result since the two components are orthogonal and the result that one component is very small near the axes of the antenna implies that the patch is capable of generating a wave with a high degree of polarization purity if the main beam is formed near one of the principle planes. It has been shown [13] that the cross-polarized component reaches a peak for look angles near the diagonals of the patch and the level of the orthogonal radiation is largely determined by the dielectric constant of the patch substrate.

2.4 SERIES FED MICROSTRIP PATCHES

There are many ways that a patch may be fed. They can be fed at a single point, as in the case of a coaxial probe inserted through the groundplane or a coplanar microstrip feed. They can also be electromagnetically coupled to a microstrip or waveguide for example. Figure 2.4-1 illustrates

a) Coplanar microstrip feed.

b) Coaxial Probe Feed

c) Coplanar Electromagnetically Coupled Feed

d) Immersed Electromagnetically Coupled Feed

e) Aperture Coupled Feed

Figure 2.4-1 Single point, single polarization feeds for patches.

a) Coplanar microstrip.

b) Coaxial Probe Feed

c) Coplanar Electromagnetically Coupled Feed

Figure 2.4-2 Dual orthogonal feeds for patches.

a) Coplanar microstrip feed.

b) Coaxial Probe Feed

c) Immersed Electromagnetically Coupled Feed

d) Aperture Coupled Feed

Slot in
Groundplane

Figure 2.4-3 Series fed single polarization feed configurations.

several common single point feeds.

Adding a second feed allows for many interesting features. If the second feed is placed, as in Figure 2.4-2, on the orthogonal side of the patch and excited in phase quadrature at an equal power level the patch will radiate a circular polarized wave. One other possibility is to connect the feed to the opposing side of the patch, as indicated in Figure 2.4-3. These last configurations represents the basic building blocks of series fed arrays.

2.4.1 CIRCUIT MODEL FOR A SERIES FED PATCH

Jones, Chow, and Seeto, [14] devised a transmission line model for the series fed patch. Referring to Figure 2.4.1-1 the model is made up of ten components. At this point each of these elements has been defined in this chapter with the exception of the term δ_n . This term represents the effective extension of the narrow feedline as a result of the change in the current density at the edge of the patch. Additionally, the presence of the feedline at the edge of the patch effectively reduces, to first order, the radiation conductance and the fringe capacitance in direct relation to the percentage of the aperture that the feedline covers.

While some of the parameters of the model can be

a) Physical Patch Element.

b) Transmission Line Sections.

c) Circuit Model of A Series Fed Patch.

Figure 2.4.1-1 Series fed patch model from Jones et al [14].

calculated with reasonable accuracy, proper characterization of this configuration requires experimental testing.

In the experimental characterization of these components the parameters G_r , δ_n , and $\Delta\ell$ can be extracted using the following expressions. The response of the patch and its feed are compared to a uniform line of the same width.

From the measured values for the return loss, insertion loss, and insertion phase of the overall construct versus the through line the parameters can be extracted. The quantity $\Delta\ell$ is derived from the measured resonant frequency by solving for $\Delta\ell$ in equation (2.30) thus

$$\Delta\ell = \frac{C}{4\sqrt{\epsilon_e}f_r} - \frac{L}{2} \quad (2.40)$$

The effective extension of the feed line due to the excess inductance of the feed to patch transition is determined from the measured insertion phase. Assuming that the patch is 180° the measured insertion phase, ϕ_{ins} , is measured relative to a uniform line and the phase over and above the amount attributable to the patch, ϕ_{exc} is a result of this effect. Thus from [14] the effective extension δ_n is given by

$$\delta_n = \frac{\lambda_n}{2\pi} \phi_{exc} \quad (2.41)$$

The final parameter to account for in this model is the radiation conductance. From the equivalent model of the circuit and simple circuit theory the conductance can be related to the measured insertion loss IL by

$$2G_r = \frac{1}{Z_0} \frac{1 - 10^{-IL/10}}{10^{-IL/10}} \quad (2.42)$$

2.4.2 TRANSMISSION MATRICES FOR THE SERIES FED COMPONENTS

To analyze the total series fed structure it is necessary to derive a transmission model of the patch. From Figure 2.4.1-1 the models required are for a uniform section of transmission line, an impedance step, and for a shunt conductance connected across a impedance junction. The models can then be inserted as applicable into a computer model and the amplitude and phase of the distribution across the aperture can be evaluated and optimized as required.

2.5 APERTURE COUPLED PATCHES

The section will detail the analytical technique used in determining the parameters associated with aperture coupled microstrip patches. The concept of using small coupling apertures is not a new one but their specific use for coupling

Figure 2.4.2-1 Wave transmission matrix for a uniform line.

$$\begin{bmatrix} v_1^+ \\ v_1^- \end{bmatrix} = \begin{bmatrix} \frac{Z_2 + Z_1}{2 Z_2} & \frac{Z_2 - Z_1}{2 Z_2} \\ \frac{Z_2 - Z_1}{2 Z_2} & \frac{Z_2 + Z_1}{2 Z_2} \end{bmatrix} \begin{bmatrix} v_2^+ \\ v_2^- \end{bmatrix}$$

Figure 2.4.2-2 Wave transmission matrix for a step in impedance.

$$\begin{bmatrix} V_1^+ \\ V_1^- \end{bmatrix} = \begin{bmatrix} \left(1 + \frac{G_r}{2} Z_1\right) & \frac{G_r}{2} Z_1 \\ -\frac{G_r}{2} Z_1 & \left(1 - \frac{G_r}{2} Z_1\right) \end{bmatrix} \begin{bmatrix} V_2^+ \\ V_2^- \end{bmatrix}$$

Figure 2.4.2-3 Wave transmission matrix for a shunt conductance.

a patch to a feed line below the patches groundplane was first proposed by Pozar [15]. Following his initial treatment Sullivan and Schaubert [16] produced a paper on the subject, where they used a thin rectangular aperture and proposed a 'full wave' moment method analysis for the structure. Pozar [17] countered with a paper using the concept of reciprocity to simplify the analysis at the expense of some precision in his calculated result. The following is a treatment of Pozar's work [17]. The Spectral Domain Green's functions and the general algorithmic details of the Method of Moments have been included as Appendix A.

2.5.1 FORMULATION OF THE APERTURE COUPLED PATCH PROBLEM

Figure 2.5.1-1 illustrates the geometry of the problem. The full analysis of the patch-aperture-feedline combination can be broken into four smaller problems: 1) patch self reactance; 2) aperture self reactance; 3) interaction between the patch and the aperture, and; 4) the perturbation of the fields on the feedline which is represented as a voltage discontinuity vector Δv . The aperture is modelled by closing the physical hole and replacing it with two equal but opposite magnetic currents over the aperture region.

Figure 2.5.1-1 Exploded view of an aperture coupled patch.

2.5.1.1 PATCH REACTION TERMS

The patch is represented by equivalent current modes over the region of metallization similar to Figure 2.5.1.1-1. With a centered aperture and a simple rectangular shape these currents can be assumed to vary sinusoidally in the x direction only [17]. The physical existence of a current edge effect on the patches is ignored and has been shown to have little effect on the end results [18]. A final point, due to the symmetry of the aperture under the patch only those current modes which are symmetrical about the centre of the patch need to be used [17]. The surface currents on the patch in its first resonance can then be assumed to be made up of a superposition of several entire domain functions which are uniform in y , and vary in x [17].

$$J_{sx} = \sum_{n=1}^N I_n f_s^n(x) f_u(y) \quad (2.43)$$

The self reaction terms for the patch are then [17]

$$Z_{nn} = - \int_{s_p} \int_{s_p} f_s^n(x) f_s^m(x_0) f_u(y) f_u(y_0) G_{EJxx}(x, y, h | x_0, y_0, h) ds ds_0 \quad (2.44)$$

While this expression does not appear very complicated it entails a great deal of other work. To date no closed form expression has been formulated which accurately describes the

a) Patch Current Modes X-Directed

b) Aperture Magnetic Current Mode Y-Directed

Figure 2.5.1.1-1 Breakdown of the patch and aperture into equivalent electric and magnetic currents respectively.

Spatial Green's Function. Instead use is made of the Fourier or Spectral Domain and $G_{EJxx}(x,y)$ is related to its Fourier representation, $G_{EJxx}(k_x, k_y)$, by [17]

$$G_{EJxx}(x,y) = \frac{1}{4\pi^2} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} \tilde{G}_{EJxx}(k_x, k_y) e^{j(k_x(x-x_0) + k_y(y-y_0))} dk_x dk_y \quad (2.45)$$

The patch reactance terms Z_{mn} are then fully represented in the Fourier domain by [17]

$$Z_{mn} = \frac{1}{4\pi^2} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} F_s^n(k_x) F_s^m(k_x) F_u^2(k_y) \tilde{G}_{EJxx}(k_x, k_y) dk_x dk_y \quad (2.46)$$

2.5.1.2 APERTURE SELF REACTION TERMS

The aperture is calculated in a similar manner. The difference is that we are now referring to magnetic currents so the result is an admittance. The self admittance of the aperture is then given as [17]

$$Y_{e_{mn}} = \frac{1}{4\pi^2} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} F_p^n(k_y) F_p^m(k_y) F_u^2(k_x) \tilde{G}_{HMyY}(k_x, k_y) \cos(y_m - y_n) dk_x dk_y \quad (2.47)$$

As indicated this results in a matrix expression for the aperture admittance. In practice the slot is often quite short electrically and this can be reduced to a single mode in the slot ($m=n=1$). The modes used in the slot are piecewise

sinusoids [17] which vary from triangular to sinusoidal currents as the slot grows in length.

2.5.1.3 INTERACTION BETWEEN THE SLOT AND PATCH

The interaction between the slot and the patch is defined by the excitation terms, V_n . These terms represent the reaction of the fields due to the magnetic current in the slot and the current modes imposed on the patch. The terms, V_n , are given as, assuming a single mode in the aperture [17]

$$V_n = \frac{1}{4\pi^2} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} F_s^n(k_x) F_p(k_y) F_u(k_y) F_u(k_x) \tilde{G}_{EMxy}(k_x, k_y) dk_x dk_y \quad (2.48)$$

2.5.1.4 FEED DISCONTINUITY VECTOR Δv

The discontinuity vector is a measure of the interaction of the feed line with the slot in the groundplane. It is given as the inner product of the slot electric field and the feedlines' magnetic field. The expression for this term is given as [17]

$$\Delta v = \frac{1}{2\pi\sqrt{Z_0}} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} F_p(k_y) F_u(k_y) \tilde{G}_{HJyx}(j_x = -\beta, k_y) dk_y \quad (2.49)$$

Figure 2.5.2-1 Equivalent electrical model of an aperture coupled patch and feedline.

2.5.2 CIRCUIT PARAMETERS

From the V and Z terms we then calculate the interaction admittance between the patch and the slot and apply it to the slot. This term Y_a is given as

$$Y_a = [V]^t [Z]^{-1} [V] \quad (2.50)$$

The series impedance then seen on the microstrip feed line of impedance Z_m is determined from

$$Z = Z_m \frac{\Delta v^2}{Y_e + Y_a} \quad (2.51)$$

The impedance in general will not be pure real so the configuration is stub tuned to arrive at the proper input match. The full circuit is shown in Figure 2.5.2-1.

2.6 OPTICAL SYNTHESIS

One of the most basic ways in which to view an array of radiating elements is to visualize the array as being made up of a fixed number of illumination sources which illuminate a particular portion of the far field. The synthesis results in a design which has a monotonic phase distribution in the near field. A monotonic phase distribution is desirable in that the resulting far field pattern will be less sensitive to changes

in frequency since the relative phasing of progressive elements remains fairly constant with changes in frequency [19]. By the same token systematic errors, such as repetitive manufacturing tolerances and temperature effects are similarly minimized.

Following the analysis of [20] the theory of the optical synthesis technique follows. In this case the synthesis is for the one dimensional case. The analysis starts by the statement of the relationship of the power at the aperture and in the far field.[20]

$$|A(z')|^2 dz' = a \frac{|E(\theta)|^2}{|G(\theta)|^2} \sin\theta d\theta \quad (2.53)$$

In the above expression $A(z')$ is the voltage distribution along the length of the array, $E(\theta)$ is the far field electric field pattern, and $G(\theta)$ is the far field pattern of a single radiating element. a is a constant which is set to ensure that the expression is dimensionally correct.

If we then combine the element and far field pattern terms we can use a total far field expression $F(\theta)$. Further, using $\cos\theta$ as the variable in equation (2.53) allows for the evaluation of the pattern outside of the visible region of the antenna pattern. The resulting expression is given as [20]

$$|A(z')|^2 dz' = a |F(\cos\theta)|^2 d(\cos\theta) \quad (2.53)$$

The source is discretized into linear elements of length dz' and each adjacent element is assumed to radiate into adjacent sectors of the far field pattern, as illustrated in Figure 2.6-1. Thus we can express the power radiated from the start of the source to a point z' on the line source as [20]

$$\int_0^{z'} |A(z')|^2 dz' = b \int_{\cos\theta_1}^{\cos\theta} |F(\cos\theta)|^2 d(\cos\theta) \quad (2.54)$$

The total radiated power is given by [20]

$$\int_0^L |A(z')|^2 dz' = b \int_{\cos\theta_1}^{\cos\theta_2} |F(\cos\theta)|^2 d(\cos\theta) \quad (2.55)$$

The result of equation (2.55) is that the left and right hand sides are essentially normalization terms thus dividing (2.54) by (2.55) we get [20]

$$\frac{\int_0^{z'} |A(z')|^2 dz'}{\int_0^L |A(z')|^2 dz'} = \frac{\int_{\cos\theta_1}^{\cos\theta} |F(\cos\theta)|^2 d(\cos\theta)}{\int_{\cos\theta_1}^{\cos\theta_2} |F(\cos\theta)|^2 d(\cos\theta)} \quad (2.56)$$

or

$$\frac{\int_0^{z'} |A(z')|^2 dz'}{\int_0^L |A(z')|^2 dz'} = \frac{\int_{\cos\theta_2}^{\cos\theta} |F(\cos\theta)|^2 d(\cos\theta)}{\int_{\cos\theta_2}^{\cos\theta_1} |F(\cos\theta)|^2 d(\cos\theta)} \quad (2.57)$$

depending on the selection of limits (see Figure 2.6-1). This expression can be evaluated numerically, graphically or

b) Effect of Reversing Angular Limits

Figure 2.6-1 Breakdown of optical synthesis model.

in closed form depending on the complexity of the defining field and source amplitude expressions. The result of evaluating this expression gives $\cos\theta$ as a function of z' . Thus the phase at any given point on the line source is given by [20]

$$\Psi(z') = \Psi(0) + \int_0^L \cos\theta(z') dz' \quad (2.58)$$

2.7 NUMERICAL OPTIMIZATION OF THE FAR FIELD

The optimization technique used in this work is a very simple perturbational technique. From the positions calculated by the phase requirements of the optical synthesis results, the positions of the radiating elements are perturbed slightly and the effect on the far field pattern is compared against some error criterion. Each patch is perturbed in turn and the change of most significance is kept. This process is repeated until either the desired solution is achieved or a minimum error condition exists.

Mathematically this can be stated

$$f'(a_1, a_2, a_3, \dots, a_N) = f(a_1, a_2, a_3, \dots, a_N) + \delta f(a_1, a_2, a_3, \dots, a_N) \quad (2.59)$$

If $f'()$ is better than $f()$ against some error criterion keep that change. Of the N changes possible keep only the best

change. In addition, the widths of the connecting lines have been allowed to become a variable. Thus changing the width of the connecting lines between radiating elements will affect the far field radiation pattern. In the same manner each feed line width is varied and the best change is kept.

CHAPTER 3

DESIGN AND TESTING OF A SERIES FED ARRAY

3.0 INTRODUCTION

This chapter provides an outline of the design and testing of a series fed microstrip array. The end use of the antenna is as an airborne radar antenna. The desired result from the synthesis of the series fed antenna is to provide a shaped beam in elevation.

At a systems level many considerations drive the antenna design. Mechanically the antenna needs to have a small aerodynamic cross section and low mass for it to be suitable for an airborne environment. Electrically the antenna should have high gain, a shaped elevation beam which will compensate for angle of incidence and slant range variations in the return power. From an economic standpoint the antenna should be relatively simple to manufacture and the design should be repeatable.

Microstrip is a technology which more or less meets these requirements. Not including the mechanical frame of the antenna, microstrip antennas are virtually two dimensional and

are much less massive than a waveguide antenna for example. The higher loss of the microstrip is offset by its ease of manufacture and repeatability of the design. The final design can be etched using standard photographic techniques to very good tolerances.

3.1 SERIES FED ELEMENT CHARACTERIZATION.

The series fed element as discussed in Chapter 2 was modelled as a cascaded network of radiation conductances steps in impedances and thin and fat transmission lines. The model of Jones et al. is repeated here in Figure 3.1-1

To characterize the parameters of this model it was necessary to fabricate several test items. An approximate sketch of one such test piece is shown in Figure 3.1-2. A through measurement is made on a line which is identical to the feed into the series element and then on the patch. Figure 3.1-3 shows a simplified test setup. The measurements were made on a network analyser and the differences in the transmission parameters between the two paths of Figure 3.1-2 yields the information with respect to the insertion loss and insertion phase of the patch de-embedded from the feed line. Five pieces were tested in turn and the extracted parameters were fitted to a polynomial expression and a plot of each are shown in Figures 3.1-4 to 3.1-8.

a) Physical Patch Element.

b) Transmission Line Sections.

c) Circuit Model of A Series Fed Patch.

Figure 3.1-1 Model for a series fed patch. [14]

Figure 3.1-2 Sample parameter extraction test piece.

Figure 3.1-3 Simplified representation of test setup.

Figure 3.1-4 Variation of microstrip characteristic impedance with the width of the line.

Figure 3.1-5 Variation of dielectric constant with feedline width.

Figure 3.1-6 Variation of radiation conductance with the difference between the feedline width and the patch width.

Figure 3.1-7 Variation of capacitive patch extension Δl with the difference between the feedline width and the patch width.

Figure 3.1-8 Variation of inductive feedline extension δ_L with the difference between the feedline width and the patch width.

3.2 ELEVATION BEAM PATTERN

In Figure 3.2-1 the geometry of the antenna - ground problem is shown. The required beam shape in elevation is a function of: 1) the far field footprint, which increases in size as the slant range increases; 2) the variation of the total path that the energy has to travel as a function of depression angle, and; 3) the variation of the backscatter as a function of the depression angle.

3.3 SYNTHESIS OF THE FAR FIELD PATTERN

In Sections 2.5 and 2.6 a technique for a phase only synthesis of a far field pattern was described. In Section 3.1 a determination of the variations of the transmission line parameters was made for the series fed patches. Using the curve fittings from that characterization process it is now possible to synthesize the desired far field pattern of the series fed array. The desired pattern is shown in Figure 3.3-1.

3.3.1 OPTICAL SYNTHESIS

Using the synthesis algorithm outlined in Sections 2.5 a FORTRAN program was written to perform the optical synthesis of the array. The algorithm for the optical synthesis actually assumes that the source is continuous and thus establishes the

ELEVATION BEAM SHAPE

INCIDENT POWER DENSITY

$$\propto GR^2$$

REFLECTED POWER

$$\propto (GR^2) R\sigma_0 (\cos(\gamma))^{-1}$$

RECEIVED POWER DENSITY

$$\propto GR^2 R\sigma_0 R^{-2} (\cos(\gamma))$$

POWER RECEIVED

$$\propto \frac{G^2 \sigma_0^2}{R^3 \cos \gamma}$$

$$R \propto (\sin \gamma)^{-1} = \csc \gamma$$

$$G(\gamma) \propto \left[\frac{\cos^{1/2} \gamma}{\sin^{3/2} \gamma \cdot 1.0000455 \cdot 10^{-3}} \right]$$

Figure 3.2-1 Illustration of the antenna to ground geometric relationship.

Desired Radiation Pattern
17 Element Series Fed Array
9.24 GHz

Figure 3.3-1 Optimal radiation pattern.

approximate geometry of the array.

The software was applied to the array and the progression from a uniformly spaced array to that of the optically synthesized result are shown in Figures 3.3.1-1 and 3.3.1-2. These are the initial and final far field patterns from the optical synthesis.

3.3.2 NUMERICAL OPTIMIZATION

As noted the optical synthesis yields only an approximate result. To continue the optimization a numerical variational technique was developed.

The purpose of this numerical optimization was threefold. Two far field parameters, the pattern itself and the ripple of that pattern in the shaped beam region, were the main priorities in the numerical design. While deviations of up to several dB could be tolerated in absolute beam shape, fast pattern ripples of only a few dB would seriously degrade antenna performance. In addition, the pattern ripple becomes more critical at high depression angles.

Uniformly Spaced
17 Element Series Fed Array
9.24 GHz

Figure 3.3.1-1 Starting pattern. Uniformly spaced series fed array.

After Optical Synthesis
17 Element Series Fed Array
9.24 GHz

Figure 3.3.1-2 Pattern after completion of optical synthesis.

The resulting error criteria for the shaped region of the beam were then given as:

- 1) A uniform weighting of the pattern at all points with the desired result that the absolute difference between the synthesized and the desired pattern should be minimized in the shaped region of the beam..
- 2) A weighting function which increases in magnitude as the depression angle increases and it is applied to the pattern ripple.

Outside the shaped sector of the beam the sidelobes were fitted to a maximum template. Of particular interest were those sidelobes which radiated energy in the region above the beam peak since in many airborne installations this leads to scattering of unwanted energy in the desired region creating ripples in the returns of the data.

To accomplish the final smoothing of the far field pattern according to the above criteria a FORTRAN program was written and applied to the geometry created by the optical synthesis. As noted in Section 2.6 this was accomplished by varying the separation between every patch during each iteration cycle. Only the best change in each iteration is kept. Figures 3.3.2-1 and 3.3.2-2 illustrate the pattern

2

Resolution test chart showing patterns of vertical and horizontal lines with numerical values: 1.0, 1.1, 1.25, 1.4, 1.6, 1.8, 2.0, 2.2, 2.5, 2.8, 3.2, 3.6, 4.0, 4.5, 5.0, 5.6, 6.3, 7.1, 8.0, 9.0, 10.

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After Coarse Optimization
17 Element Series Fed Array
9.24 GHz

Figure 3.3.2-1 Synthesized pattern after coarse optimization of the array.

After Fine Optimization
17 Element Series Fed Array
9.24 GHz

Figure 3.3.2-2 Pattern after completion of numerical synthesis.

changes as the optimization takes place.

3.4 PATTERN TESTING

Once the array was synthesized it underwent a series of far field measurements. A simplified sketch of the test facility is illustrated in Figure 3.4-1. The copolarized patterns at 9.19, 9.24 and 9.29 GHz were measured and are shown in Figures 3.4-2, 3.4-3 and 3.4-4 respectively. In addition, the cross polarized pattern at 9.24 GHz is shown in Figure 3.4-5.

In evaluating the plots several criteria must be applied. For the copolarized cases the main criterion is how well the measured patterns match the designed patterns. Upon examining the copolarized plots the patterns at 9.19 GHz and 9.24 GHz show a good deal of correlation with the designed pattern. At 9.29 GHz the pattern appears to be breaking up and forming large ripples in the shaped region. The final plot, that of the cross polarized reception at 9.24 GHz is an indication of the polarization purity of the generated pattern. At a system level, which is outside the scope of this thesis, the design goal was to provide at least 20 dB suppression of the cross polarized energy. From examining Figure 3.4-5 it is clear that

3.4-1 Simplified sketch of pattern acquisition setup.

3.4-2 Co-polarized pattern of 17 element series fed linear array at 9.19 GHz.

3.4-3 Co-polarized pattern of 17 element series fed linear array at 9.24 GHz.

3.4-4 Co-polarized pattern of 17 element series fed linear array at 9.29 GHz.

3.4-5 Cross polarized pattern of 17 element series fed linear array at 9.24 GHz.

the measured result exceeds this.

3.5 THE FINAL ARRAY

Figure 3.5-1 shows a one half scale plot of the final array geometry. The actual size is slightly shorter than 40 cm.

3.5-1 Scaled plot of the final 17 element series fed array.
Actual array is approximately 40 cm in length.

CHAPTER 4

DESIGN AND TESTING OF AN APERTURE COUPLED PATCH

4.0 INTRODUCTION

In Chapter 3 the design of a series fed array was outlined. This chapter will discuss a feed design intended to excite the orthogonal polarization from the same array.

4.1 SOURCES OF CROSS POLARIZED RADIATION IN A MICROSTRIP ANTENNA.

Before proceeding with the selection of a microstrip radiating element, it is necessary to understand where cross-polarized energy is generated in a microstrip patch array. There are three main sources.

The first contributor of cross polarized energy is energy which is radiated into the orthogonal polarization from the patches themselves. As shown in Chapter 2 this is generally not significant so long as the main beam of the antenna is oriented close to one of the principal planes of the microstrip patch element. This effect has been investigated and verified by Hansen [13] for both rectangular and circular patches. The other source of corruption of the polarization

purity in an array is due to the launching of surface waves at the edges of a microstrip patch.

Feed radiation is a pervasive source of extraneous radiation in microstrip antennas. While uniform lines generally can be considered as non-radiating, discontinuities and curved lines will launch both surface and space waves. The space waves generated are a direct source of cross-polarized radiation. The surface waves generated cause more subtle problems in that once generated they are free to reenter the array at any point and thus can be a considerable source of cross polarized radiation. As will be illustrated, careful selection of substrates, properly designed feed networks, and the use of shielding where necessary, can reduce this problem to acceptable levels.

The last and potentially most difficult source of cross-polarized radiation is attributable to coupling of the feed networks which provide the excitation of the two orthogonal polarizations. Any energy coupled between the two channels can be radiated in the orthogonal polarization state. Several different feed configurations have been examined experimentally in an effort to find a configuration which minimizes this problem.

4.2 ELEMENT SELECTION

The square patch has the advantage that the orthogonal polarizations are excited by modes which are inherently orthogonal due to the rectangular geometry of the patch. This means that misalignment of the feed positions are less likely to cause cross polarized radiation than they would in the case of the circular patch where a rotation of the feed point translates directly into a rotation of the polarization generated by that feed.

4.3 FEED SELECTION

Chapter 3 showed how a series fed array is capable of generating a shaped beam. A series fed array has the additional property that almost without exception all discontinuities are directly involved with generating the far field pattern and thus little extraneous radiation is produced. The orthogonal feed is another matter. The easiest way to feed the other polarization would be to print the orthogonal feed on the radiating skin using a coupled line. Each coupling point has several discontinuities associated with it and as a result would be capable of generating a great deal of crosspolarized radiation and thus reduce the performance of both polarizations. In addition line to line coupling of adjacent lines can be substantial.

Two possible solutions to this problem are to either shield the orthogonal feed line or move the orthogonal feed to the rear of the antenna where the groundplane of the radiating skin can act as a electromagnetic shield. While the first is problematical due to the large amounts of metal that would be placed on the radiating skin the second solution is not without its problems. Coupling the power through the ground plane is a non-trivial task.

4.3.1 PROBE COUPLING

The standard technique for moving power from back to front is by way of a pin which directly connects the feed line on the back to the patch on the front. Figure 4.3.1-1 shows a sketch of the model which was constructed and investigated experimentally. Several models of this were constructed and the isolation of the two feeds were made. It was found that the cross coupling between the feeds was on the order of 13 dB. This is unacceptably high and thus precludes the use of the arrangement.

4.3.2 APERTURE COUPLING

To overcome this difficulty a different feed technique for the horizontal channel was investigated. This feed type

Figure 4.3.1-1 Conventional probe coupling to achieve orthogonal polarization.

consists of an aperture cut in the ground plane which couples the feed line to the patch electromagnetically. This configuration was initially proposed by Pozar [15] and is illustrated in Figure 4.3.2-1. The rationale for choosing this excitation technique is that it approximates a longitudinal slot cut in the broad wall of a magnetic walled parallel plate waveguide (Figure 4.3.2-2) since a microstrip line or a patch can be represented approximately in this manner by applying equation (2.24). To first order the current in a parallel plate guiding structure flows along the longitudinal axis of the guide. From waveguide theory a longitudinal slot does not disturb the field structure in the guide and as a result does not tend to radiate any significant amount of power. Using this analogy the fields which are consistent with power flowing along the vertical axis of the vertical polarization channel will not appreciably excite a longitudinal slot cut in the ground plane. Further reciprocity dictates that energy entering on the horizontal feed line which is coupled through the slot should not couple into the orthogonal direction. To utilize this configuration it was necessary to be able to determine a geometry which yields an input impedance that was matched to the feed line impedance. To do this a Fortran program which calculates the associated parameters had to be developed. The source code is given as Appendix B.

When designing an aperture coupled feed, several

Figure 4.3.2-1 Novel hybrid dual polarized feed.

Figure 4.3.2-2 Parallel plate waveguide analogy applied to microstrip.

variables are available to the designer. The obvious ones include the patch width and length, and the length and width of the slot. In addition to these the substrate thickness and dielectric constant can be changed. The change of impedance with different lengths of aperture and changes in frequency are shown in Figures 4.3.2-3 and 4.3.2-4 respectively.

Using the model as a starting point, several test pieces similar to Figure 4.3.2-1 were constructed and the aperture coupled patch design was iterated until a reasonable impedance match was obtained. Figure 4.3.2-5 shows a plot of the measured S_{11} on a dB scale superimposed on the results of the computer model.

4.4 SMALL ARRAY DESIGN

With the single patch feed finalized the coupled line feed could be constructed. Figures 4.4-1 a) and b) show the drawings of the board which was constructed after two iterations. The iterations came about due to the lack of a good analytical model of the microstrip T junction. The design aim was to achieve a coupled power level equal to the coupled power per patch in the series feed.

Figure 4.3.2-3 Variation of input impedance of the aperture coupled patch with a change in aperture length.

Figure 4.3.2-4 Variation of input impedance of the aperture coupled patch with a change in frequency.

Figure 4.3.2-5 Measured versus calculated return loss for a single aperture coupled microstrip patch.

Figure 4.4-1 1:1 plot of the upper substrate of the small array.

Figure 4.4-2 1:1 Plot of the horizontal feed board of the small array.

4.4.1 PATTERN TESTS OF THE SMALL ARRAY

The far field patterns were measured for each of the two polarizations states. A simple sketch of the measurement range is shown in Figure 4.4.1-1. The co-polarized and cross polarized reception of radiation of each of the feed channels are shown in Figures 4.4.1-2 to 4.4.1-5. In this case the measurements were made to demonstrate that the small array was radiating energy into the desired polarization and to establish the cross polarized energy suppression of the two feed channels. Examining the measured results show that the desired radiation was present and that the small array displayed reasonable cross polarization characteristics.

4.4.2 CHANNEL TO CHANNEL ISOLATION

In this test the power coupled between the feed channels of the small array were measured at its ports. The aim of these measurements was to establish the level of on-board cross polarization. Figure 4.4.2-1 illustrates the test piece and shows the port designations used when the isolation measurements were made. Figures 4.4.2-2 and 4.4.2-3 show the isolation between the input and the two orthogonal outputs. By using symmetry and reciprocity the remaining cases are covered in these plots as well.

Figure 4.4.1-1 Simplified sketch of the far field pattern measurement range.

Figure 4.4.1-2 Measured co polarization pattern for vertically fed four element array.

Figure 4.4.1-3 Measured cross polarization pattern for vertically fed four element array.

Figure 4.4.1-4 Measured co polarization pattern for horizontally fed four element array.

Figure 4.4.1-5 Measured cross polarization pattern for horizontally fed four element array.

Front

Back

Figure 4.4.2-1 Illustration of the port designations for the two feeds on the small array.

Figure 4.4.2-2 Measured cross polarization isolation between ports 1 and 3 on the small array.

Figure 4.4.2-3 Measured cross polarization isolation between ports 1 and 4 on the small array.

CHAPTER 5

CONCLUSIONS

5.0 OVERALL PERFORMANCE

Chapters 3 and 4 illustrate the measured results of the development undertaken in this thesis. In Chapter 3 a series fed microstrip patch array was designed and the phase only synthesis was shown to produce excellent correlation of the design and achieved radiation patterns. The discrepancies can largely be attributed to an inadequate element pattern model, in the sense that it did not roll off fast enough at large angles from broadside. Further, the lack of mutual coupling in the synthesis model may account for the higher than expected first sidelobe.

The original intent of this work was to fully integrate the orthogonal polarization with the series fed array. While time and cost constraints did not allow this to happen, the results of the small array test indicate that the coupled patch performed well and the isolation characteristics were exceptional.

The experimental results of this thesis indicate that it is feasible to generate two essentially independent

polarizations from the same radiating elements using the design techniques described in this thesis.

5.2 FUTURE WORK

The most immediate extension of this work would be to complete the integration of the orthogonal feed. In addition, substantial work could be invested in increasing the bandwidth of the aperture feed. Other concepts worthy of further study include single point aperture feeds which generate circular polarizations from the patch. The possibilities for aperture coupling appear numerous and will probably lead to several interesting variations.

APPENDIX A

SPECTRAL DOMAIN METHOD OF MOMENTS

A.0 INTRODUCTION

The method of moments has been used as an analysis technique for some time. The work of Harrington [21] and others helped to bring this technique to the forefront as a powerful tool for the analysis of microwave structures. This method results in a rigorous 'full wave' solution with an accuracy limited only by the level of approximation used in the initial model. The solution is rigorous because the result is derived from a full mathematical treatment of the models. In the case of microstrip antennas this involves the use of dyadic Green's functions in the spectral or Fourier transform domain as opposed to the space domain since, to date, there exist no Green's functions that are applicable to the problem in the space domain.

Much of the early work in analyzing microstrip structures using method of moments was done by T. Itoh. In addition Itoh was the first to outline the approach of analyzing the microstrip resonator in the spectral domain [22].

A.1 GENERAL FORMULATION

To discuss the formulation of a Method of Moments an outline of Harrington's [21] introduction to the subject will be outlined. In general an electromagnetic problem may be formulated as

$$L(f)=g \quad (A.1)$$

where L is the operator, g is the driving or source function and, f is the field or response associated with the given driving expression. The problem is then further broken down into one of synthesis of the applicable operator L , where f and g are given, or analysis, where L and g are given and the response of the system, f , is required.

The method of moments provides a solution to the analysis problem where L and g are known. The response function f , is expanded into a series of known functions with unknown coefficients. i.e.

$$f=\sum_n \alpha_n f_n \quad (A.2)$$

With an infinite summation this expression will approach the exact solution to within an arbitrarily small error. However, for most situations a finite series will suffice and

the limit on the number of elements in the series is generally limited by the computing power available and the desired speed and accuracy of the solution. Assuming a linear operator L , equation (A.1) can be rewritten as

$$\sum_n \alpha_n L(f_n) = g \quad (\text{A.3})$$

At this point it is necessary to establish testing functions w_m and these are applied to the left and right hand side as an inner product of equ (A.3) with the result

$$\sum_n \alpha_n \langle w_m, L(f_n) \rangle = \langle w_m, g \rangle \quad m=1,2,3,\dots \quad (\text{A.4})$$

The resulting set of equations can be expressed in matrix form as

$$[l_{mn}] [\alpha_n] = [g_m] \quad (\text{A.5})$$

where

$$[l_{mn}] = \begin{bmatrix} \langle w_1, Lf_1 \rangle & \langle w_2, Lf_2 \rangle & \dots \\ \langle w_1, Lf_2 \rangle & \langle w_2, Lf_2 \rangle & \dots \\ \vdots & \vdots & \ddots \end{bmatrix} \quad (\text{A.6})$$

$$[\alpha_n] = \begin{bmatrix} \alpha_1 \\ \alpha_2 \\ \alpha_3 \\ \vdots \\ \vdots \end{bmatrix} \quad (\text{A.7})$$

$$[g_m] = \begin{bmatrix} \langle w_1, g \rangle \\ \langle w_2, g \rangle \\ \langle w_3, g \rangle \\ \vdots \\ \vdots \end{bmatrix} \quad (\text{A.8})$$

Then by inverting $[l_{mn}]$ the coefficients α_n are given by

$$[\alpha_n] = [l^{-1}][g_n] \quad (A.9)$$

The resulting solution is exact or an approximation depending on the choice of expansion functions in the problem formulation. For the case where $w_n = f_n$ the solution has been shown to be convergent [21] and is referred to as Galerkin's method. In addition, the choice of modes consistent with Galerkin's method results in a highly diagonalized matrix. Finally, choosing expansion functions which reasonably approximate reality yields a solution which will converge with relatively few expansion terms.

A.2 SPECTRAL DOMAIN GREEN'S FUNCTIONS

This section will present a reduced version of the derivation of Sullivan and Schaubert [23]. In addition only the Green's functions necessary for the analysis of this thesis will be included.

In the moment method analysis of the aperture coupled patch it is necessary to define some Spectral Domain Green's functions. The analysis of Sullivan and Schaubert starts by introducing the current quantities that are applicable to the coupled patch problem. These are electric currents in the x direction and the magnetic currents in the y direction. These represent the patch at the air dielectric boundary and the

slot in the ground plane respectively.

The Green's functions required are those which give the x directed electric field at the location of the patch and the y directed magnetic field at the aperture due to an x directed electric current on the patch. It is also necessary to have an expression which relates the x directed electric fields of the patch and the y directed magnetic field on the aperture to the magnetic current in the slot. We will start with the Green's functions for the electric current.

A.2.1 ELECTRIC CURRENT GREEN'S FUNCTIONS

For the case of an infinitesimal current in x at the air dielectric boundary the geometry can be expressed in the following differential system of equations. [23]

$$[\nabla^2 + \epsilon_r k_o^2] A_{x1}(x, y, z) = -\delta(x-x_o) \delta(y-y_o) \delta(z-h), \quad 0 < z < h \quad (A.10)$$

$$[\nabla^2 + \epsilon_r k_o^2] A_{z1}(x, y, z) = 0, \quad 0 < z < h \quad (A.11)$$

$$[\nabla^2 + k_o^2] A_{x2}(x, y, z) = 0, \quad z > h \quad (A.12)$$

$$[\nabla^2 + k_o^2] A_{z2}(x, y, z) = 0, \quad z > h \quad (A.13)$$

where

$$k_o^2 = \omega^2 \epsilon_o \mu_o \quad (A.14)$$

and ω , ϵ_o , and μ_o , are the radian frequency, the free space permittivity and permeability respectively. ϵ_r is the

dielectric constant of the substrate. Equations (A.10) - (A.13) are then Fourier transformed to the spectral domain with the result [23]

$$\left[\frac{d^2}{dz^2} + k_1^2 \right] \bar{A}_{x1}(k_x, k_y, z) = -\delta(z-h) e^{-j(k_x x_0 + k_y y_0)} \quad 0 < z < h \quad (\text{A.15})$$

$$\left[\frac{d^2}{dz^2} + k_1^2 \right] \bar{A}_{z1}(k_x, k_y, z) = 0, \quad 0 < z < h \quad (\text{A.16})$$

$$\left[\frac{d^2}{dz^2} + k_1^2 \right] \bar{A}_{x2}(k_x, k_y, z) = 0, \quad z > h \quad (\text{A.17})$$

$$\left[\frac{d^2}{dz^2} + k_1^2 \right] \bar{A}_{z2}(k_x, k_y, z) = 0, \quad z > h \quad (\text{A.18})$$

where

$$\bar{A}_{x1,2}(k_x, k_y, z) = \iint_{-\infty}^{\infty} \bar{A}_{x1,2}(x, y, z) e^{-j(k_x x_0 + k_y y_0)} dx dy \quad (\text{A.19})$$

$$\bar{A}_{z1,2}(k_x, k_y, z) = \iint_{-\infty}^{\infty} \bar{A}_{z1,2}(x, y, z) e^{-j(k_x x_0 + k_y y_0)} dx dy \quad (\text{A.20})$$

The quantities k_x and k_y are spectral variables. Solving (A.15 - A.20) with respect to the boundary conditions then yields the following magnetic vector potentials in the spectral domain. [23]

$$\bar{A}_{x1}(k_x, k_y, z) = \frac{1}{T_e} \sin(k_1 z) e^{-j(k_x x_0 + k_y y_0)} \quad (\text{A.21})$$

$$\bar{A}_{z1}(k_x, k_y, z) = \frac{k_x(\epsilon_r - 1) \sin(k_1 h)}{T_e T_m} \cos(k_1 z) e^{-j(k_x x_0 + k_y y_0)} \quad (\text{A.22})$$

$$\bar{A}_{x2}(k_x, k_y, z) = \frac{1}{T_e} \sin(k_1 h) e^{-jk_2(z-h)} e^{-j(k_x x_0 + k_y y_0)} \quad (\text{A.23})$$

$$\bar{A}_{z2}(k_x, k_y, z) = \frac{k_x(\epsilon_r - 1) \sin(k_1 h) \cos(k_1 h)}{T_e T_m} e^{-jk_2(z-h)} e^{-j(k_x x_0 + k_y y_0)} \quad (\text{A.24})$$

At the interface we get the required result that the vector potentials are equal. The electric field components at $z = h$ from an x-directed dipole are then derived from [23]

$$E_x(k_x, k_y, z) = \frac{1}{j\omega\epsilon_r\epsilon_0} \left[(\epsilon_r k_0^2 - k_x^2) \tilde{A}_x + jk_x \frac{d}{dz} \tilde{A}_z \right] \quad (A.25)$$

$$E_y(k_x, k_y, z) = \frac{1}{j\omega\epsilon_r\epsilon_0} \left[jk_y \frac{d}{dz} \tilde{A}_z \right] \quad (A.26)$$

The required Green's functions in the space domain are then

The Green's function for x-directed electric fields from x-directed current sources. [23]

$$G_{EJxx}(x, y, h | x_0, y_0, h) = \iint_{-\infty}^{\infty} Q_{EJxx}(k_x, k_y) e^{j(k_x(x-x_0) + k_y(y-y_0))} dk_x dk_y \quad (A.27)$$

$$Q_{EJxx}(k_x, k_y) = -j \frac{Z_0}{4\pi^2 k_0} \left[\frac{(\epsilon_r k_0^2 - k_x^2) k_2 \cos(k_1 h)}{T_e T_m} + \frac{j(k_0^2 - k_x^2) k_1 \sin(k_1 h)}{T_e T_m} \right] \sin(k_1 h) \quad (A.28)$$

where

$$T_e = k_1 \cos(k_1 d) + jk_2 \sin(k_1 d), \quad (A.29)$$

$$T_m = \epsilon_r k_2 \cos(k_1 d) + jk_1 \sin(k_1 d), \quad (A.30)$$

$$k_1^2 = \epsilon_r k_0^2 - \beta^2, \quad \text{Im } k_1 < 0 \quad (A.31)$$

$$k_2^2 = k_0^2 - \beta^2, \quad \text{Im } k_2 < 0 \quad (A.32)$$

$$\beta^2 = k_x^2 + k_y^2, \quad (A.33)$$

$$k_0 = \omega \sqrt{\mu_0 \epsilon_0} = 2\pi/\lambda_0, \quad (A.34)$$

$$Z_0 = \sqrt{\mu_0 / \epsilon_0}, \quad (A.35)$$

A.2.2 MAGNETIC CURRENT GREEN'S FUNCTIONS

From this point we can also define the y directed magnetic field due to the x directed electric current element. The signs of k_1 and k_2 are to ensure the proper path of integration. We now have from classical theory and transformed into the spectral domain [23]

$$H_y = \frac{d}{dz}A_x - jk_x A_z \quad (A.37)$$

The resulting function then relates the Y directed patch current to the Y directed aperture magnetic field. [23]

$$G_{HJyx}(x, y, 0 | x_0, y_0, h) = \iint_{-\infty}^{\infty} Q_{HJyx}(k_x, k_y) e^{j(k_x(x-x_0)+k_y(y-y_0))} dk_x dk_y \quad (A.38)$$

$$Q_{HJyx}(k_x, k_y) = -j \frac{1}{4\pi^2} \left[\frac{k_x^2 (\epsilon_r - 1) \sin(k_1 d)}{T_e T_m} + \frac{k_1}{T_e} \right] \quad (A.39)$$

The one remaining quantity is the expression which relates the field quantities to the fictitious magnetic current in the aperture. The analysis is similar to the above but the magnetic vector potentials A are replaced by electric vector potentials F.

From the analysis the resulting expressions are for the y directed magnetic field in the aperture due to the y directed magnetic current in the slot [23]

$$G_{HMYY}(x, y, 0 | x_0, y_0, 0) = \int_{-\infty}^{\infty} Q_{HMYY}(k_x, k_y) e^{j(k_x(x-x_0)+k_y(y-y_0))} dk_x dk_y \quad (A.40)$$

$$Q_{HMYY}(k_x, k_y) = \frac{1}{4\pi^2 k_0 Z_0} \left[\frac{(\epsilon_r k_0^2 - k_y^2)(k_1 \cos(k_1 d) + j k_2 \epsilon_r \sin(k_1 d))}{k_1 T_m} - \frac{k_1 k_y^2 (\epsilon_r - 1)}{T_e T_m} \right] \quad (A.41)$$

Finally, the expression which relates the slot current to the X directed patch fields [23]

$$G_{EMXY}(x, y, 0 | x_0, y_0, 0) = \int_{-\infty}^{\infty} Q_{EMXY}(k_x, k_y) e^{j(k_x(x-x_0)+k_y(y-y_0))} dk_x dk_y \quad (A.42)$$

$$Q_{EMXY}(k_x, k_y) = j \frac{1}{4\pi^2} \left[\frac{k_x^2 (\epsilon_r - 1) \sin(k_1 d)}{T_e T_m} - \frac{k_1}{T_e} \right] \quad (A.43)$$

APPENDIX B

B.0 INTRODUCTION

This appendix provides a brief overview of the software which was developed in this research for calculating the parameters associated with the aperture coupled feed. While the equations were outlined in Appendix A this appendix will discuss some of the finer points of the coding.

B.1 NUMERICAL INTEGRATION

The equations of Appendix A present several challenges with respect to their evaluation numerically. The integrand is oscillatory and the order of the oscillations increases with the order of the basis functions used over the patch. In addition the denominator goes to zero at those locations where the integration variable β is such that the T_e or the T_m terms go to zero. In general only the T_m term is of any relevance due to the presence of the 0 cutoff slab mode TM0.

B.2 CONVERGENCE

Convergence of the integrals was shown to occur for $\beta > 150k_0$ [16]. To ensure that the oscillatory nature of the

integrand was not a factor it was generally required that the integrand be evaluated using a high order gaussian quadrature technique, and that the integrand be evaluated in steps of β_0 , where β_0 is the propagation constant of the TM₀ mode in the slab. This results in the evaluation of approximately 4000 points for each integral. This is compared to a rectangular rule approach which was found to require several tens of thousands of points to achieve the same convergence.

B.3 EVALUATION OF THE TM₀ POLE.

The propagation constant of the TM₀ mode is determined by finding the zero of the T_m expression. A FORTRAN program was written using the Newton - Raphson root finding technique to determine the zero of T_m . As a part of the algorithm it was necessary to determine the derivative of T_m with respect to β . The resulting expression is given as T_m' .

$$\beta_{new} = \beta_{old} - \frac{T_m(\beta)}{T_m'(\beta)} \quad (B.1)$$

where

$$T_m'(\beta) = -\beta((\epsilon_r \cos(k_1 d)/k_2 - \epsilon_r * k_2 d \sin(k_1 d)/k_1) + j(\sin(k_1 d)/k_1 + d \cos(k_1 d))) \quad (B.2)$$

With this the root is then found iteratively through repeated evaluation of equation B.1.

Once the location of the zero is known with respect to β there is then the problem remains of how to handle it numerically. This was addressed by Pozar [18] who suggested that the integrand be broken into three parts. The three components represent the real, pole contribution, and the imaginary component of the integral respectively.

B.4 EVALUATING THE PARAMETERS OF THE MODEL

The remaining software was written to evaluate the V_n , Z_{mn} , ΔV , and Y_e terms. The source code for these calculations is given. Each of the programs follows the same basic logic and only the specific equations, for example the Green's functions, are changed in each. Each program reads the same geometry file, thus allowing for streamlined entry of physical parameters, initializes the integration variables and then processes the result. Each program creates a small file which contains the matrix result of the calculations.

After each of the files has been executed a final piece of software is executed. For completeness this has been included at the end of the code listings. Essentially it inverts the patch self impedance matrix and then carries out the required multiplications, thus resulting in the series impedance that the slot patch pair presents to the feedline.

C*****

C Impedanc.for

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C Program Function -

C This program is responsible for calculating the self and
C mutual interaction terms for the imposed current distributions
C on the patch. The currents are assumed to flow in the X direction
C only and is uniform across the width of the patch.

C External Functions

C TE - This function evaluates the expression for Te as a
C function of the integration variable Beta
C TM - This function evaluates the expression for Tm as a
C function of the integration variable Beta
C TMPRIME - This function is the analytical derivative of Tm
C and it is evaluated at Beta
C FU - Fourier Transform of a uniform current source.
C FS - Fourier Transform of a Sinusoidal current source.
C GEJXX - Spectral Domain Green's function relating the
C electric current along X on the patch to the
C electric field along X on the patch.

C Program Variables

C Z(m,n) - Matrix of self and mutual reactances of the patch
C currents.
C K1 , K2 - Dependent variables. Function of BETA, K0, and ER.
C J - Square root of -1.
C D - Thickness of the substrate.
C ER - Relative dielectric constant of the substrate material.
C ERE - Assumed dielectric constant of the slot. Average of
C the effective dielectric constant of the patch and the
C feedline.
C EPSEFF - Effective dielectric constant of the feedline.
C FREQ - Holds the frequency of operation.
C LAP - Length of the aperture.
C WAP - Width of the aperture.
C PI - 3.1416
C K0 - Free space wave number.
C FUKY - Temporary variable. Holds the result of one calculation
C of function FU.
C FSMKX - Temporary variable. Holds the result of one calculation
C of function FS.
C KEFF - Slot wave number.
C KX - Secondary integration variable.
C KY - Secondary integration variable.
C ZM - Impedance of the microstrip feedline.
C C - Speed of light in free space.
C ETA - Intrinsic impedance of free space.
C WF - Width of the microstrip feed line.
C XR - Root locations in Gaussian Quadrature integration.
C WT - Weighting constants for Gaussian Integration.
C ALP - Primary integration variable.
C AWT - ALP weighting.
C BT - Primary integration variable.
C BTWT - BT weighing.

```

C      BETAO - Pole value of beta.
C      LP - Length of the patch.
C      WP - Width of the patch.
C      CTn - Counting variables.
C      NA - Number of basis modes on patch.
C      UPPER - number of integration cells.

```

```

C

```

```

C*****

```

```

PROGRAM IMPEDANCE
EXTERNAL TM,TMPRIME,FU,FS,TE,GEJXX
DOUBLE COMPLEX Z(12,12),GEJXX,K1,K2,TE,TM,TMPRIME,J
DOUBLE PRECISION D,ER,FREQ,LP,WP,PI,K0,XR(30),WT(60),
1FUKY,FSMKX,FS,FU,KX,KY,C,ETA,ALP(60),AWT,BT(30),BETA0,BTWT,
2LAP,WAP,WF,ZM,ERE,EPSEFF
INTEGER NA,M,N,CT0,CT1,CT2,UPPER

```

```

C*****

```

```

C      Define standard constants.

```

```

C*****

```

```

J=DCMPLX(0.0,1.0)
PI=DACOS(-1.0)
C=1/DSQRT(PI*35.416E-19)
ETA=C*PI*4E-07

```

```

C*****

```

```

C      Read the geometry of the problem.

```

```

C*****

```

```

OPEN (UNIT=201,FILE='GEO.DAT',STATUS='OLD')
READ (201,*) LP,WP
WRITE (*,*) ' LENGTH OF THE PATCH      - ', LP
WRITE (*,*) ' WIDTH OF THE PATCH      - ', WP
READ (201,*) LAP,WAP
READ (201,*) WF,ZM
READ (201,*) D,ER,ERE,EPSEFF
WRITE (*,*) ' SUBSTRATE THICKNESS      - ', D
WRITE (*,*) ' DIELECTRIC CONSTANT      - ', ER
READ (201,*) FREQ
WRITE (*,*) ' CENTRE FREQUENCY [GHZ]   - ', FREQ
FREQ=FREQ*1E09
READ (201,*) NA
WRITE (*,*) ' NUMBER OF CURRENT MODES - ', NA
READ (201,*) UPPER
CLOSE(201)

```

```

OPEN (UNIT=202,FILE='NEWT.DAT',STATUS='OLD')
READ (202,*) BETAO
WRITE (*,*) ' BETA AT TMO POLE        - ', BETAO
CLOSE(202)

```

```

C*****

```

```

C      Define the wavenumber for free space.

```

```

C*****

```

```

K0 = 2*PI*FREQ/C

```

```

DO 10 M=1,NA
DO 20 N=1,NA
  Z(M,N)=DCMPLX(0.0)

```

```

20 CONTINUE

```

```

10 CONTINUE

```

```

C*****
C   Read in the roots and weighting coefficients for the gaussian
C   quadrature integration.
C*****
OPEN (UNIT=203,FILE='XR30.DAT' ,STATUS='OLD')
OPEN (UNIT=204,FILE='WGT30.DAT',STATUS='OLD')
DO 30 CT0=30,1,-1
    READ (203,*) XR(CT0)
    READ (204,*) WT(CT0)
30  CONTINUE
    CLOSE(204)
    CLOSE(203)

C*****
C   Start the main integration loop.
C*****
AWT=PI/8.0
DO 40 CT0=1,30
    ALP(CT0) = AWT*XR(CT0) + PI/8.0
    ALP(CT0+30)= ALP(CT0) + PI/4.0
    WT(CT0+30) = WT(CT0)
40  CONTINUE

    OPEN(UNIT=205,FILE='IMP.OUT',STATUS='UNKNOWN')

    DO 50 CT0=1,UPPER
C*****
C   Define the points of integration.
C*****
        SELECT CASE(CT0)
        CASE(1)
            BTWT=0.999*BETA0/2.0
            DO 51 CT1=1,30
                BT(CT1)=BTWT*XR(CT1)+0.999*BETA0/2.0
51         CONTINUE
        CASE(2)
            BTWT=0.999*BETA0/2.0
            DO 52 CT1=1,30
                BT(CT1)=BTWT*XR(CT1)+3.001*BETA0/2.0
52         CONTINUE
        CASE(3:)
            BTWT=2*BETA0
            DO 53 CT1=1,30
                BT(CT1)=BTWT*XR(CT1)+(4*CT0-8)*BETA0
53         CONTINUE
        END SELECT
        DO 60 CT1=1,60
        DO 70 CT2=1,30
            KX=BT(CT2)*DCOS(ALP(CT1))
            KY=BT(CT2)*DSIN(ALP(CT1))
            K1=DCONJG(CDSQRT(DCMPLX(ER*K0**2-BT(CT2)**2)))
            K2=DCONJG(CDSQRT(DCMPLX( K0**2-BT(CT2)**2)))
            DO 80 M=1,NA
                FSMKX=FS((2*M-1),KX,LP/2.0)
                FUKY =FU(KY,WP)**2
                DO 90 N=M,NA
                    Z(M,N)=Z(M,N)+AWT*BTWT*WT(CT1)*WT(CT2)*BT(CT2)*
1                     GEJXX(BT(CT2),KX,K0,K1,K2,D,ER,TH)*
2                     FSMKX*FUKY*FS((2*N-1),KX,LP/2.0)
90        CONTINUE

```

```

80      CONTINUE
70      CONTINUE
60      CONTINUE
      WRITE (*,999) Z((NA-1),NA),Z(NA,NA)
50      CONTINUE
999     FORMAT(' ',6F13.5)

```

130

```

C*****
C      Integration routine for determining the contribution of the
C      surface wave pole.
C*****

```

```

      DO 100 CT1=1,60
        KX=BETA0*DCOS(ALP(CT1))
        KY=BETA0*DSIN(ALP(CT1))
        K1=DCONJG(CDSQRT(DCMPLX(ER*K0**2-BETA0**2)))
        K2=DCONJG(CDSQRT(DCMPLX( K0**2-BETA0**2)))
        DO 110 M=1,NA
          FSMKX=FS((2*M-1),KX,LP/2.0)
          FUKY =FU(KY,WP)**2
          DO 120 N=M,NA
            Z(M,N)=Z(M,N)-J*PI*AWT*WT(CT1)*BETA0*
1          GEJXX(BETA0,KX,K0,K1,K2,D,ER,TMPRIME)*
2          FSMKX*FUKY*FS((2*N-1),KX,LP/2.0)
120      CONTINUE
110      CONTINUE
100      CONTINUE

```

```

C*****
C      Complete the final multiplication and output result to IMP.OUT
C*****

```

```

      DO 130 M=1,NA
        DO 140 N=M,NA
          Z(M,N)=Z(M,N)*J*ETA/(K0*PI**2)
          Z(N,M)=Z(M,N)
140      CONTINUE
130      CONTINUE

```

```

      DO 150 M=1,NA
        DO 160 N=1,NA
          WRITE (205,*) M,N,Z(M,N)
          WRITE ( *,*) M,N,Z(M,N)
160      CONTINUE
150      CONTINUE

```

```

      CLOSE(205)
      STOP
      END

```

```

C*****
C      GEJXX - Spectral Domain Green's function relating the
C               electric current along X on the patch to the
C               electric field along X over the patch.
C*****

```

```

      DOUBLE COMPLEX FUNCTION GEJXX(BETA,KX,K0,K1,K2,D,ER,TM)
      DOUBLE COMPLEX    K1,K2,TE,TM,J,PROD
      DOUBLE PRECISION  K0,D,ER,PI,KX,BETA
      PI=DACOS(-1.0)
      PROD=K1*D
      J=DCMPLX(0.0,1.0)

```

```

      GEJXX=(CDSIN(PROD)/TE(D,K1,K2))
      +*((ER*K0**2-KX**2)*K2*CDCOS(PROD)+
      +   J*(K0**2-KX**2)*K1*CDSIN(PROD))/TM(D,K1,K2,BETA,ER)
      RETURN
      END

```

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```

C*****
C      FS-      Fourier Tranform of a Sinusoidal current source.
C*****

```

```

      DOUBLE PRECISION FUNCTION FS(MODE,KX,A)
      DOUBLE PRECISION KX,A,PI
      INTEGER MODE
      PI=DACOS(-1.0)
      FS=MODE*PI*DCOS(KX*A)/(A*((MODE*PI/(2.0*A))**2-KX**2))
      RETURN
      END

```

```

C*****
C      FU -      Fourier Tranform of a uniform current source.
C*****

```

```

      DOUBLE PRECISION FUNCTION FU(KY,W)
      DOUBLE PRECISION KY,W
      FU=DSIN(KY*W/2.0)/(KY*W/2.0)
      RETURN
      END

```

```

C*****
C      TE -      This function evaluates the expression for Te as a
C                  function of the integration variable Beta
C*****

```

```

      DOUBLE COMPLEX FUNCTION TE(D,K1,K2)
      DOUBLE COMPLEX K1,K2
      DOUBLE PRECISION D
      TE=K1*CDCOS(K1*D)+DCMPLX(0.0,1.0)*K2*CDSIN(K1*D)
      RETURN
      END

```

```

C*****
C      TM -      This function evaluates the expression for Tm as a
C                  function of the integration variable Beta
C*****

```

```

      DOUBLE COMPLEX FUNCTION TM(D,K1,K2,BETA,ER)
      DOUBLE COMPLEX K1,K2
      DOUBLE PRECISION D,BETA,ER
      TM=ER*K2*CDCOS(K1*D)+DCMPLX(0.0,1.0)*K1*CDSIN(K1*D)
      RETURN
      END

```

```

C*****
C      TMPRIME - This function is the analytical derivative of Tm
C                  and it is evaluated at Beta
C*****

```

```

      DOUBLE COMPLEX FUNCTION TMPRIME(D,K1,K2,BETA,ER)
      DOUBLE COMPLEX K1,K2
      DOUBLE PRECISION D,BETA,ER
      TMPRIME=-BETA*((ER*CDCOS(K1*D)/K2-ER*K2*D*CDSIN(K1*D)/K1)+
      +   DCMPLX(0.0,1.0)*(CDSIN(K1*D)/K1+D*CDCOS(K1*D)))
      RETURN
      END

```



```

C WP -      Width of the patch.
C CTn -     Counting variables.
C NA -      Number of basis modes on patch.
C UPPER -   number of integration cells.
C
C*****

```

```

PROGRAM DELV

```

```

EXTERNAL TM,TMPRIME,FU,FP,TE,GHJYX

```

```

DOUBLE COMPLEX DV,TM,GHJYX,K1,K2,TMPRIME,TE,J

```

```

DOUBLE PRECISION D,ER,ERE,EPSEFF,FREQ,LAP,WAP,PI,KO,FUKY,
1FPKY,FP,FU,KEFF,KX,KY(30),ZM,C,ETA,KYWT,WF,XR(30),WT(30),BETA,
2LP,WP

```

```

INTEGER CT0,CT1,CT2,NA,UPPER

```

```

C*****
C   Define standard constants.
C*****

```

```

J=DCMPLX(0.0,1.0)
PI=DACOS(-1.0)
C=1/DSQRT(PI*35.416E-19)
ETA=C*PI*4E-07

```

```

C*****
C   Read the geometry of the problem.
C*****

```

```

OPEN (UNIT=201,FILE='GEO.DAT',STATUS='OLD')
READ (201,*) LP,WP
READ (201,*) LAP,WAP
WRITE (*,*) ' LENGTH OF THE APERTURE - ', LAP
WRITE (*,*) ' WIDTH OF THE APERTURE - ', WAP
READ (201,*) WF,ZM
WRITE (*,*) ' WIDTH OF THE FEEDLINE - ', WF
WRITE (*,*) ' FEEDLINE IMPEDANCE - ', ZM
READ (201,*) D,ER,ERE,EPSEFF
WRITE (*,*) ' SUBSTRATE THICKNESS - ', D
WRITE (*,*) ' DIELECTRIC CONSTANT - ', ER
WRITE (*,*) ' SLOT CONSTANT - ', ERE
WRITE (*,*) ' FEEDLINE CONSTANT - ', EPSEFF
READ (201,*) FREQ
WRITE (*,*) ' CENTRE FREQUENCY [GHZ] - ', FREQ
FREQ=FREQ*1E09
READ (201,*) NA
READ (201,*) UPPER
CLOSE(201)

```

```

C*****
C   Define the wavenumbers for free space, the slot, and the
C   propagation constant along the feed line.
C*****

```

```

KO - 2*PI*FREQ/C
KEFF - KO*DSQRT(ERE)
KX - -KO*DSQRT(EPSEFF)

```

```

C*****
C    Read in the roots and weighting coefficients for the gaussian
C    quadrature integration.
C*****

```

```

OPEN (UNIT=203,FILE='XR30.DAT',STATUS='OLD')
OPEN (UNIT=204,FILE='WGT30.DAT',STATUS='OLD')
DO 30 CT0=30,1,-1
    READ (203,*) XR(CT0)
    READ (204,*) WT(CT0)
30  CONTINUE
CLOSE(204)
CLOSE(203)

```

```

C*****
C    Start the main integration loop.
C*****

```

```

OPEN(UNIT=205,FILE='DV.OUT',STATUS='UNKNOWN')

DV=DCMPLX(0.0)

```

```

DO 50 CT0=-UPPER,UPPER-1

```

```

C*****
C    Define the points of integration.
C*****

```

```

    KYWT=125.0/2.0
    DO 51 CT1=1,30
        KY(CT1)=KYWT*XR(CT1)+(2*CT0+1)*125.0/2.0
51  CONTINUE
    DO 70 CT2=1,30
        BETA=DSQRT(KX**2+KY(CT2)**2)
        K1=DCONJG(CDSQRT(DCMPLX(ER*K0**2-BETA**2)))
        K2=DCONJG(CDSQRT(DCMPLX(  K0**2-BETA**2)))

        FPKY=FP(KEFF,KY,LAP/2.0)
        FUKY =FU(KY,WF)

```

```

C*****
C    Continue summing of the variable dv.
C*****

```

```

    DV=DV+KYWT*WT(CT2)*
1      GHJYX(BETA,KX,K1,K2,D,ER,TH)*
2      FPKY*FUKY
70  CONTINUE
50  CONTINUE
999  FORMAT(' ',4F13.5)

```

```

C*****
C    Complete the final multiplication and output result to DV.OUT
C*****

```

```

DV = DV/(2*PI*DSQRT(ZM))

```

```

WRITE (205,*) DV
WRITE (  *,*) DV

```

```

CLOSE(205)
STOP
END

```

```

C*****
C GHJYX - Spectral Domain Green's function relating the
C electric current along X on the feed line to the
C magnetic field along Y over the aperture.
C*****

```

```

DOUBLE COMPLEX FUNCTION GHJYX(BETA,KX,K1,K2,D,ER,TM)
DOUBLE COMPLEX K1,K2,TE,TM,J,PROD
DOUBLE PRECISION D,ER,PI,KX,BETA
PI=DACOS(-1.0)
PROD=K1*D
J=CMPLX(0.0,1.0)
GHJYX=(-1.0/TE(D,K1,K2))*
+((J*KX**2*(ER-1.0)*CDSIN(PROD)/TM(D,K1,K2,BETA,ER))-K1)
RETURN
END

```

```

C*****
C FP - Fourier Transform of a Piece Wise Sinusoidal current
C source.
C*****

```

```

DOUBLE PRECISION FUNCTION FP(KE,KY,H)
DOUBLE PRECISION KE,KY,H,PI
PI=DACOS(-1.0)
FP=2*KE*(DCOS(KY*H)-DCOS(KE*H))/((KE**2-KY**2)*DSIN(KE*H))
RETURN
END

```

```

C*****
C FU - Fourier Transform of a uniform current source.
C*****

```

```

DOUBLE PRECISION FUNCTION FU(KY,W)
DOUBLE PRECISION KY,W
FU=DSIN(KY*W/2.0)/(KY*W/2.0)
RETURN
END

```

```

C*****
C TE - This function evaluates the expression for Te as a
C function of the integration variable Beta
C*****

```

```

DOUBLE COMPLEX FUNCTION TE(D,K1,K2)
DOUBLE COMPLEX K1,K2
DOUBLE PRECISION D
TE=K1*CDCOS(K1*D)+DCMPLX(0.0,1.0)*K2*CDSIN(K1*D)
RETURN
END

```

```

C*****
C TM - This function evaluates the expression for Tm as a
C function of the integration variable Beta
C*****

```

```

DOUBLE COMPLEX FUNCTION TM(D,K1,K2,BETA,ER)
DOUBLE COMPLEX K1,K2
DOUBLE PRECISION D,BETA,ER
TM=ER*K2*CDCOS(K1*D)+DCMPLX(0.0,1.0)*K1*CDSIN(K1*D)
RETURN
END

```

```

C*****

```

C TPRIME - This function is the analytical derivative of Tm
C and it is evaluated at Beta

C*****

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```
DOUBLE COMPLEX FUNCTION TPRIME(D,K1,K2,BETA,ER)
DOUBLE COMPLEX      K1,K2
DOUBLE PRECISION D,BETA,ER
TPRIME--BETA*((ER*CDCOS(K1*D)/K2-ER*K2*D*CDSIN(K1*D)/K1)+
+          DCMLX(0.0,1.0)*(CDSIN(K1*D)/K1+D*CDCOS(K1*D)))
RETURN
END
```

```

C*****
C
C   Newton.for
C
C   Program Function -
C
C       This program is responsible for calculating the zero of the
C   Tm expression.
C
C   External Functions
C
C       TM -      This function evaluates the expression for Tm as a
C                 function of the integration variable Beta
C       TMPRIME - This function is the analytical derivative of Tm
C                 and it is evaluated at Beta
C
C   Program Variables
C
C       K1 , K2 -  Dependent variables. Function of BETA, K0, and ER.
C       D -       Thickness of the substrate.
C       ER -       Relative dielectric constant of the substrate material.
C       ERE -      Assumed dielectric constant of the slot. Average of
C                 the effective dielectric constant of the patch and the
C                 feedline.
C       EPSEFF -   Effective dielectric constant of the feedline.
C       FREQ -     Holds the frequency of operation.
C       LAP -      Length of the aperture.
C       WAP -      Width of the aperture.
C       PI -       3.1416
C       K0 -       Free space wave number.
C       ZM -       Impedance of the microstrip feedline.
C       WF -       Width of the microstrip feed line.
C       BETA -     Independent variable.
C       LP -       Length of the patch.
C       WP -       Width of the patch.
C       CTO -      Counting variable.
C       NA -       Number of basis modes on patch.
C       UPPER -   number of integration cells.
C
C*****

```

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```

PROGRAM NEWTON
DOUBLE COMPLEX    TM,TMPRIME,K1,K2
DOUBLE PRECISION BETA,D,ER,K0,FREQ,LP,WP,LAP,WAP,WF,ZM,ERE,
1EPSEFF
INTEGER NA,CTO,UPPER

```

```

C*****
C   Define PI.
C*****
PI=DACOS(-1.0)

```

```

C*****
C   Read the geometry of the problem.
C*****
OPEN (UNIT=201,FILE='GEO.DAT',STATUS='OLD')
READ (201,*) LP,WP
READ (201,*) LAP,WAP

```

```

      READ (201,*) WF,ZM
      READ (201,*) D,ER,ERE,EPSEFF
      WRITE (*,*) ' SUBSTRATE THICKNESS      - ', D
      WRITE (*,*) ' DIELECTRIC CONSTANT      - ', ER
      READ (201,*) FREQ
      WRITE (*,*) ' CENTRE FREQUENCY [GHZ]    - ', FREQ
      FREQ=FREQ*1E09
      READ (201,*) NA
      READ (201,*) UPPER
      CLOSE(201)

```

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```

C*****
C   Define the wavenumber for free space.
C*****

```

```

      KO=2*PI*FREQ*DSQRT(8.854E-12*4*PI*1E-07)
      PRINT *,KO

```

```

C*****
C   Initial guess at the root.
C*****

```

```

      BETA=1000.0

```

```

C*** ~*****
C   Implement Newton Raphson Root finding algorithm.
C*****

```

```

      DO 10 CT0=1,50
      K1=DCONJG(CDSQRT(DCMPLX((ER*KO**2-BETA**2),0.0)))
      K2=DCONJG(CDSQRT(DCMPLX(( KO**2-BETA**2),0.0)))

      BETA=BETA-TM(D,K1,K2,BETA,ER)/TMPRIME(D,K1,K2,BETA,ER)

      WRITE(*,100) BETA,TM(D,K1,K2,BETA,ER)

```

```

10    CONTINUE

```

```

C*****
C   Save the result.
C*****

```

```

      OPEN (UNIT=205,FILE='NEWT.DAT',STATUS='UNKNOWN')
      WRITE (205,*) BETA
      CLOSE (205)

```

```

      STOP
100   FORMAT(' ',3F25.12)
      END

```

```

C*****
C   TM - This function evaluates the expression for Tm as a
C         function of the integration variable Beta
C*****

```

```

      DOUBLE COMPLEX FUNCTION TM(D,K1,K2,BETA,ER)
      DOUBLE COMPLEX K1,K2
      DOUBLE PRECISION D,BETA,ER
      TM=ER*K2*CDCOS(K1*D)+DCMPLX(0.0,1.0)*K1*CDSIN(K1*D)
      RETURN

```

END

```
C*****
C      TMPRIME - This function is the analytical derivative of Tm
C              and it is evaluated at Beta
C*****
```

139

```
DOUBLE COMPLEX FUNCTION TMPRIME(D,K1,K2,BETA,ER)
  DOUBLE COMPLEX K1,K2
  DOUBLE PRECISION D,BETA,ER
  TMPRIME=-BETA*((ER*CDCOS(K1*D)/K2-ER*K2*D*CDSIN(K1*D)/K1)+
+      DCMPLX(0.0,1.0)*(CDSIN(K1*D)/K1+D*CDCOS(K1*D)))
  RETURN
END
```

C*****

C

C Patchad.for

C

C Program Function -

140

C

C This program is responsible for processing the outputs of the
C programs IMPEDANC.FOR, VOLT.FOR, YAD.FOR and DELTA.FOR. The
C program reads the computed results of each and then calculates
C the series impedance seen on the feedline.

C

C Program Variables

C Z - Matrix of self and mutual reactances of the patch
C current modes.
C V - Matrix of coupling values between the slot and the
C patch current modes.
C TEMP - Temporary transfer matrix.
C RESULT - Holds the result of the inverse of the Z matrix
C times the square of the V matrix.
C YE - Aperture self admittance term.
C DV - Discontinuity vector on the feed line.
C INDEXn - Dummy variables.
C M,N Array indices.
C J,K Counting variables.
C NA Number of patch current modes.
C NP Number of patch modes to process.

C

C*****

PROGRAM INVERTER

DOUBLE COMPLEX Z(1:12,1:24),V(1:12),TEMP(1:12),RESULT,

+YE,DV

INTEGER INDEX1,INDEX2,M,N,NA,J,K,NP

C*****

C Read voltage and patch reactance terms.

C*****

OPEN (UNIT=87,FILE='IMP.OUT',STATUS='OLD')

OPEN (UNIT=98,FILE='V.OUT',STATUS='OLD')

REWIND (87)

REWIND (98)

READ *, NA,NP

DO 11 M=1,NP

READ (98,*) INDEX1,V(M)

11 CONTINUE

DO 10 M=1,NA

DO 20 N=1,NA

READ (87,*) INDEX1,INDEX2,Z(M,N)

IF ((M.GT.NP).OR.(N.GT.NP)) THEN

Z(M,N)=DBLE(0.0)

ENDIF

20 CONTINUE

10 CONTINUE

C*****

C Set up for matrix inversion.

C*****

DO 30 M=1,NP


```

      Z(M,M+NP)=CMPLX(1.0,0.0)
30  CONTINUE
      DO 31 N=2*NP,1,-1
          Z(1,N)=Z(1,N)/Z(1,1)
31  CONTINUE
      DO 60 I=1,NP-1
          DO 70 M=I+1,NP
              DO 80 N=2*NP,I,-1
                  Z(M,N)=(Z(I,I)*Z(M,N)-Z(M,I)*Z(I,N))
80  CONTINUE
70  CONTINUE
          DO 81 N=2*NP,I+1,-1
              Z(I+1,N)=Z(I+1,N)/Z(I+1,I+1)
81  CONTINUE
60  CONTINUE
      DO 160 I=NP,2,-1
          DO 170 M=I-1,1,-1
              DO 180 N=2*NP,I,-1
                  Z(M,N)=(Z(I,I)*Z(M,N)-Z(M,I)*Z(I,N))
180 CONTINUE
170 CONTINUE
160 CONTINUE
      DO 190 J=1,NP
          DO 200 K=(NP+1),2*NP
              Z(J,K-NP)=Z(J,K)
200 CONTINUE
190 CONTINUE
      DO 191 J=1,NP
          TEMP(J)=CMPLX(0.0,0.0)
191 CONTINUE
      DO 290 M=1,NP
          DO 300 N=1,NP
              TEMP(M)=TEMP(M)+Z(M,N)*V(N)
300 CONTINUE
          PRINT *,TEMP(M)
290 CONTINUE
      RESULT=CMPLX(0.0,0.0)
      DO 301 N=1,NP
          RESULT=RESULT+TEMP(N)*V(N)
301 CONTINUE
      PRINT *,RESULT
      CLOSE(87)
      CLOSE(98)

```

```

C*****
C   Output the admittance term in the slot due to the patch.
C*****

```

```

      OPEN (UNIT=99,FILE='PATAD.OUT',STATUS='UNKNOWN')
      WRITE (99,*) RESULT
      CLOSE(99)

```

```

C*****
C   Read in the remaining calculated terms and find the series
C   impedance.
C*****

```

```

      OPEN (UNIT=1009,FILE='YE.OUT',STATUS='OLD')
      READ (1009,*) YE
      CLOSE (1009)

```

```
OPEN (UNIT=1009,FILE='DV.OUT',STATUS='OLD')  
READ (1009,*) DV  
CLOSE (1009)
```

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```
WRITE (*,*) 50*DV**2/(RESULT+2*YE)
```

```
STOP
```

```
39  FORMAT(' ',2G14.5)
```

```
51  FORMAT(' ',2F6.2)
```

```
END
```

Volt. for

Program Function -

External Functions

Program Variables

```

C      BETAO -      Pole value of beta.
C      LP -        Length of the patch.
C      WP -        Width of the patch.
C      CTn -       Counting variables.
C      NA -        Number of basis modes on patch.
C      UPPER -     number of integration cells.
C
C*****

```

144

PROGRAM VOLTAGE

EXTERNAL TM,TMPRIME,FU,FS,FP,GEMXY,TE

DOUBLE COMPLEX V(12),GEMXY,K1,K2,TE,TM,TMPRIME,J

DOUBLE PRECISION D,ER,ERE,FREQ,LP,WP,LAP,WAP,BETA0,PI,K0,XR(30)
 1,WT(60),AWT,BTWT,ALP(60),BT(30),FUKY,FPKY,FP,FS,FU,KEFF,KX,KY,C
 2,ETA,WF,ZM,EPSEFF

INTEGER NA,N,CT0,CT1,CT2,UPPER

```

C*****
C      Define standard constants.
C*****

```

J=DCMPLX(0.0,1.0)
 PI=DACOS(-1.0)
 C=1/DSQRT(PI*35.416E-19)
 ETA=C*PI*4E-07

```

C*****
C      Read the geometry of the problem.
C*****

```

```

OPEN (UNIT=201,FILE='GEO.DAT',STATUS='OLD')
  READ (201,*) LP,WP
  WRITE (*,*) ' LENGTH OF THE PATCH      - ', LP
  WRITE (*,*) ' WIDTH OF THE PATCH       - ', WP
  READ (201,*) LAP,WAP
  WRITE (*,*) ' LENGTH OF THE APERTURE   - ', LAP
  WRITE (*,*) ' WIDTH OF THE APERTURE    - ', WAP
  READ (201,*) WF,ZM
  READ (201,*) D,ER,ERE,EPSEFF
  WRITE (*,*) ' SUBSTRATE THICKNESS      - ', D
  WRITE (*,*) ' DIELECTRIC CONSTANT      - ', ER
  WRITE (*,*) ' SLOT CONSTANT            - ', ERE
  READ (201,*) FREQ
  WRITE (*,*) ' CENTRE FREQUENCY [GHZ]   - ', FREQ
  FREQ=FREQ*1E09
  READ (201,*) NA
  WRITE (*,*) ' NUMBER OF CURRENT MODES - ', NA
  READ (201,*) UPPER
CLOSE(201)

```

```

OPEN (UNIT=202,FILE='NEWT.DAT',STATUS='OLD')
  READ (202,*) BETAO
  WRITE (*,*) ' BETA AT TMO POLE        - ', BETAO
CLOSE(202)

```

```

C*****

```

C Define the wavenumber for free space.

C*****

KO = 2*PI*FREQ/C
KEFF = KO*DSQRT(ERE)

145

DO 20 N=1,NA
V(N)=DCMPLX(0.0)

20 CONTINUE

C*****

C Read in the roots and weighting coefficients for the gaussian
C quadrature integration.

C*****

OPEN (UNIT=203,FILE='XR30.DAT',STATUS='OLD')
OPEN (UNIT=204,FILE='WGT30.DAT',STATUS='OLD')
DO 30 CTO=30,1,-1
READ (203,*) XR(CTO)
READ (204,*) WT(CTO)

30 CONTINUE
CLOSE(204)
CLOSE(203)

AWT=PI/8.0
DO 40 CTO=1,30
ALP(CTO) = AWT*XR(CTO) + PI/8.0
ALP(CTO+30)= ALP(CTO) + PI/4.0
WT(CTO+30) = WT(CTO)

40 CONTINUE

C*****

C Start the main integration loop.

C*****

OPEN(UNIT=205,FILE='VOLT.OUT',STATUS='UNKNOWN')

DO 50 CTO=1,UPPER

C*****

C Define the points of integration.

C*****

SELECT CASE(CTO)
CASE(1)
BTWT=0.999*BETA0/2.0
DO 51 CT1=1,30
BT(CT1)=BTWT*XR(CT1)+0.999*BETA0/2.0
51 CONTINUE
CASE(2)
BTWT=0.999*BETA0/2.0
DO 52 CT1=1,30
BT(CT1)=BTWT*XR(CT1)+3.001*BETA0/2.0
52 CONTINUE
CASE(3:)
BTWT=2*BETA0
DO 53 CT1=1,30
BT(CT1)=BTWT*XR(CT1)+(4*CTO-8)*BETA0
53 CONTINUE

```

END SELECT
DO 60 CT1=1,60
DO 70 CT2=1,30
  KX=BT(CT2)*DCOS(ALP(CT1))
  KY=BT(CT2)*DSIN(ALP(CT1))
  K1=DCONJG(CDSQRT(DCMPLX(ER*KO**2-BT(CT2)**2)))
  K2=DCONJG(CDSQRT(DCMPLX(KO**2-BT(CT2)**2)))

  FPKY=FP(KEFF,KY,LAP/2.0)
  FUKY =FU(KY,WP)*FU(KX,WAP)
  DO 90 N=1,NA
    V(N)=V(N)+AWT*BTWT*WT(CT1)*WT(CT2)*BT(CT2)*
1      GEMXY(BT(CT2),KX,KO,K1,K2,D,ER,TM)*
2      FPKY*FUKY*FS((2*N-1),KX,LP/2.0)
90    CONTINUE

70    CONTINUE
60    CONTINUE
  WRITE (*,999) V(1),V(NA)
50    CONTINUE
999  FORMAT(' ',4F13.5)

C*****
C  Integration routine for determining the contribution of the
C  surface wave pole.
C*****

DO 100 CT1=1,60
  KX=BETA0*DCOS(ALP(CT1))
  KY=BETA0*DSIN(ALP(CT1))
  K1=DCONJG(CDSQRT(DCMPLX(ER*KO**2-BETA0**2)))
  K2=DCONJG(CDSQRT(DCMPLX(KO**2-BETA0**2)))
  FPKY=FP(KEFF,KY,LAP/2.0)
  FUKY =FU(KY,WP)*FU(KX,WAP)
  DO 120 N=1,NA
    V(N)=V(N)-J*PI*AWT*WT(CT1)*BETA0*
1      GEMXY(BETA0,KX,KO,K1,K2,D,ER,TMPRIME)*
2      FPKY*FUKY*FS((2*N-1),KX,LP/2.0)
120  CONTINUE
100  CONTINUE

C*****
C  Complete the final multiplication and output result to VOLT.OUT
C*****

DO 140 N=1,NA
  V(N)=V(N)/(PI**2)
140  CONTINUE

DO 160 N=1,NA
  WRITE (205,*) N,V(N)
  WRITE ( *,*) N,V(N)
160  CONTINUE

CLOSE(205)
STOP
END

C*****
C  GEMXY - Spectral Domain Green's function relating the

```

```

C          magnetic current along Y of the slot to the
C          electric field along X over the patch.
C*****

```

147

```

DOUBLE COMPLEX FUNCTION GEMXY(BETA,KX,KO,K1,K2,D,ER,TM)
DOUBLE COMPLEX K1,K2,TE,TM,J,PROD
DOUBLE PRECISION KO,D,ER,PI,KX,BETA
PI=DACOS(-1.0)
PROD=K1*D
J=CMPLX(0.0,1.0)
GEMXY=(-1.0/TE(D,K1,K2))*
+((J*KX**2*(ER-1.0)*CDSIN(PROD)/TM(D,K1,K2,BETA,ER))-K1)
RETURN
END

```

```

C*****
C      FP -      Fourier Tranform of a Piece Wise Sinusoidal
C               current source.
C*****

```

```

DOUBLE PRECISION FUNCTION FP(KE,KY,H)
DOUBLE PRECISION KE,KY,H,PI
PI=DACOS(-1.0)
FP=2*KE*(DCOS(KY*H)-DCOS(KE*H))/((KE**2-KY**2)*DSIN(KE*H))
RETURN
END

```

```

C*****
C      FS-      Fourier Tranform of a Sinusoidal current source.
C*****

```

```

DOUBLE PRECISION FUNCTION FS(MODE,KX,A)
DOUBLE PRECISION KX,A,PI
INTEGER MODE
PI=DACOS(-1.0)
FS=MODE*PI*DCOS(KX*A)/(A*((MODE*PI/(2.0*A))**2-KX**2))
RETURN
END

```

```

C*****
C      FU -      Fourier Tranform of a uniform current source.
C*****

```

```

DOUBLE PRECISION FUNCTION FU(KY,W)
DOUBLE PRECISION KY,W
FU=DSIN(KY*W/2.0)/(KY*W/2.0)
RETURN
END

```

```

C*****
C      TE -      This function evaluates the expression for Te as a
C               function of the integration variable Beta
C*****

```

```

DOUBLE COMPLEX FUNCTION TE(D,K1,K2)
DOUBLE COMPLEX K1,K2
DOUBLE PRECISION D
TE=K1*CDCOS(K1*D)+DCMPLX(0.0,1.0)*K2*CDSIN(K1*D)
RETURN
END

```

```

C*****
C      TM -      This function evaluates the expression for Tm as a
C                function of the integration variable Beta
C*****

```

148

```

      DOUBLE COMPLEX FUNCTION TM(D,K1,K2,BETA,ER)
      DOUBLE COMPLEX   K1,K2
      DOUBLE PRECISION D,BETA,ER
      TM=ER*K2*CDCOS(K1*D)+DCMPLX(0.0,1.0)*K1*CDSIN(K1*D)
      RETURN
      END

```

```

C*****
C      TMPRIME - This function is the analytical derivative of Tm
C                and it is evaluated at Beta
C*****

```

```

      DOUBLE COMPLEX FUNCTION TMPRIME(D,K1,K2,BETA,ER)
      DOUBLE COMPLEX   K1,K2
      DOUBLE PRECISION D,BETA,ER
      TMPRIME=-BETA*((ER*CDCOS(K1*D)/K2-ER*K2*D*CDSIN(K1*D)/K1)+
+      DCMPLX(0.0,1.0)*(CDSIN(K1*D)/K1+D*CDCOS(K1*D)))
      RETURN
      END

```


C*****

C
C Yad.for

149

C Program Function -

C This program is responsible for calculating the self reaction
C terms between the slot magnetic current and the fields in the
C slot.

C External Functions

C TE - This function evaluates the expression for Te as a
C function of the integration variable Beta

C TM - This function evaluates the expression for Tm as a
C function of the integration variable Beta

C TMPRIME - This function is the analytical derivative of Tm
C and it is evaluated at Beta

C FU - Fourier Transform of a uniform current source.

C FP - Fourier Transform of a Piece Wise Sinusoidal current
C source.

C GHMY - Spectral Domain Green's function relating the
C magnetic current along Y of the slot to the
C magnetic field along Y in the slot.

C Program Variables

C YE - Mutual reactances of the slot current.

C K1 , K2 - Dependent variables. Function of BETA, K0, and ER.

C J - Square root of -1.

C D - Thickness of the substrate.

C ER - Relative dielectric constant of the substrate material.

C ERE - Assumed dielectric constant of the slot. Average of
C the effective dielectric constant of the patch and the
C feedline.

C EPSEFF - Effective dielectric constant of the feedline.

C FREQ - Holds the frequency of operation.

C LAP - Length of the aperture.

C WAP - Width of the aperture.

C PI - 3.1416

C K0 - Free space wave number.

C FUKY - Temporary variable. Holds the result of one calculation
C of function FU.

C FPKY - Temporary variable. Holds the result of one calculation
C of function FP.

C KEFF - Slot wave number.

C KX - Secondary integration variable.

C KY - Secondary integration variable.

C ZM - Impedance of the microstrip feedline.

C C - Speed of light in free space.

C ETA - Intrinsic impedance of free space.

C WF - Width of the microstrip feed line.

C XR - Root locations in Gaussian Quadrature integration.

C WT - Weighting constants for Gaussian Integration.

C ALP - Primary integration variable.

C AWT - ALP weighting.

C BT - Primary integration variable.

C BTWT - BT weighing.

C BETA0 - Pole value of beta.

C LP - Length of the patch.

C WP - Width of the patch.
 C CTn - Counting variables.
 C NA - Number of basis modes on patch.
 C UPPER - number of integration cells.

150

C*****

PROGRAM APCON

EXTERNAL TM, TMPRIME, FU, FP, TE, GHMY

DOUBLE COMPLEX YF, GHMY, K1, K2, TE, TM, TMPRIME, J

DOUBLE PRECISION D, ER, ERE, FREQ, WAP, LAP, BETA0, PI, KO, FUKX, FPKY,
 +FP, FU, KEFF, XR(30), WT(60), KX, KY, C, ETA, ALP(60), BT(30), BTWT, AWT,
 +EPSEFF, LP, WP, WF

INTEGER CTO, CT1, CT2, UPPER, NA

C*****

C Define standard constants.

C*****

J=DCMPLX(0.0,1.0)
 PI=DACOS(-1.0)
 C=1/DSQRT(PI*35.416E-19)
 ETA=C*PI*4E-07

C*****

C Read the geometry of the problem.

C*****

OPEN (UNIT=201, FILE='GEO.DAT', STATUS='OLD')
 READ (201,*) LP, WP
 READ (201,*) LAP, WAP
 WRITE (*,*) ' LENGTH OF THE APERTURE - ', LAP
 WRITE (*,*) ' WIDTH OF THE APERTURE - ', WAP
 READ (201,*) WF, ZM
 READ (201,*) D, ER, ERE, EPSEFF
 WRITE (*,*) ' SUBSTRATE THICKNESS - ', D
 WRITE (*,*) ' DIELECTRIC CONSTANT - ', ER
 WRITE (*,*) ' SLOT CONSTANT - ', ERE
 READ (201,*) FREQ
 WRITE (*,*) ' CENTRE FREQUENCY [GHZ] - ', FREQ
 FREQ=FREQ*1E09
 READ (201,*) NA
 READ (201,*) UPPER
 CLOSE(201)

OPEN (UNIT=202, FILE='NEW.DAT', STATUS='OLD')
 READ (202,*) BETA0
 WRITE (*,*) ' BETA AT TMO POLE - ', BETA0
 CLOSE(202)

C*****

C Define the wavenumber for free space and the slot.

C*****

KO = 2*PI*FREQ/C
 KEFF = KO*DSQRT(ERE)

YE=DCMPLX(0.0)

```
C*****
C   Read in the roots and weighting coefficients for the gaussian      151
C   quadrature integration.
C*****

OPEN (UNIT=203,FILE='XR30.DAT',STATUS='OLD')
OPEN (UNIT=204,FILE='WGT30.DAT',STATUS='OLD')
DO 30 CT0=1,30
  READ (203,*) XR(CT0)
  READ (204,*) WT(CT0)
30  CONTINUE
CLOSE(204)
CLOSE(203)

AWT=PI/8.0
DO 40 CT0=1,30
  ALP(CT0) = AWT*XR(CT0) + PI/8.0
  ALP(CT0+30)= ALP(CT0) + PI/4.0
  WT(CT0+30) = WT(CT0)
40  CONTINUE

C*****
C   Start the main integration loop.
C*****

OPEN(UNIT=205,FILE='YE.OUT',STATUS='UNKNOWN')

DO 50 CT0=1,UPPER
C*****
C   Define the points of integration.
C*****

  SELECT CASE(CT0)
  CASE(1)
    BTWT=0.999*BETA0/2.0
    DO 51 CT1=1,30
      BT(CT1)=BTWT*XR(CT1)+0.999*BETA0/2.0
51    CONTINUE
  CASE(2)
    BTWT=0.999*BETA0/2.0
    DO 52 CT1=1,30
      BT(CT1)=BTWT*XR(CT1)+3.001*BETA0/2.0
52    CONTINUE
  CASE(3:)
    BTWT=2*BETA0
    DO 53 CT1=1,30
      BT(CT1)=BTWT*XR(CT1)+(4*CT0-8)*BETA0
53    CONTINUE
  END SELECT
DO 60 CT1=1,60
DO 70 CT2=1,30
  KX = BT(CT2)*DCOS(ALP(CT1))
  KY = BT(CT2)*DSIN(ALP(CT1))
  K1 = DCONJG(CDSQRT(DCMPLX(ER*K0**2-BT(CT2)**2)))
  K2 = DCONJG(CDSQRT(DCMPLX( K0**2-BT(CT2)**2)))
  FPKY = FP(KEFF,KY,LAP/2.0)**2
  FUKX = FU(KX,WAP)**2
  YE=YE+AWT*BTWT*WT(CT1)*WT(CT2)*BT(CT2)*
```

```

1      GHMYY(BT(CT2),KY,K0,K1,K2,D,ER,TM)*
2      FPKY*FUKX
70     CONTINUE
60     CONTINUE
      WRITE (*,999) YE
50     CONTINUE
999    FORMAT(' ',2F13.5)

```

152

```

C*****
C      Integration routine for determining the contribution of the
C      surface wave pole.
C*****

```

```

      DO 100 CT1=1,60
      KX=BETA0*DCOS(ALP(CT1))
      KY=BETA0*DSIN(ALP(CT1))
      K1=DCONJG(CDSQRT(DCMPLX(ER*K0**2-BETA0**2)))
      K2=DCONJG(CDSQRT(DCMPLX(  K0**2-BETA0**2)))
      FPKY = FP(KEFF,KY,LAP/2.0)**2
      FUKX = FU(KX,WAP)**2
      YE=YE-J*PI*AWT*WT(CT1)*BETA0*
1      GHMYY(BETA0,KY,K0,K1,K2,D,ER,TMPRIME)*
2      FPKY*FUKX
100    CONTINUE

```

```

C*****
C      Complete the final multiplication and output result to VOLT.OUT
C*****

```

```

      YE=-YE*J/(K0*ETA*PI**2)

```

```

      WRITE (205,*) YE
      WRITE ( *,*) YE

```

```

      CLOSE(205)
      STOP
      END

```

```

C*****
C      GEMKY - Spectral Domain Green's function relating the
C      magnetic current along Y of the slot to the
C      magnetic field along Y in the slot.
C*****

```

```

      DOUBLE COMPLEX FUNCTION GHMYY(BETA,KY,K0,K1,K2,D,ER,TM)
      DOUBLE COMPLEX  K1,K2,TE,TM,J,PROD
      DOUBLE PRECISION K0,D,ER,PI,KY,BETA
      PI=DACOS(-1.0)
      PROD=K1*D
      J=CMPLX(0.0,1.0)
      GHMYY=(J*(K1*CDCOS(PROD)+J*K2*ER*CDSIN(PROD))*
+(ER*K0**2-KY**2)/(K1*TM(D,K1,K2,BETA,ER))-
+(J*KY**2*K1*(ER-1.0)/TM(D,K1,K2,BETA,ER))/TE(D,K1,K2))
      RETURN
      END

```

```

C*****
C      FP -      Fourier Transform of a Piece Wise Sinusoidal
C      current source.
C*****

```

```

DOUBLE PRECISION FUNCTION FP(KE,KY,H)
DOUBLE PRECISION KE,KY,H,PI
PI=DACOS(-1.0)
FP=2*KE*(DCOS(KY*H)-DCOS(KE*H))/((KE**2-KY**2)*DSIN(KE*H))
RETURN
END

```

153

```

C*****
C      FU -      Fourier Tranform of a uniform current source.
C*****

```

```

DOUBLE PRECISION FUNCTION FU(KY,W)
DOUBLE PRECISION KY,W
FU=DSIN(KY*W/2.0)/(KY*W/2.0)
RETURN
END

```

```

C*****
C      TE -      This function evaluates the expression for Te as a
C                function of the integration variable Beta
C*****

```

```

DOUBLE COMPLEX FUNCTION TE(D,K1,K2)
DOUBLE COMPLEX K1,K2
DOUBLE PRECISION D
TE=K1*CDCOS(K1*D)+DCMPLX(0.0,1.0)*K2*CDSIN(K1*D)
RETURN
END

```

```

C*****
C      TM -      This function evaluates the expression for Tm as a
C                function of the integration variable Beta
C*****

```

```

DOUBLE COMPLEX FUNCTION TM(D,K1,K2,BETA,ER)
DOUBLE COMPLEX K1,K2
DOUBLE PRECISION D,BETA,ER
TM=ER*K2*CDCOS(K1*D)+DCMPLX(0.0,1.0)*K1*CDSIN(K1*D)
RETURN
END

```

```

C*****
C      TMPRIME - This function is the analytical derivative of Tm
C                and it is evaluated at Beta
C*****

```

```

DOUBLE COMPLEX FUNCTION TMPRIME(D,K1,K2,BETA,ER)
DOUBLE COMPLEX K1,K2
DOUBLE PRECISION D,BETA,ER
TMPRIME=-BETA*((ER*CDCOS(K1*D)/K2-ER*K2*D*CDSIN(K1*D)/K1)+
+ DCMPLX(0.0,1.0)*(CDSIN(K1*D)/K1+D*CDCOS(K1*D)))
RETURN
END

```

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END

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FIN