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NOVEL MICROSTRIP ANTENNAS

by

Aldo Petosa, B.Eng.

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

Ottawa-Carleton Institute for Electrical Engineering
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"NOVEL MICROSTRIP ANTENNAS"

submitted by Aldo Petosa, B.Eng.
in partial fulfilment of the requirements for
the degree of Master of Engineering

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August, 1991

ABSTRACT

Novel microstrip patch antennas have been investigated for use in mobile cellular communications and mobile satellite communications. Techniques for reducing the resonant frequency of a centre-fed circular microstrip cavity have been investigated. Interlaced circumferential slots were added, resulting in reductions in the resonant frequency of up to 50%. The performance of these antennas when mounted on the roof of an automobile has been analyzed using the Geometric Theory of Diffraction. Tchebychev impedance matching networks have been investigated to improve the bandwidth of these modified antennas.

A cylindrical array of circular polarized microstrip patches has been designed and constructed for use in mobile satellite communications. The array geometry consisted of 18 L-band circularly-polarized square microstrip elements which were modified to reduce the size of the resonant patch. A high-gain beam was formed which was electronically steered around the array.

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LIST OF PRINCIPAL SYMBOLS

SYMBOL	DESCRIPTION
a	Radius of Microstrip Disk Antenna
a_e	Effective Radius of Disk Antenna
$A(s', s)$	Spreading Factor at a Distance s from diffraction point Q , due to a source at a distance s' from Q
$(AF)_n$	Normalized Array Factor
AP	Antenna Pattern
c	Speed of Light in Free Space (3.0×10^8 m/s)
d	Distance
D	Diffraction Coefficient
D_h	Hard Edge Diffraction Coefficient
D^i	Incident Edge Diffraction Coefficient
D^r	Reflected Edge Diffraction Coefficient
$E^d(s)$	Diffacted Field at Distance s from Diffraction Point
E_{GO}	Geometric Optics Field
$E_i(Q)$	Incident Field at Diffraction Point Q
E_θ, E_ϕ	Field Intensities
$F(x)$	Fresnel Integral
F_e	Figure of Merit for Cellular Communications
f_r	Resonant Frequency
h	Substrate Height

$I(n)$	Current Amplitude of n^{th} Array Element
$J_n(x)$	n^{th} Mode of the Bessel Function of the First Kind
k_{mn}	m^{th} Zero of the First Derivative of order n Bessel Function of the First Kind
k_o	Free Space Wave Number
N	Number of Array Elements
$P(\theta, \phi)$	Radiation Intensity
r	Far-Field Distance
V_a	Voltage at the Edge of the Microstrip Disk Antenna
Z_{in}	Input Impedance
Z_o	Characteristic Impedance of Microstrip Line
Z_{out}	Output Impedance
β	Additional Phase Shift Between Array Elements
ϵ_r	Relative Permittivity
(θ, ϕ)	Polar Co-ordinates
λ_o	Free Space Wavelength
Ψ	Array Factor Phase Term

1. INTRODUCTION

1.1 Introduction

In the past several years, microstrip antennas have become widely used in a variety of areas such as cellular and satellite communications, airborne radar and microwave landing systems to name but a few. This popularity can be attributed to the number of attractive advantages microstrip antennas offer: they are light-weight, low-profile antennas which can be made conformal to many surfaces. Their manufacturing is relatively simple and they can be easily integrated with other printed circuits making them inherently low-cost antennas [1]. There are however, certain disadvantages with this technology: microstrip antennas have low efficiency and small bandwidth, and problems may arise from extraneous radiation from feed lines, junctions and surface waves. Furthermore, there are no rigorous design equations that exist in closed form for these antennas, and much research has been carried out and is still under way to develop models or apply numerical techniques to predict their performance (example [2] - [10]).

The work in this thesis deals with novel designs of microstrip elements for use as a mobile vehicle antenna for cellular communications and in a innovative array configuration for mobile satellite communications.

1.2 Objectives

This thesis investigates novel microstrip antenna designs for applications in mobile cellular and mobile satellite communications. Circular microstrip antennas are modified to improve their performance with respect to such parameters as size, radiation patterns and bandwidth such that they are better suited for applications as roof-top mounted vehicle antennas for mobile cellular communications. The effects of vehicle roof-tops (which represent finite ground planes to the microstrip antennas) on the radiation patterns of the microstrip antennas are investigated using the Geometric Theory of Diffraction.

Modifications are also made to square microstrip radiating elements to reduce their resonant length for a given frequency. These modified elements are used in a unique cylindrical array for mobile satellite communications.

1.3 Organization

The thesis is divided into two major sections. The first focuses on circular disk elements which have been modified to serve as roof-top mounted vehicle antennas for cellular communications. In Chapter 2, several modifications are investigated. These antennas are constructed and measured and their performance is compared to a commercially available cellular vehicle antenna. Chapter 3 discusses the improvement in bandwidth which may be achieved by the addition of impedance matching networks to the modified antennas. Chapter 4 deals with the

analysis of the effects of finite ground planes on the radiation patterns of the modified microstrip antennas. The Geometric Theory of Diffraction is used to predict these effects and the theoretical predictions are compared to measured results.

The second section (Chapter 5) details the investigation of a modified circular-polarized square microstrip element for use in an eighteen-element cylindrical mobile array for satellite communications. The cylindrical array produces a high-gain beam which is electronically steered to track the satellite. The array has been designed, constructed and tested, and the measured results are compared to theoretical predictions.

The final chapter summarizes the work done in this thesis and draws conclusions and recommendations for future work.

2. CELLULAR VEHICLE ANTENNA

2.1 Introduction

Mobile cellular communications is a rapidly expanding area of research fuelled by increased market demands due to the popularity and convenience of mobile phones for business and personal use. Mobile cellular communications involve the transmission and reception of signals between a vehicle and a nearby base station, one of many stations interspersed throughout the city in regions known as cells (hence the name cellular communications). The base station antennas operate at frequencies between 800 MHz and 900 MHz and are vertically polarized. Thus one of the requirements for a mobile cellular antenna is vertical polarization. (Note the simplifying assumption is made that the signals arriving at the mobile antenna remain vertically polarized, which is almost never totally the case since reflections off buildings and other city obstacles tend to alter the polarization of the signals.) It is necessary for the vehicle antenna to generate an omnidirectional radiation pattern in the azimuth plane with the peak of the beam directed close to the horizon. Such a pattern is produced, for example, by a dipole whose length is equal to or less than one wavelength. Figure 2.1 shows the normalized elevation plane patterns of various length dipoles. A figure of merit applicable for a mobile antenna is the directivity into a volume enclosed by the spherical polar coordinates $60^\circ \leq \theta \leq 90^\circ$, and $0^\circ \leq \phi \leq 360^\circ$ [11]. This is expressed mathematically as:

Figure 2.1 Normalized Elevation Pattern of Dipoles in Free Space, With Accompanying Figure of Merit F_0 .

(Eq. 2.1)

$$F_{\theta} = \frac{\int_0^{2\pi} \int_{\pi/3}^{\pi/2} P(\theta, \phi) \sin\theta d\theta d\phi}{\int_0^{2\pi} \int_0^{\pi/2} P(\theta, \phi) \sin\theta d\theta d\phi}$$

Thus, one wishes to maximize the performance of the vehicle antenna by concentrating the radiation pattern within the envelope specified above. For the case of a dipole, this is achieved by increasing the length to one wavelength as depicted in Figure 2.1 where the figure of merit increases from 0.696 for a quarter wavelength dipole to 0.891 for a full wavelength dipole. For microstrip antennas, the same effect can be achieved by other means, which will be discussed later in Section 2.4. Attention is now focused on the performance of a commercially available vehicle antenna.

2.2 Commercial Vehicle Antenna

A typical commercial antenna which was investigated in this work consists of a wire antenna in series with an inductive coil as shown in Figure 2.2. An adhesive material is applied to the base of the antenna allowing it to be attached to the outside of one of the car windows (usually the rear). The capacitive coupler is then attached on the opposite side of the window and the received signal is coupled into a coaxial line to the handset. It is readily apparent that such an installation is far from ideal. Since the antenna is located at the rear of the automobile, there will be interference from the entire

Figure 2.2 Sketch of Commercially Available Mobile Cellular Antenna.

front body of the vehicle, resulting in a degradation of the omnidirectional radiation pattern. Far more important however, the monopole antenna is not attached to any ground plane, thus its radiation pattern will strongly depend on the exact mounting location and will not be ideal. The normalized radiation pattern in the elevation plane of the commercial antenna in the absence of a ground plane was measured and is shown in Figure 2.3. As expected, the beam peak is directed towards the horizon. A pattern cut in the azimuth plane was also taken, revealing a nearly omnidirectional pattern, with ± 2.25 dB variation in level. These patterns are deceiving for they do not take into account the effects of ground planes which would be present in the forms of the vehicle roof-top or body. The antenna was therefore mounted on a square ground plane (50 cm x 50 cm) and the radiation patterns were measured. Figure 2.4 depicts the deterioration in performance when the ground is added. More power is diverted out of the region of interest (away from the horizon) towards the broadside region of the ground plane. This degradation would increase in the presence of a larger ground plane such as that of a vehicle roof-top. The effects of finite ground planes will be discussed in greater detail in the chapter on Geometric Theory of Diffraction. The rest of this chapter focuses on microstrip antennas and how they may be modified to provide adequate performance as a cellular vehicle antenna.

Figure 2.3 Normalized Elevation Pattern of Commercial Antenna at 850 MHz in the Absence of a Ground Plane.

Figure 2.4 Normalized Elevation Pattern of Commercial Antenna at 850 MHz Centred on 50cm x 50cm Ground Plane.

2.3 Microstrip Antennas

The use of a microstrip antenna seems a reasonable alternative to the commercial vehicle antenna for many of the reasons that have already been listed, such as their light weight, low profile and low cost. A roof-mounted microstrip antenna would be nearly conformal to the car surface and the risk of accidental damage would be greatly decreased compared to the exposed and susceptible commercial antenna. The reason why a microstrip antenna has not previously proved viable is its size. It will be shown that the size of the microstrip disk antenna examined is a function of the frequency of operation, the substrate permittivity and the mode in which the antenna is excited. To produce a radiation pattern appropriate for mobile cellular applications, the disk must operate at a higher order mode which results in an increase in size for a given frequency. At 850 MHz, the radius of a centre-fed disk is in the order of 20 cm divided by the square root of the substrate permittivity. Thus it would seem reasonable to use a substrate with high permittivity in order to decrease the dimensions of the antenna. Unfortunately, the bandwidth of microstrip antennas is inversely proportional to the substrate permittivity: the higher the permittivity, the lower the bandwidth. Thus a low-permittivity substrate should be used to provide adequate bandwidth, but the size of the disk then becomes unrealistic for use as a vehicle antenna. A prime focus of this work was on techniques to reduce the size of the antenna without sacrificing bandwidth. The approach taken was to begin with a standard patch design which produced the appropriate radiation

pattern and then to modify this antenna in an attempt to reduce its size. The standard antenna is the centre-fed circular disk. Its characteristics will first be reviewed in the next section before proceeding to the modified designs.

2.4 The Centre-fed Circular Disk

Figure 2.5 (a) shows a typical microstrip disk. It consists of a circular copper disk (or patch) supported by a grounded dielectric substrate. The disk can be fed via a coaxial feed (as shown in the diagram) or by a microstrip line on the same plane as the patch. There are several models which have been developed to predict the performance of the microstrip disk [12]. The simplest and most widely used model - the cavity model - is described here and is used in this work.

The microstrip disk can be thought of as a resonant cavity as shown in Figure 2.5 (b). Electric walls bound the top and bottom surfaces, while a magnetic wall bounds the rest of the cavity. By solving the wave equation with the boundary conditions of the cavity model, the resonant frequency of the disk can be determined [12] and is given by [13,14]:

$$f_r = \frac{k_{mn}c}{2\pi a\sqrt{\epsilon_r}} \quad (\text{Eq. 2.2})$$

Figure 2.5 Microstrip Disk Antenna.
(a) Top and Side Views
(b) Cavity Model

where

$$a_e = \sqrt{1 + \frac{2h}{\pi a \epsilon_r} \ln\left(\frac{\pi a}{2h}\right) + 1.7726} \quad (\text{Eq. 2.3})$$

The m and n subscripts correspond to the TM_{mn} modes of the resonant cavity, where n represents the azimuthal variation in the magnetic field while m represents the radial variation. The far-field radiation pattern is derived by using image theory to replace the ground plane by an equivalent magnetic current. Then the vector electric potential can be calculated by integrating this magnetic current over the aperture [12]. The resulting radiation patterns are:

$$E_\theta = j^n \frac{V_a k_o}{2} \frac{e^{-jk_o r}}{r} \cos n\phi (J_{n+1} - J_{n-1}) \quad (\text{Eq. 2.4})$$

$$E_\phi = j^n \frac{V_a k_o}{2} \frac{e^{-jk_o r}}{r} \cos\theta \sin n\phi (J_{n+1} + J_{n-1}) \quad (\text{Eq. 2.5})$$

where $J_n = J_n(k_o a \sin \theta)$

The figure of merit for various modes can be determined by calculating the directivity in the region of interest described in Section 2.1. This was done by Vaughan, in [15] for modes n=0 to n=4. The resulting figure of merit curves have been reproduced in Figure 2.6 for convenience. For a disk operating in a given mode, it is apparent that

Figure 2.6 Element Figure of Merit (Percentage Directivity into an area $0^\circ \leq \phi \leq 360^\circ$, $60^\circ \leq \theta \leq 90^\circ$) called Mobile Communications Scenario (MCS) for Pure Mode Circular Microstrip Disks. Different curves are for different azimuthal modes and labelled points are substrate relative permittivity required for cavity model patch antenna. From [15].

by decreasing the radius of the disk (i.e. by increasing the permittivity of the substrate) the figure of merit increases. In other words, as the dielectric permittivity is increased, more radiation energy is directed down towards the horizon, as shown in Figure 2.7. All the modes except the $n=1$ mode (which is the mode most commonly excited for microstrip disks) have nulls in the radiation pattern at broadside. Vaughan [15] goes on to design, construct and measure a two-port microstrip disk excited in the $n = 3$ mode. Although it provides the desired radiation pattern, the radius of the disk was 14.25 cm. (This can be calculated from (Eq. 2.1) using a relative permittivity of 2.5 at a frequency of 890 MHz, and $k_{13} = 4.201$.) A disk of such a large radius is undesirable for a vehicle antenna and methods of reducing the size would be beneficial. This thesis looks at several methods of reducing the size of a centre-fed (i.e. $n=0$ mode) disk. The centre-fed disk was chosen for it requires only one feed and can provide an adequate radiation pattern. A centre-fed circular disk was manufactured with a radius of 6.0 cm on a 1.252 cm thick nylon substrate (ϵ_r of 2.5) truncated to the edge of the patch and mounted on a square ground plane of dimensions 50 cm x 50 cm. The radius chosen was somewhat arbitrary and represents an acceptable size for mounting on the roof-top of a vehicle. The other parameters were chosen based upon material availability and manufacturing costs and considerations and although they may not be optimal in terms of electrical characteristics they are adequate to demonstrate the size-reduction techniques to be discussed.

The radiation pattern and return loss of the centre-fed disk were

Figure 2.7 Effect of Substrate Permittivity on Normalized Elevation Patterns of a Centre-Fed ($n = 0$ mode) Microstrip Disk.

measured and are shown in Figures 2.8 and 2.9 respectively. The resonant frequency of the disk was measured to be 1.85 GHz with a 10 dB bandwidth of 110 MHz (adequate for cellular communications). The radiation pattern extending beyond $\theta = 90^\circ$ arises from the presence of the finite ground plane and will be examined in detail in Chapter 4.

From these patterns, one cannot yet justly compare the radiation pattern of the microstrip disk with that of the commercial antenna on the same ground plane (Figure 2.4). The two patterns were not taken at the same frequency, thus although the ground plane used in each case was physically the same, electrically it appeared much smaller in the case of the commercial antenna (at 850 MHz) than for the microstrip antenna (at 1850 MHz). It will be shown later that the radiation pattern of the microstrip disk is improved by increasing the size of the ground plane (either physically or electrically).

Modifications (both successful and unsuccessful) will now be discussed which reduce the resonant frequency of the centre-fed patch, while attempting to maintain the radiation pattern and bandwidth.

2.5 Modified Disks

Much work has been carried out on modifying circular patches to alter their resonant frequencies, input impedance or radiation patterns. McIlvenna and Kernweis [16] added stubs to the patch to reduce the resonant frequencies of the dominant modes. Unfortunately, for the mode of interest (TM_{10}), as the frequency was lowered, the central null of the radiation pattern began to fill in and it

Figure 2.8 Normalized Elevation Pattern of Centre-Fed Microstrip Disk at 1850 MHz.

Figure 2.9 Return Loss of Centre-Fed Microstrip Disk.

eventually degraded to the broadside beam of the TM_{11} mode which does not produce an appropriate radiation pattern for cellular communications.

Addition of a coaxial stub tuner [17] was shown to produce dual frequency operation. The example cited was that of an unloaded disk with resonance at 2.5 GHz made to resonate at 2.3 GHz and 2.7 GHz using a coaxial stub tuner. Thus a reduction of 8% in frequency was achieved. However a reduction in the vicinity of 50% is required to make microstrip disks of reasonable size for vehicular mounted applications.

Disks of varying size have been stacked on top of one another [18] to increase the bandwidth of operation but once again this would not prove viable for vehicular antennas due to the added height and the fact that large disk radii are still required.

In this work, three techniques are investigated to reduce the resonant frequency of a disk of given size: (1) the inductive loaded disk, (2) the shorted disk and (3) the slotted disk. The first method, although plausible in theory, proved unsuccessful in practise. The second and third methods worked with varying degrees of success. These techniques will now be discussed in detail.

2.5.1 Inductively Loaded Disks

In the context of input impedance and return loss, the centre-fed disk antenna can be thought of as a resonant circuit and can be modelled as a series LRC circuit. At resonance, the reactive components cancel and the resulting input impedance is real. Above resonance,

there is an inductive reactance while below resonance the reactance is capacitive. Thus, in theory, if one wishes to lower the resonant frequency of the disk, one could add series inductance to cancel the reactive component. To investigate this effect, the antenna return loss was modelled as a series LRC circuit using Touchstone. Values for the lumped R, L and C components were chosen in order to emulate the measured return loss of the unloaded centre-fed disk. The resulting circuit and corresponding return loss are shown in Figure 2.10 (a) and Figure 2.11 respectively. To study the effects of inductive loading, a lumped inductor was added in series to the circuit, as shown in Figure 2.10 (b). In order to shift the resonant frequency down to the required range the value of the series inductor was found to be 175 nH. As can be seen from the return loss curve of the loaded antenna in Figure 2.11, the inductor not only reduces the resonant frequency, but has the undesired effect of narrowing the bandwidth. The practical realization of the lumped series inductor turned out to be a problem. The use of a chip inductor was ruled out due to its inability to handle the required power levels. An attempt was made to use a wire-wound ferrite bead. The bead was soldered at one end to the coaxial feed, and at the other end to the patch surface, as shown in Figure 2.12. This configuration proved unsuccessful for two major reasons: it was mechanically unstable and thus unreliable. Of equal or greater importance was the fact that the ferrite bead had a large resistive component at the desired frequency, and thus did not behave as a suitable inductor.

It was then decided to replace the ferrite bead with a spiral

(a)

(b)

Figure 2.10 Circuit Model for Return Loss of Microstrip Disk.
 (a) Unloaded Disk
 (b) Inductively Loaded Disk

Figure 2.11 Modelled Return Loss of Loaded and Unloaded Disk.

Figure 2.12 Microstrip Disk with Ferrite Bead Load.

inductor, similar to the ones used in MMIC applications. The idea was to print the inductor on the patch surface, thus eliminating the mechanical problems posed by the ferrite bead. A spiral inductor was modelled using Touchstone. However, given the desired frequency range and the large substrate thickness of the antenna, a spiral trace printed on the patch surface did not behave as a discrete inductor, but as a transmission line. The theoretical predictions were verified by testing several spiral traces on the antenna patches. It was further found that the traces interfered with the radiation patterns in an undesired manner.

Thus, although in theory, the addition of a lumped inductor reduces the resonant frequency of the centre-fed disk, no practical implementation was found and the idea was abandoned in order to pursue more promising alternatives.

2.5.2 The Shorted Disk

It is known that one can halve the size of a half-wavelength resonant rectangular patch by introducing an electric wall along the centre of the patch. The electric wall satisfies all the boundary conditions of the half wavelength patch, and thus a short-circuited quarter-wavelength patch will resonate at the same frequency as a normal half-wavelength patch. The idea of reducing the size of the patch by introducing an electric wall was extended to the circular disk.

Figure 2.13 shows a diagram of the short-circuited disk antenna.

Figure 2.13 Shorted Microstrip Disk Antenna.

Four brass screws act as the electric wall. The centre feed was extended via four tracks to the edge of the disk to improve the input impedance match. Thus, this disk no longer acts as one antenna, but as an array of four elements, where each element is a quarter section of the disk. To obtain more insight into the performance of such an array, one can visualize the elements as being quarter-wavelength shorted rectangular patches, or half-wavelength patches as shown in Figure 2.14. Referring to this figure, taking a pattern cut in the A-A plane, the disk can be modelled by two quarter-wavelength shorted patches (Figure 2.14b). To predict the approximate pattern, the patches are replaced by isotropic elements separated by a half wavelength (Figure 2.14c) and standard array theory can then be used. From Balanis [19], for an N element linear array, the normalized array factor is given by:

$$(AF)_n = \frac{1}{N} \frac{\sin(\frac{N}{2}\psi)}{\sin(\frac{1}{2}\psi)} \quad (\text{Eq. 2.6})$$

where

$$\psi = k d \sin\theta + \beta \quad (\text{Eq. 2.7})$$

Taking $N = 2$, and with $d = 0.5\lambda$, Eq. 2.6 reduces to:

$$(AF)_n = \cos\left[\frac{1}{2}(\pi\sin(\theta) + \pi)\right] \quad (\text{Eq. 2.8})$$

(a)

(b)

(c)

Figure 2.14 Simple Model for Shorted Disk Antenna Pattern.
 (a) Shorted Disk
 (b) $\lambda/4$ Shorted Patch
 (c) Isotropic Two-Element Array

In the above equation, $\beta = \pi$, since the feeds on the patches are separated by a half wavelength and are thus 180 degrees out of phase. Plotting (Eq. 2.8) produces a radiation pattern with a null in the broadside direction, as desired. Thus it might be expected that a modification of the square element of Figure 2.14 to the circular form of Figure 2.13 might produce a similar radiation pattern. This was indeed the case as can be seen from the measured radiation pattern of Figure 2.15. The shorted antenna was resonant at a frequency of 417 MHz, a reduction to 23% with respect to the original unmodified disk. Unfortunately, both the radiation bandwidth and impedance bandwidth were very narrow. Figure 2.16 compares the return loss of the shorted disk to that of the unmodified disk. Improvements in this narrow return loss can be made by the addition of impedance matching networks, and this will be investigated in Chapter 3. Before proceeding with bandwidth enhancement techniques however, one more method of frequency reduction will be investigated: the slotted disk.

2.5.3 The Slotted Disk

The third method used to reduce the resonant frequency of the disk is based on the premise that increasing the path length travelled by the surface current on the patch will result in a drop in resonant frequency. This idea was realized by cutting concentric slots in the disk surface as shown in Figure 2.17. The surface current leaving the central feed point is forced to flow around the slots, thereby increasing the path length. To determine the effect of the slots on the

Figure 2.15 Normalized Elevation Pattern of Shorted Disk at 430 MHz.

Figure 2.16 Return Loss of Shorted Disk Compared To Unmodified Centre-Fed Disk.

Figure 2.17 Slotted Disk.

resonant frequency of the disk, three antennas were constructed with concentric slots spaced 1.0 cm, 0.5 cm and 0.25 cm apart as shown in Figure 2.18. Theoretically, one would predict that the closer the slots are spaced, the longer the path length and thus the lower the resonant frequency. The resonant frequency of each antenna was measured and, as is shown in Table 2.1, the measurements confirm this prediction.

SLOT SPACING (cm)	RESONANT FREQUENCY (MHz)
Solid Disk	1850
1.0	1525
0.5	1178
0.25	987

Table 2.1 Resonant frequency of solid and slotted disks.

The normalized radiation patterns of each antenna were measured and are shown in Figures 2.19 to 2.21. As can be seen, when the number of slots is increased, the resonant frequency drops and the radiation patterns do not deteriorate to the lower order TM_{11} mode (peak at broadside). Pattern cuts in the azimuth plane revealed omni-directional performance with the worst case power deviation being ± 2 dB.

The effect of the gap between slots was also briefly investigated. Using the disk with 0.5 cm slot spacing (Figure 2.18b), the gap length was decreased from 1.0 cm to 0 cm (i.e. the slots became concentric rings). There was negligible effect on the resonant frequency and impedance as the gap size was decreased. Only when the gap size was zero (and the slots became continuous) did the results

(a)

(b)

$g = 1.0 \text{ cm}$

(c)

Figure 2.18 Fabricated Slotted Disks.
(a) Slots spaced 1.00 cm apart
(b) Slots spaced 0.50 cm apart
(c) Slots spaced 0.25 cm apart

Figure 2.19 Normalized Elevation Pattern of Slotted Disk
(1.0 cm spacing) at 1400 MHz.

Figure 2.20 Normalized Elevation Pattern of Slotted Disk
(0.50 cm spacing) at 1175 MHz.

Figure 2.21 Normalized Elevation Pattern of Slotted Disk
(0.25 cm spacing) at 900 MHz.

change. At this limit, the basic configuration is altered. No longer can the surface current flow from the centre to the edge of the disk. Instead, the geometry becomes one of a small disk encircled by a number of parasitic rings. The resonant frequency thus becomes that of the small disk which is dramatically higher and is not appropriate for use as a vehicle mounted antenna.

This method of adding slots to reduce the resonant frequency of the disk is superior to the shorted disk method. By simply varying the spacing between slots, one can control the resonant frequency of the disk, reducing it by as much as 50%. This is a versatile property which may find a wide variety of applications anywhere where size constraints are present. The only drawback is the reduction in impedance bandwidth which accompanies the reduction in frequency. Figures 2.22 to 2.24 show the return loss of the three slotted disk antennas. It can be seen that the impedance bandwidth is inversely proportional to the number of slots. To help improve the bandwidth, a multi-stage impedance matching network can be used. The next chapter investigates the effects of such a network.

Figure 2.22 Return Loss of Slotted Disk (1.0 cm spacing).

Figure 2.23 Return Loss of Slotted Disk (0.5 cm spacing).

Figure 2.24 Return Loss of Slotted Disk (0.25 cm spacing).

3. IMPEDANCE MATCHING

3.1 Introduction

Both the shorted disk of Section 2.4 and the slotted disks of Section 2.5 had impedance bandwidths that were much narrower than the original unmodified disk. To increase the bandwidth, an impedance matching network can be added to the antenna. In this chapter the effects of a quarter wavelength transformer and of a Tchebychev matching network on the impedance bandwidth of the slotted disk of Figure 2.18 (c) will be investigated.

In order to design a broadband matching network, the input impedance of the antenna must first be known. The input impedance of the slotted disk was measured using an HP 8753 network analyzer and is shown below in Table 3.1.

FREQUENCY (MHz)	INPUT IMPEDANCE (Ω)
935	10.3 - j 32.2
945	10.2 - j 28.0
955	10.2 - j 22.0
965	10.3 - j 16.0
975	10.8 - j 9.2
985	11.4 - j 1.9
987	11.5
995	12.3 + j 6.1
1005	13.8 + j 15.9
1015	16.6 + j 27.8
1025	21.4 + j 42.8
1035	30.0 + j 62.0

Table 3.1 Input impedance versus frequency of slotted disk.

The reflection coefficients at the above frequencies were also measured and they were used to generate a Touchstone data file to simulate the return loss of the slotted disk antenna. The return loss of the unmatched disk is plotted in Figure 3.1, and as can be seen, the bandwidth is very narrow and the return loss is poor. The effects on the return loss of the two impedance matching techniques are now investigated.

3.2 Quarter Wavelength Transformer

The first matching network investigated was the simple quarter wavelength transformer. From transmission line theory, the well-known impedance matching equation can be derived:

$$Z_o = \sqrt{Z_{in} Z_{out}} \quad (\text{Eq. 3.1})$$

At the centre frequency of 987 MHz, the input impedance of the slotted antenna is 11.5Ω and the desired output impedance is 50Ω , thus Z_o is 24Ω . The slotted disk was then modelled in Touchstone with the above quarter-wavelength transformer and the return loss is plotted in Figure 3.2. The improvement over the unmatched antenna is significant, yet the bandwidth is still not as wide as that of the solid disk. To widen the bandwidth even further, a Tchebychev matching network was investigated.

Figure 3.1 Return Loss of Slotted Disk (0.25 cm spacing).

Figure 3.2 Return Loss of Slotted Disk with Quarter Wavelength Impedance Transformer.

3.3 Tchebychev Matching Network

The principles of using low-pass filter prototypes for matching networks are well documented by Matthaei, Young and Jones [20], and the general procedure will be only briefly outlined here. An appropriate impedance matching network is shown in Figure 3.3 and consists of quarter wavelength lines and shunt stubs. To demonstrate the technique, a three-resonator filter was chosen where one of the resonators is the radiating element itself. (More elements can be used to improve bandwidth and reduce in-band ripples, but at the expense of additional space requirements.) To determine the admittance values of the lines and stubs, one uses a prototype low-pass impedance network as shown in Figure 3.4. To determine the values of g_1 through g_4 , one must first find R_A and δ , where R_A is the input impedance of the antenna at resonance and δ is the ratio of R_A to the imaginary part of the impedance at the band edges. From the measured values of the slotted disk in Table 3.1:

$$R_A = 11.5\Omega$$

$$\delta = 0.5 (\delta_1 + \delta_u) = 0.266$$

$$\text{where } \delta_1 = R_A / 32.2 \text{ for the lower band edge,}$$

$$\text{and } \delta_u = R_A / 62.0 \text{ for the upper band edge.}$$

The value of δ is taken as the average value for the upper and lower band edges (δ_u and δ_1) since these two values are not identical.

The series stub Z_1 , the shunt stubs Y_k , and the connecting lines $Y_{k,k+1}$ are all a quarter-wavelength long at midband

Figure 3.3 Impedance Matching Network formed from Stubs and Lines.
From [20].

Figure 3.4 A Load with a Low-Pass Impedance-Matching Network (case of $n = 4$). From [20].

Knowing R_A and δ , one can determine the values g_1 through g_4 using Figure 4.09-7 from [20]. These values are given below in Table 3.2.

g_0	= 1
w_1	= 1
g_1	= 3.8
g_2	= 0.47
g_3	= 2.4
g_4	= 0.36

Table 3.2 Conductance values for prototype low-pass impedance-matching network.

To derive the required stub admittances, the design equations of Table 11.09-1 from [20] are used. For the case of $n=3$, these equations reduce to:

$$Y_2 = G_A \left(N_{23} - \frac{J_{23}}{G_A} \right) \quad (\text{Eq. 3.2})$$

$$Y_3 = G_A \left(C_3'' \tan(\theta_1) + N_{23} - \frac{J_{23}}{G_A} \right) \quad (\text{Eq. 3.3})$$

$$Y_{23} = G_A \left(\frac{J_{23}}{G_A} \right) \quad (\text{Eq. 3.4})$$

where:

$$\theta_1 = \frac{\pi}{2} \left(1 - \frac{w}{2} \right) \quad (\text{Eq. 3.5})$$

(Eq. 3.6)

$$C_3' = C_3 - g_2$$

(Eq. 3.7)

$$C_3 = g_2 g_4 \frac{R_A}{R_B}$$

(Eq. 3.8)

$$\frac{J_{23}}{G_A} = \frac{1}{g_0} \sqrt{\frac{C_3}{g_3}}$$

(Eq. 3.9)

$$N_{23} = \sqrt{\left(\frac{J_{23}}{G_A}\right)^2 + \left(\frac{g_2 \tan(\theta_1)}{g_0}\right)^2}$$

and

w is the desired fractional bandwidth,

$G_A = 1 / R_A$ is the load admittance,

R_B is the desired input impedance.

Using the values in Table 3.2, the above equations become:

$$w = (1035 - 935) / 987 = 0.101$$

$$\theta_1 = 1.4915$$

$$C_3 = 0.1987$$

$$C_3' = -0.2713$$

$$J_{23}/G_A = 0.2877$$

$$N_{23} = 5.9217$$

and thus the line and shunt admittances become:

$$Y_2 = 0.50$$

$$Y_3 = 0.193$$

$$Y_{23} = 0.025$$

As a design example, the above impedance matching network can be fabricated on microstrip and the above admittances can be translated to corresponding microstrip line elements of appropriate length and width. For the common alumina substrate (with relative permittivity of 10 and substrate thickness of 0.060") the corresponding microstrip dimensions were determined (using LineCalc) and are shown below in Table 3.3.

IMPEDANCE (Ω)	WIDTH (mm)	$\lambda/4$ at 987 MHz (mm)	ϵ_{eff}
50.00	1.38	29.73	6.52
40.00	2.16	29.02	6.84
5.18	31.99	24.95	9.26
2.00	88.73	24.34	9.74
20.73	6.03	27.23	7.77
8.00	19.58	25.45	7.57

Table 3.3. Microstrip line dimensions for Tchebychev network.

As can be seen in the above table, the impedance values of 5.18 Ω and 2.00 Ω (which correspond to Y_2 and Y_3) translate into microstrip lines with unacceptably large widths. To decrease this dimension, the impedance of the lines can be increased by a factor of four by taking

the following steps: replacing the single shorted stub by two stubs in parallel, each stub having twice the impedance; and by replacing the quarter wavelength shorted stubs with half wavelength open stubs (with double the impedance). The resulting stub widths are shown as the last two entries in Table 3.3 and their values are far more reasonable.

In order to account for the discontinuities at impedance steps, junctions and line ends, the Tchebychev matching network was optimized using Touchstone. The final values are presented in the Touchstone file in Figure 3.5 and the matching network is shown in Figure 3.6. The theoretical return loss of the microstrip radiator with the Tchebychev matching network is plotted in Figure 3.7. Comparing the return loss of this network to that of the quarter wavelength transformer (Figure 3.2), it is clear that the Tchebychev matching network sacrifices the mid-band return loss for a better return loss over a wider band. The theoretical 10 dB return loss bandwidth of the Tchebychev network is double that of the quarter wavelength match. Further improvements in this performance can be achieved by adding more resonant sections to the Tchebychev network and by increasing the substrate permittivity to mitigate the additional space requirements.

The same procedure was carried out for the shorted disk (Figure 2.13) and yielded similar results. These bandwidth results are encouraging but there are still problems. From the Tchebychev network described above, a bandwidth in the order of 4% is achievable. However, for cellular communications, a bandwidth in the order of 10% is desirable. A solution to this problem may be to fabricate the matching

```

DIM
    FREQ GHZ
    LNG CM

VAR
    LL1 # 2.0 5.682 7.5
    LL2 # 2.0 6.310 7.5

CKT
    MSUB ER=10.0 H=C.1524 T=0.0015 RHO=1 RGH=C

! Unmatched Antenna
    S1PA 1 0 SLOTDISK.S1P
    DEF1P 1 DISK

! Antenna with Quarter Wavelength Transformer: 11 ohms to 50 ohms
    MLIN 1 2 W=0.491 L=2.759
    S1PA 2 0 SLOTDISK.S1P
    DEF1P 1 QDISK

! Antenna with Tchebychev Matching Network
    S1PA 10 0 SLOTDISK.S1P
    MLIN 10 1 W=0.138 L # 3.0 6.262 7.0
    MCROS 1 2 3 4 W1=0.138 W2=1.958 W3=0.216 W4=1.958
    MLEF 2 W=1.958 L ^ LL1
    MLEF 4 W=1.958 L ^ LL1
    MLIN 3 5 W=0.216 L # 1.0 1.078 4.0
    MCROS 5 6 7 8 W1=0.216 W2=0.603 W3=0.138 W4=0.603
    MLEF 6 W=0.603 L ^ LL2
    MLEF 8 W=0.603 L ^ LL2
    MLIN 9 W=0.138 L=1.0
    DEF1P 9 MATCHED

FREQ
    SWEEP 0.930 1.030 0.002

OUT
    DISK DB[S11] GR1
    QDISK DB[S11] GR1
    MATCHED DB[S11] GR1

GRID
    GR1 -20 0 2

OPT
    RANGE 0.970 1.010
    MATCHED DB[S11] = -11.5

```

Figure 3.5 Touchstone Circuit File for Optimized Tchebychev Matching Network.

Figure 3.6 Optimized Tchebychev Matching Network.

Figure 3.7 Return Loss of Slotted Disk with Optimized Tchebychev Matching Network.

network on a thicker substrate with very high dielectric constant. (Ceramic substrates are currently available with relative permittivity in the hundreds.) Not only would the size of the network be significantly reduced, but extra sections could then be added to increase the bandwidth. Pursuit of this area of bandwidth enhancement is left to future work.

4. GEOMETRIC THEORY OF DIFFRACTION

4.1 Introduction

The equations (Eq. 2.3) and (Eq. 2.4) for the radiation pattern of a circular microstrip disk were derived for the case of the disk located on an infinite ground plane. However, the roof-top of a vehicle cannot be considered an infinite ground plane when operating at frequencies from 800 MHz to 900 MHz. At this frequency band, the roof-top would measure in the order of five or six wavelengths in each dimension. The effects of the finite ground plane on the antenna radiation pattern, which are not accounted for by the above equations, can be predicted by utilizing the Geometric Theory of Diffraction (GTD). This theory was first developed by Keller [21] as an extension of Geometric Optics (which deals with direct, reflected and refracted rays) to predict the diffraction of optical rays by edges or slots. The theory has been improved and extended over the years [22] to include electromagnetic waves. These waves are essentially treated as rays and the same principles of diffraction are applied to these rays as is done with the optical rays. The salient principles of GTD will now be reviewed and will then be applied to predict the radiation patterns of the microstrip disks mounted on finite ground planes in order to simulate the effects of vehicle roof-tops.

4.2 The Conducting Wedge: A GTD Canonical Problem

The Geometric Theory of Diffraction is a high-frequency asymptotic technique which can be utilized to solve complicated geometries. This is done by resolving the complex geometries into smaller, simpler components with known solutions. These simple geometries are known as canonical problems and since diffraction is a local phenomenon, a canonical problem with the same local geometry as the object under investigation can be used to solve the diffraction field at that location. In order to introduce the principles of GTD, the two-dimensional canonical problem of a conducting wedge will be investigated.

The geometry of the conducting wedge problem is shown in Figure 4.1. It consists of an infinite line source (either electric or magnetic) placed near a two-dimensional electric conducting wedge of angle $(2 - n)\pi$ radians. One can treat the field at an observation point P as the sum of three fields: the direct field, the reflected field and the diffracted field. Furthermore, depending on the location of the observation point, either the direct, reflected or both fields may be zero due to the blockage by the conducting wedge. Thus, the area surrounding the wedge can be categorized into three regions which are characterized by the fields present in each region as shown below in Table 4.1. If (as is the case with Geometric Optics) the diffracted field were not taken into account, it is obvious that there would be discontinuities along the boundary separating Regions I and II (Reflected Shadow Boundary) and Regions II and III (Incident Shadow

Figure 4.1 The Conducting Wedge GTD Canonical Problem. From [23].

REGION	SECTOR	FIELDS PRESENT
I	$(0 < \phi < \pi - \phi')$	Direct, Reflected, Diffracted
II	$(\pi - \phi' < \phi < \pi + \phi')$	Direct, Diffracted
III	$(\pi + \phi' < \phi < n\pi)$	Diffracted

Table 4.1. Field regions surrounding the 2-D conducting wedge. [23]

Boundary). To remedy this situation, the diffracted fields must be included. The diffracted fields will modify the fields in all three regions, removing the discontinuities and filling in the shadow region. When the source is located a distance s' from the point of diffraction, and the observation point a distance s , the diffracted field can be expressed as [23]

$$\vec{E}^d(s) = \vec{E}^i(Q) \cdot \vec{D} A(s', s) e^{jks} \quad (\text{Eq. 4.1})$$

where $A(s', s)$ is the spreading factor given by:

$$A(s', s) = \begin{cases} \frac{1}{\sqrt{s}} & \text{for plane, cylindrical, or conical wave incident} \\ \sqrt{\frac{s'}{s(s+s')}} & \text{for spherical wave incident} \end{cases}$$

The above equation states that the diffracted field at the observation point at a distance s from the diffracted edge, is determined by the

value of the incident wave at the point of diffraction Q, the diffraction coefficient and the spreading factor. The diffraction coefficient is, in general, a dyadic and is usually determined either by asymptotic solutions to canonical problems or by experimentation.

The geometry which applies to the disk antenna mounted on a finite ground is that of an incident ray contacting the edge of a wedge as shown in Figure 4.2. This figure shows the two possible scenarios: for the case of a normal incident ray, Figure 4.2a shows that the diffracted rays form a plane normal to the diffracting edge. For the case of an oblique incident ray, the diffracted rays form a cone whose angle equals the incident angle (assuming a homogeneous medium) as shown in Figure 4.2b. In the case of the disk antenna mounted on the square ground plane, the elevation pattern cuts were taken normal to the ground plane, which corresponds to diffraction due to normal incidence (Figure 4.2a). Thus, the edge diffraction coefficient for normal incidence is the one to be determined. This diffraction coefficient has been determined and is given by [24]:

$$D(\rho', \beta^-, \beta^+, n) = D^i(\rho', \beta^-, n) \pm D^r(\rho', \beta^+, n) \quad (\text{Eq. 4.2})$$

where

$$D^i(\rho', \beta^-, n) = K_{en} \{ C^+(\beta^-, n) F[k\rho' g^+(\beta^-) + C^-(\beta^-, n) F[k\rho' g^-(\beta^-)] \}$$

$$D^r(\rho', \beta^+, n) = K_{en} \{ C^+(\beta^+, n) F[k\rho' g^+(\beta^+) + C^-(\beta^+, n) F[k\rho' g^-(\beta^+)] \}$$

(a) Normal incidence

(b) Oblique incidence

Figure 4.2 Diffraction Mechanism by a Conducting Wedge for Normal and Oblique Incidences. From [23]

$$K_{en} = \frac{-e^{-j\pi/4}}{2n\sqrt{2\pi k}}$$

$$C^+(\beta, n) = \cot((2\pi + \beta)/(2n))$$

$$C^-(\beta, n) = \cot((2\pi - \beta)/(2n))$$

$$F(x) = 2j\sqrt{x}e^{jx} \int_{\sqrt{x}}^{\infty} e^{-jt^2} dt$$

$$\beta^+ = \phi + \phi'$$

$$\beta^- = \phi - \phi'$$

$$g^+ = 1 + \cos[(\phi \pm \phi') - 2n\pi N^+]$$

$$g^- = 1 + \cos[(\phi \pm \phi') - 2n\pi N^-]$$

where N^+ or N^- is a positive or negative integer or zero, whichever most closely satisfies the equation:

$$2n\pi N^+ - (\phi \pm \phi') = +\pi \quad (\text{for } g^+)$$

$$2n\pi N^- - (\phi \pm \phi') = -\pi \quad (\text{for } g^-)$$

The $\pm$ in (Eq. 4.2) depends on the polarization of the electric line source. For soft polarization (electric line source parallel to the edges of the conducting wedge), the lower sign (-) is used, while for hard polarization (magnetic line source horizontal to conducting wedge) the upper sign (+) is used.

Once the edge diffraction coefficient is computed, the diffracted field can be determined from (Eq. 4.1). The elevation plane radiation patterns for circular aperture antennas on finite ground planes have been simulated by Cockrell and Pathak [25] using the diffracted fields determined by the above equations. Their method will now be used (with

slight modifications) in this work to predict the elevation plane patterns of the microstrip disk antennas mounted on a finite rectangular ground plane.

4.3 Effects of Finite Ground Planes on Disk Antenna Patterns

Figure 4.3 depicts the microstrip disk antenna centred on a square finite ground plane. For an elevation plane radiation pattern taken at $\phi = 90^\circ$, the pattern can be predicted by summing the field due to geometric optics (E_{GO}) and the fields diffracted at the two edges (E^D_1 and E^D_2). The diffracted field at angle θ from the edge at Q_1 is expressed as:

$$E^D_1(\theta) = E^i(Q_1) \cdot D_n(\rho' = d_1, \beta, n=2) \cdot A(d_1, r_1) \cdot e^{-jk r} \quad (\text{Eq. 4.3})$$

Now the incident field at Q_1 is:

$$E^i(Q_1) = 0.5 E_{GO}(r=d_1, \theta=\pi/2) \quad (\text{Eq. 4.4})$$

The factor of 0.5 is used since the incident and reflected fields merge for grazing incidence [25]; therefore only one half the total field contributes to the incident field.

By normalizing (Eq. 2.4) E_{GO} is obtained:

$$E_{GO}(r, \theta) = (J_{n+1}(k a \sin(\theta)) - J_{n+1}(k a \sin(\theta))) \cdot e^{-jk r} / r \quad (\text{Eq. 4.5})$$

Figure 4.3 Direct and Diffracted Fields of a Centre-Fed Microstrip Disk on a Finite Ground Plane.

and for $r=d_1$, $\theta=\pi/2$, $n=0$, this becomes:

$$E_{G0}(Q_1) = 2J_1(ka) \cdot e^{-jkd_1/d_1}$$

The hard diffraction coefficient (D_h) is the appropriate one to use since the E-fields of the microstrip disk are vertically polarized. For this configuration we have:

$$D_h(\rho'=d_1, \beta, n=2) = D^i(\rho'=d_1, \beta^-, n=2) + D^r(\rho'=d_1, \beta^+, n=2) \quad (\text{Eq.4.6})$$

Now, for the angle of incidence from the source towards the point of diffraction equal to zero degrees (i.e. $\phi' = 0^\circ$) then:

$$\beta^- = \phi_1 - \phi' = \phi_1 + \phi' = \beta^+ = \beta$$

Thus:

$$D_h(\rho'=d_1, \beta, n=2) = 2D^i(\rho'=d_1, \beta, n=2) = 2D^r(\rho'=d_1, \beta, n=2) \quad (\text{Eq.4.7})$$

$$\text{Also, } A(d_1, r_1) = \sqrt{d_1}/r_1 \quad (\text{Eq.4.8})$$

Now, for far-field measurements the following approximations can be made:

$$r_1 \approx r - d \sin\theta \quad \text{for phase terms}$$

$$r_1 \approx r \quad \text{for amplitude terms}$$

Thus the diffracted field of (Eq. 4.3) becomes:

$$E_1^d(\theta) = 2J_1(ka) \cdot D^1(\rho'=d_1, \beta=\theta+\pi/2, n=2) \cdot \frac{e^{-jkd(1+\sin\theta)}}{\sqrt{d_1}} \cdot e^{-jkr} \quad (\text{Eq. 4.9})$$

Similarly, the diffracted field from the edge at Q_2 can be derived as:

$$E_2^d(\theta) = 2J_1(ka) \cdot D^1(\rho'=d_2, \beta=\pi/2-\theta, n=2) \cdot \frac{e^{-jkd(1+\sin\theta)}}{\sqrt{d_2}} \cdot e^{-jkr} \quad (\text{Eq. 4.10})$$

A computer program (GTD.FOR) was written in Microsoft FORTRAN to compute the geometric optics field and diffracted fields for a centre-fed microstrip disk on a finite ground plane. A listing of the source code along with program documentation is found in Appendix 1. To estimate the semi-infinite Fresnel Integral necessary to obtain the diffraction coefficients, a numerical algorithm reported by Boersma [26] was utilized. The GTD.FOR program was then used to predict the normalized elevation patterns of the various disks presented in Chapter 2.

Figure 4.4 shows the superposition of the Geometric Optics Field and the diffracted fields for the case of the centre-fed disk of Figure 2.4. As can be seen, the Geometric Optics field is only valid in the region $\theta \leq 90^\circ$; the diffracted fields are responsible for the back lobes in the region $90^\circ < \theta \leq 180^\circ$. Figure 4.5 compares the predicted and measured patterns of the centre-fed microstrip disk at 1850 MHz mounted on a 50 cm x 50 cm ground plane. There is good agreement between the two patterns, with the GTD curve following the back lobe

Figure 4.4 Superposition of Direct and Diffracted Fields for Normalized Elevation Pattern of Centre-Fed Disk on Finite Ground Plane.

Figure 4.5 Theoretical and Measured Normalized Elevation Pattern of Centre-Fed Disk at 1850 MHz on 50cm x 50cm Ground Plane.

levels quite closely and accurately predicting the number of lobes. Figure 4.6 compares the measured and simulated elevation patterns for the slotted disk with 1.0 cm slot spacing (Figure 2.18a) at 1400 MHz. Once again the GTD is seen to correctly predict the number of back lobes. Two trends in the elevation pattern (which will be seen to continue in later figures) occur as the frequency is altered. For a fixed size ground plane, the number of back lobes decreases as the frequency decreases and the level of back lobes increases as the frequency decreases. The second trend is intuitively obvious; at lower frequencies, the electrical size of the ground plane decreases and thus more power "leaks" out around the ground plane into the back lobes. Conversely, if the frequency were increased, the ground plane would seem electrically larger and less power would be radiated into the back lobes. Taking this situation to the limit, if the frequency were increased enough, the ground plane would appear infinite in extent, and no power would be radiated in the back lobe region. Thus there would be no diffracted fields and the radiation pattern would simply be that of the geometric optics field.

The first trend observed (the increase in the number of back lobes with increasing frequency) can be rationalized by treating the two diffraction points (one on each edge of the ground plane) as isotropic sources radiating in phase. The increase in frequency is equivalent to an increase in separation between the two sources. The further the separation, the more ripples there will be in the interferometer radiation pattern (this can be seen by plotting (Eq.

Figure 4.6 Theoretical and Measured Normalized Elevation Pattern of Slotted Disk (1.0 cm slot spacing) at 1400 MHz on 50cm x 50cm Ground Plane.

2.6) with increasing values of d). These ripples correspond to the lobes in the diffracted field patterns. Conversely, if the isotropic sources are brought closer together (i.e. decreasing the frequency) the number of lobes decreases until in the limit the two sources are superimposed and an isotropic pattern is obtained.

One drawback to the GTD is that it is a high-frequency asymptotic technique; that is, the GTD works well when the dimensions of the antenna and ground plane are large compared to the wavelength. As these dimensions are reduced, the asymptotic assumptions made for determining the diffraction coefficients begin to break down and the GTD predictions begin to deteriorate. This can be seen in Figure 4.7 which compares the measured elevation pattern for the slotted disk with 0.5 cm slot spacing at 1175 MHz with the predicted GTD pattern. Discontinuities appear at the shadow boundaries ($\theta = 90^\circ$) and at $\theta = 0^\circ$. Nevertheless, away from the shadow regions, the agreement is still quite good. However, as the frequency is further reduced, as is the case in Figure 4.8 (the slotted disk with 0.25 cm slot spacing at 900 MHz) and Figure 4.9 (the shorted disk at 430 MHz) the asymptotic assumption become further and further in error and GTD fails to accurately predict the measured values. At these frequencies (where antenna dimensions are comparable to the wavelength), one would have to turn to alternate methods (such as the method of moments) to predict the effects of the finite ground planes on the radiation patterns.

To conclude this chapter, the GTD will be used to predict the effects of a ground plane whose size more closely corresponds to that

Figure 4.7 Theoretical and Measured Normalized Elevation Pattern of Slotted Disk (0.5 cm slot spacing) at 1175 MHz on 50cm x 50cm Ground Plane.

Figure 4.8 Theoretical and Measured Normalized Elevation Pattern of Slotted Disk (0.25 cm slot spacing) at 900 MHz on 50cm x 50cm Ground Plane.

Figure 4.9 Theoretical and Measured Normalized Elevation Pattern of Shorted Disk at 430 MHz on 50cm x 50cm Ground Plane.

of an actual vehicle roof-top. Figure 4.10 presents two normalized elevation patterns: the first is a repetition of the measured pattern of the centre-fed disk antenna given in Figure 2.8. The antenna was mounted on a ground plane of dimensions $3\lambda \times 3\lambda$. For a better approximation of a vehicle roof-top, a ground plane of $6\lambda \times 6\lambda$ would be more appropriate. The second pattern in Figure 4.10 is the GTD prediction for the antenna mounted on this larger ground plane. The results are interesting in that the GTD predicts the elevation pattern to have more energy concentrated towards the horizon which is a desired result for cellular communications. These results are to be expected since as the ground plane is increased, it behaves more and more like an infinite ground plane, and the radiation pattern approaches that of Figure 2.7 (the case with $\epsilon_r = 2.5$). This is in contrast to the case with the commercial antenna investigated in Chapter 2, where the introduction of the ground plane caused a deterioration in the radiation pattern (Figure 2.4).

Figure 4.10 Theoretical Elevation Patterns of Centre-Fed Disk at 850 MHz on Different-Sized Ground Planes.

5. ANTENNA ARRAY FOR MOBILE SATELLITE COMMUNICATIONS

5.1 Introduction

Mobile satellite communications can be thought of as an extension of city-bound mobile cellular communications. One of the variety of applications for mobile satellite communications is for the transmission of voice and data information to commercial vehicles travelling between cities which must maintain close contact with headquarters. This chapter focuses on a novel microstrip antenna array designed to be mounted on a large commercial vehicle for use in mobile satellite communications.

5.2 Antenna Requirements

In order to be applicable to mobile satellite communications, an antenna must meet several requirements. It must be able to track the location of the satellite as the vehicle changes position; thus the beam of the antenna must be steered by either electrical or mechanical means. The antenna must have a wide bandwidth (1530 MHz to 1660.5 MHz for this particular application). The antenna must have a high gain in order to provide a good signal to noise ratio to relax the specifications on the transceiver performance. The antenna must be circularly polarized in order to conform to the satellite antenna polarization. Finally, the antenna must be relatively compact, for mounting on a commercial vehicle. A current available antenna for such an application consists of a five-element array which is mechanically

steered to track the satellite. Such a configuration suffers from several disadvantages including its low reliability and high cost and the large volume required. An alternative novel microstrip antenna array will now be examined.

5.3 The Cylindrical Patch Array

The Cylindrical Patch Array (CPA) consists of an array of eighteen circularly-polarized microstrip elements mounted on a cylindrical cone as shown schematically in Figure 5.1. The array produces a high-gain beam which is electronically steered in the azimuthal plane for satellite tracking. The beam is fixed in the elevation plane such that the beam peak is aligned with the satellite. The advantages of this array include: the absence of moving parts (due to electronic steering); it occupies a relatively small volume, and is relatively inexpensive.

There are a host of design issues which must be addressed and resolved when setting out to fabricate such an array; issues from mechanical integrity to electrical performance. Only the major issues pertaining to the array geometry, antenna elements and consequent effects on radiation patterns will be discussed in detail in this work. Other important factors which are directly related to the antenna (such as the feed structure and phase shifters) will also be discussed but in less depth.

Figure 5.1 Schematic of the Cylindrical Patch Array.

2

of/de

2

PM-1 3½"x4" PHOTOGRAPHIC MICROCOPY TARGET
NBS 1010a ANSI/ISO #2 EQUIVALENT

5.4 CPA Design issues

5.4.1 Mechanically- vs. Electronically-Steered Arrays

There are two conventional methods used for steering beams: mechanical steering and electronic steering (which includes techniques such as phase scanning, beam switching and frequency scanning), each of which has advantages and disadvantages. Using mechanical steering, the actual antenna is, in general, less complex than for electronically steered arrays. However, much effort must be placed into the mechanical beam steering unit which will usually contribute a significant portion of the overall costs. Many problems involving the performance and reliability of the moving parts may arise when dealing with mechanical steering. One must also consider the overall swept volume required by the antenna as it moves to track the satellite, thereby adding to the overall dimensions. (The swept volume is significant due to aerodynamic drag and aesthetic factors.)

Electronically steered antennas, on the other hand, require no moving parts and thus tend to be more reliable. However, the actual antenna is, in general, more complex than its mechanically steered counterpart since either a planar array of phased elements or a three dimensional array is necessary for hemispherical coverage. Thus while the prime focus of a mechanically-steered antenna is on mechanical design, that of an electronically-steered array is on such factors as power distribution to the elements, phase shifter design and control, and element spacing. Much of the complexity of electronically-steered

array can be mitigated if beam pointing can be restricted to only one polar co-ordinate instead of two. For steering in one polar co-ordinate, only a linear array of phased elements is required. This has been done for the case of the Cylindrical Patch Array. By placing the radiating elements around an inclined cylinder, the beam peak is aligned to point towards the satellite, thus no steering is required in the elevation plane. The beam need only be steered in the azimuthal plane therefore only a linear array is required. (In the case of the CPA, the linear array has been wrapped around a cylinder.) Geometric considerations for this arrangement will now be examined.

5.4.2 Geometrical Considerations

The dimensions of the CPA cylinder are shown in Figure 5.2. The size of the cylinder is a compromise between the conflicting requirements of minimizing the total space occupied by the array and maximizing the number of radiating elements to increase the number of possible discrete beam positions for increased resolution. The size of the microstrip element is dictated by the frequency range and bandwidth requirements. For adequate coverage, it was decided that 36 beam positions were required (i.e. the beam could be steered around the array in 10° increments). In order to obtain 36 beam positions, 18 radiating elements are required (as will be seen in 5.4.3). The shape of the microstrip element is that of a modified square patch, the choice of which will be described in 5.4.5. To help reduce the mutual coupling between elements, the patches are placed such that the

Figure 5.2 CPA Dimensions.

diagonals are collinear (Figure 5.2) thus dictating the mid-way circumference of the inclined cylinder. Once this circumference has been established, the minimum and maximum diameters of the CPA, along with the vertical height are then determined by simple geometry.

5.4.3 Mode of Operation

In order to produce a high-gain beam in a particular direction, an appropriate number of adjacent elements must be turned on. For example, in Figure 5.3, if three elements (1 2 3) produce beam A, then turning off element 1 and turning on 4 would form beam B, shifted over by ϕ° . Similarly, turning off element 2 and turning on 5 would produce beam C, shifted by $2\phi^\circ$ from beam A. For a fixed number of adjacent elements turned on (three in the above example), the total number of beam positions available is N, where N is the total number of elements. Thus, for the 18 element CPA, only 18 beam positions are available, spaced 20° apart. To double the number of beam positions, the number of elements which are turned on is varied, as shown in Figure 5.4. In this manner, 36 beam positions, spaced 10° apart can be achieved.

Another consideration is the number of elements which must be turned on to produce an acceptable beam pattern. The trade-off lies between beam shape and gain. Initially, by turning on more elements, a narrower beam can be achieved but, due to the circular geometry, there comes a point where the gain begins to drop off as more elements are turned on. (Ultimately, if all the elements are turned on, an omnidirectional pattern results.) To investigate this trade-off, the

ELEMENTS ON	BEAM
1 2 3	A
2 3 4	B
3 4 5	C

Figure 5.3 Beam Steering for a Fixed Number of Elements On.

ELEMENTS ON	BEAM
1 2 3	A
1 2 3 4	AB
2 3 4	B
2 3 4 5	BC
3 4 5	C

Figure 5.4 Beam Steering for a Variable Number of Elements On.

azimuthal plane radiation patterns for various numbers of patterns were simulated using standard array techniques as described in the next section.

5.4.4 Simulated Azimuthal Radiation Patterns

In order to obtain a first-order prediction the azimuthal pattern and gain of the CPA as a function of the number of elements, the CPA geometry was approximated as a circular ring of elements as shown in Figure 5.5 and standard array techniques were applied [19]. The approximate azimuthal pattern is given by:

$$AP = \frac{1}{N_{on}} \sum_{n=1}^{N_{on}} I(n) e^{j\psi(n)} \quad (\text{Eq. 5.1})$$

where the phase term is given by:

$$\psi(n) = x(n) \cos(\phi) + y(n) \sin(\phi)$$

and the magnitude is approximated by:

$$I(n) = \begin{cases} [\cos(\phi - \phi'(n))]^{2.5}, & \text{if } \cos(\phi - \phi') \geq 0 \\ 0 & \text{if } \cos(\phi - \phi') < 0 \end{cases}$$

The amplitude factor $I(n)$ is used as an approximation to the radiation pattern of a microstrip square patch. Using (Eq. 5.1) the approximate azimuthal patterns were determined. Figure 5.6 shows the azimuthal patterns for 3 to 6 elements turned on. The patterns for each element has been shifted so that the beam peaks are aligned for ease of

Figure 5.5 Circular Ring Array.

Figure 5.6 Simulated Azimuthal Patterns for 3 to 6 Elements.

comparison. As can be seen, as the number of elements turned on is increased, the beam becomes narrower. This trend continues until the fifth element. When the sixth is turned on, the beam begins to broaden once again. The narrowest beams are obtained with four and five elements turned on and thus it was decided to switch between four and five elements to obtain the desired beam 10° beam shifts. The narrow beam widths are desirable to avoid interference with any other satellite in the vicinity.

5.4.5 CPA Radiating Element

The size and shape of the microstrip element to be used for the CPA depend on the frequency of operation, the required bandwidth and the required polarization. As has been mentioned in previous chapters, the impedance bandwidth of a microstrip patch is a function of its thickness and the dielectric constant of the substrate. In order to maximize the bandwidth, the element should have a large thickness and low substrate permittivity. To achieve the desired bandwidth (1530–1660.5 MHz) a honeycomb substrate (with $\epsilon_r \approx 1$) of 1.252 cm thickness was chosen.

The shape of the radiating element was chosen to minimize the axial ratio in order to provide the best reception of circularly polarized signals. There are a few ways to produce circularly polarized waves using microstrip elements [27–31]; the most common approach (and the one which provides the widest bandwidth) is to feed the microstrip patch at two points, with one point phase shifted by 90° as shown in

Figure 5.7. For the case of a square patch, the feed points lie along the perpendicular bisectors of two adjacent sides of the patch, while for a circular patch, the points lie on radii spaced 90° apart. To provide the 90° phase shift, the phase shifter of Figure 5.8 was used. The phase shifter was fabricated on 0.060" Duroid ($\epsilon_r = 2.3$) and the ground plane of the phase shifter also served as the ground plane of the radiating elements as shown in Figure 5.9. Two pins were fed from the phase shifter up through the honeycomb substrate and soldered onto the patch surface to provide the circularly polarized wave. The radiating element used was the modified square patch shown in Figure 5.10. Experimental modifications were carried out in order to reduce the size of the patch for the given resonant frequency. Using the design equations from Bahl and Bhartia [32], and for the substrate thickness and permittivity mentioned above, at 1595 MHz a standard square patch would have a resonant length of 7.7 cm, while a circular patch would have a resonant diameter of 8.9 cm. The modified square patch had a resonant length of 6.2 cm. The return loss of the modified patch connected to the phase shifter was measured and is shown in Figure 5.11. A better than 10 dB return loss was achieved across the bandwidth of interest which was an acceptable match. The axial ratio of the modified element was then measured by using a spinning dipole source mounted on a 2' diameter parabolic dish. The resulting pattern at mid-band is presented in Figure 5.12. Table 5.1 summarizes the axial ratios at mid-band and at the band edges. As can be seen, the maximum axial ratio is 2 dB over the bandwidth of interest, which met

Figure 5.7 Circularly Polarized Patches.
(a) Square Patch
(b) Circular Patch

Figure 5.8 Phase Shifter for Circular Polarization.

Figure 5.9 Patch and Phase Shifter Geometry.

Figure 5.10 Modified Square Patch

Figure 5.11 Measured Return Loss for Patch and Phase Shifter.

Figure 5.12 Measured Axial Ratio of Modified Square Patch at 1595 MHz.

the required specifications.

FREQUENCY (MHZ)	MODIFIED SQUARE PATCH
1530	2
1595	1.5
1660	2

Table 5.1 Axial Ratio (in dB) of the radiating elements.

The modified square patch was chosen over the conventional square and circular patches since it provided comparable performance with respect to axial ratio, beam shape and impedance bandwidth, yet at a significantly reduced size. Using the modified element, the overall dimensions of the CPA were substantially reduced.

5.5 CPA AZIMUTHAL PATTERNS

A five-element model was first constructed using the modified elements. A simple five-way power splitter (Figure 5.13) was designed and fabricated to provide power to each element. The radiation patterns were then measured in the azimuthal plane normal to the elements (i.e. at an elevation angle of 30° from the horizon) for 3, 4 and 5 elements on. (The unused branches of the power splitter were terminated with matched loads.) The normalized patterns are presented in Figures 5.14 to 5.16 along with the respective predicted patterns of Figure 5.6. As can be seen, the measured beam widths are approximately 50% wider than

Figure 5.13 Five-Way Power Splitter.

Figure 5.14 Measured Azimuthal Pattern for 3 Elements On.

Figure 5.15 Measured Azimuthal Pattern for 4 Elements On.

Figure 5.16 Measured Azimuthal Pattern for 5 Elements On.

those predicted using (Eq. 5.1). These results can be attributed to the fact that this equation simplifies the geometry and radiating element; it was used only to determine the general trends of pattern beam width versus the number of elements.

In order to narrow the beam width and lower the sidelobes to more desirable levels, it was decided to provide a limited amount of phase shifting (in the form of a one-bit phase shifter) to the elements. Once again, (Eq. 5.1) was used to predict the general effects of adding phase shifts. For the case of four elements on, a phase β was introduced to the two middle elements and the effects on the array pattern were simulated. Similarly, for the case of five elements, the same phase shift β was introduced to the three middle elements. The objective was to find an optimum value of β which produced a narrowing of the beam and a lowering of the sidelobes for both cases. The simulated effects of various phase shifts on the five element array is shown in Figure 5.17. Based on numerous trials, it was decided to incorporate a phase shift of 50° into the CPA.

A phase shifter was designed as shown in Figure 5.18 and several were fabricated. P-I-N diodes were used to turn on or turn off the phase shifters when appropriate (using computer controlled logic). Measurements were taken using a full 18-element prototype (Figure 5.19). The prototype array was fed with an 18-way power switch (Figure 5.20). Through the use of P-I-N diode switches, power is channelled into any four or five adjacent elements around the cylinder thus creating the electronically-steerable high-gain beam. The azimuthal

Figure 5.17 Simulated Azimuthal Patterns for 5 Elements On With Various Phase Shifts.

Figure 5.18 Fabricated Phase Shifter.

Figure 5.19 Cylindrical Patch Array.

Figure 5.20 18-Way Power Switch.

patterns for four and five elements were measured and are shown in Figures 5.21 and 5.22. As predicted, the beam widths in both cases are narrower and sidelobes are reduced. The theoretical predictions are now much closer to measured values in the region around the beam peak, while the measured sidelobes are significantly lower than predicted. These lower sidelobe levels seemed to indicate there might be amplitude tapering to the power supplied to the elements (i.e. the outer elements received less power than the middle elements). S_{21} measurements were carried out on the phase shifters and it was found that the insertion loss with the phase shifters off was approximately 1 dB lower than when they were on, thus confirming the amplitude taper. These measured results were included in the simple array model and the resultant pattern for 5 elements on is plotted in Figure 2.23. As can be seen, the sidelobe levels are reduced and are much closer to the measured values.

To determine the axial ratio of the array, cross-polarization measurements were taken at mid-band for four and five elements on as shown in Figures 5.24 and 5.25 respectively. At boresight, the cross-polarization levels were -25 dB for four elements on and -21 dB for five elements on. These correspond to axial ratios of 0.96 dB and 1.5 dB respectively (see Appendix 2 for conversion calculations from cross-polarization to axial ratio).

Finally, a gain standard was used to determine the gain of the CPA. Table 5.1 shows the resulting gains at mid-band and at the band edges for the CPA with four and five elements turned on. The CPA

Figure 5.21 Measured Azimuthal Pattern with 4 Elements On.

Figure 5.22 Measured Azimuthal Pattern with 5 Elements On.

Figure 5.23 Simulated Azimuthal Pattern with 5 Elements On and Amplitude Taper.

Figure 5.24 Cross Polarization Pattern With 4 Elements On.

Figure 5.25 Cross Polarization Pattern With 5 Elements On.

surpassed the specification of 11dBic over the entire bandwidth.

ELEMENTS ON	1530 MHz	1595 MHz	1660 MHz
4	11.6	11.5	12.0
5	12.0	12.3	11.3

Table 5.1 Gain Measurements (dBic) for the CPA.

6. SUMMARY AND RECOMMENDATIONS

6.1 SUMMARY

This thesis has investigated several novel microstrip patches for use in mobile cellular communications and mobile satellite communications.

In Chapter 2, the disadvantages of a commercially available mobile cellular antenna were discussed. A typical commercial antenna was investigated and it was discovered that the radiation pattern was adversely affected by the presence of a ground plane simulating the effects of a vehicle roof-top. Due to many attractive features, it was decided to investigate microstrip antennas as a possible alternative to the commercial antenna. The centre-fed circular disk was first examined and was found to produce the desired omni-directional pattern. Unfortunately, the size of the disk radius required for operation at the desired frequency band was prohibitive for vehicle mounting. Techniques for reducing the disk size were thus pursued.

The first technique involved the addition of a lumped series inductor to lower the resonant frequency of the disk. A wire-wound ferrite bead was used, but it proved unsuccessful both mechanically and electrically. The use of printed spiral inductors was also investigated. Touchstone models were used to predict the performance of spiral inductors printed on the disk surface. Given the substrate thickness and permittivity of the disk antenna, the spiral traces behaved like distributed lines and not lumped elements. These

predictions were verified with several fabricated antennas. A deterioration in radiation pattern was also noted.

The second approach for reducing the resonant frequency involved the addition of shorting posts to produce electric walls. Array theory was used to predict the radiation pattern. A reduction to 23% was measured in the resonant frequency and there was no deterioration in the radiation pattern. Unfortunately, a large reduction in impedance bandwidth accompanied the decrease in resonant frequency.

The third technique involved cutting slots in the disk surface to increase the path length of the surface current thereby reducing the resonant frequency. A reduction to 50% was attained by cutting slots spaced 0.25 cm apart. This method proved quite versatile, since one could vary the resonant frequency simply by varying the slot spacing. Once again, the impedance bandwidth was reduced with frequency.

Impedance matching networks were then investigated to improve the impedance bandwidth of the modified microstrip disks. Both quarter wavelength transformers and Tchebychev matching networks were examined using Touchstone models. With the quarter wavelength transformer, a 10 dB bandwidth of 2% was attained, while the three-resonator Tchebychev network produced a 4% bandwidth. To minimize the space requirements of such networks, it was apparent that a high permittivity substrate should be used. This would allow for an increased number of resonators in the Tchebychev network in order to increase the bandwidth closer towards 10%, which is desirable for cellular communications.

The Geometric Theory of Diffraction (GTD) was used to predict the performance of the radiation patterns of the disk antennas in the presence of finite ground planes. The GTD provided good agreement with measured data when the antenna and ground plane dimensions were greater than one wavelength. The GTD was also used to predict the effects of ground planes whose size approximately equalled actual vehicle rooftops. The GTD predicted that antenna patterns would be improved over larger ground planes, which is not the case for the commercial antennas.

The second section of this thesis dealt with the design, fabrication and testing of an 18-element microstrip array for mobile satellite communications. The Cylindrical Patch Array (CPA) was designed to operate at L-band (1530 MHz to 1660.5 MHz) and consisted of microstrip elements which produced a high-gain beam, electronically steered to track the satellite. The microstrip elements consisted of novel modified square patches to decrease the dimensions of the resonant element and thus allow for a more compact array. Many important design issues related to the radiating elements and the resulting antenna patterns were discussed. Antenna patterns were simulated using simple array theory and predictions were compared to measured data. A five-element model was first constructed to verify the predicted patterns. The measured beam widths were found to be significantly wider than anticipated. Phase shifters were thus added to narrow the beam width and lower the side lobe levels. The 18 element CPA was constructed and the relevant antenna measurements were taken to

ensure that all designs specifications were met.

Recommendations for Future Work

There is much work that can still be carried out on the modified centre-fed disks for mobile communications. Such work might include the following:

- (1) More experimental work can be performed to derive design equations relating slot spacing to resonant frequency. The investigation can be expanded to determine the effects of changing the radius of the disk and changing the dielectric constant. The effects of non-uniform spacing on resonant frequency and radiation patterns can also be investigated.
- (2) Numerical analysis techniques (such as the method of moments) can be developed to determine the surface currents (and consequently the input impedance) for the slotted and shorted disks.
- (3) Matching networks on high-permittivity substrates can be pursued in an attempt to enhance impedance bandwidth performance of the modified disks.
- (4) Numerical techniques could be applied to predict the effects of finite ground planes on microstrip disks at frequencies where the GTD is no longer valid.

For the Cylindrical Patch Array, future work might include:

- (1) Further modifications to the element geometry could be pursued to further improve the axial ratio.

(2) More rigorous analysis of the array can be investigated, incorporating such factors as the finite ground plane of the radiating elements and the cylindrical array geometry into the analysis.

APPENDIX 1

DOCUMENTATION FOR GTD.FOR

Purpose:

This FORTRAN program GTD.FOR uses the Geometric Theory of Diffraction to model the radiation patterns of a circular microstrip disk on a finite square ground plane.

Procedure:

- (1) The geometric optics E-field (EGO) is computed for the microstrip disk using the equations derived for the simple cavity model (Eq. 2.3)
- (2) The Fresnel integral (subroutine FRNELS) is computed to determine the edge diffraction coefficients.
- (3) The diffracted fields (ED1, ED2) are determined using the diffraction coefficients and the geometric optics field.
- (4) The total field (ET) is calculated by summing the contribution from the geometric optics field plus the contribution from the diffracted fields.

Subroutines:

CF: computes the edge diffraction coefficients

FRNELS: approximates the infinite Fresnel integral

BESJ : computes the n^{th} -order Bessel function of the first kind

Key Variables:

ANG: angle of diffracted ray

D1, D2: distance from centre of disk to edges 1 and 2 (D1 = D2 for the case of the disk centred on a square ground plane)

ED1, ED2: diffracted fields from edges 1 and 2

EGO: geometric optics field

ET: total field (EGO + ED1 + ED2)

F: frequency

FN: angular size of wedge (FN = 2 for the limit
where the wedge becomes a plane)

K: wave number

KA: $KA = 2\pi/\lambda a \sqrt{\epsilon_r}$, where a is the radius
of the microstrip disk

RCF, UCF: real and imaginary parts of diffraction
coefficient

TH: theta

```

PROGRAM GTD
-----
*
*           GTD.FOR
*
*   Purpose:  Uses Geometric theory of diffraction to model a
*             circular microstrip disk on a finite square ground
*             plane.
*
*-----

* Variables
  DOUBLE PRECISION RCF, UCF, ANG, FN
  DOUBLE PRECISION F, K, PI, TH, D1, D2
  DOUBLE PRECISION ARG, FMAX
  DOUBLE PRECISION C1, C2, C3
  DOUBLE PRECISION EGO(-180:180), KA
  COMPLEX*16 DI
  COMPLEX*16 ED1(-180:180), ED2(-180:180), ED(-180:180)
  COMPLEX*16 ET(-180:180)
  DOUBLE PRECISION ETP(-180:180)
  COMPLEX*16 J
  REAL D

  OPEN (UNIT = 50, FILE = 'GTD.DAT', STATUS = 'OLD')

* Constants
  PI = DACOS(-1.0DG)
  J = (0.0, 1.0)

  D = 1E-3
  IER = 10

* In wavelengths

* For Centre-fed Disk
  D1 = 1.54
  D2 = 1.54
  F = 1.850E9

* For Slotted Disk (1.0 cm slot spacing)
*   D1 = 1.17
*   D2 = 1.17
*   F = 1.40E9

* For Slotted Disk (0.5 cm slot spacing)
*   D1 = .979
*   D2 = .979
*   F = 1.175E9

* For Slotted Disk (0.25 cm slot spacing)
*   D1 = 0.75
*   D2 = 0.75
*   F = .900E9

* For Shorted Disk
*   D1 = 0.358
*   D2 = 0.358
*   F = 0.430E9

```

```

      K = 2 * PI / (3E8 / F)

      DO I = -180, 180
        EGO(I) = 1E-5
      END DO

*   Er = 2.5
      KA = 0.39 * 2 * PI
      FMAX = 0.0

*   Calculate E-Fields from Geometric Optics
      DO I = -90, 90, 3
        TH = I * PI / 180.0
        ARG = DABS(KA * DSIN(TH))
        CALL BESJ(ARG, 1, EGO(I), D, IER)
        EGO(I) = CABS(EGO(I))
        IF (FMAX .LE. EGO(I)) FMAX = EGO(I)
      END DO

*   Compute diffracted E-Fields using GTD

*   FN = 2 for a plane
      FN = 2
      CALL BESJ(KA, 1, C1, D, IER)
      C2 = -1 / (SQRT(D1 / K * 2 * PI) * (2 * FN * SQRT(2 * PI * K)))
      C3 = -1 / (SQRT(D2 / K * 2 * PI) * (2 * FN * SQRT(2 * PI * K)))

      DO I = -180, 180, 3
        TH = I * PI / 180.0

*   Compute diffracted field at Edge 1
        IF (I .LT. 90) THEN
          ANG = TH + PI / 2.0
        ELSE
          ANG = TH - 3 * PI / 2.0
        END IF
        CALL CF(RCF, UCF, D1, ANG, FN)
        DI = CMPLX(RCF, UCF)
        ARG = 2 * PI * (D1 - D1 * DSIN(TH)) + PI / 4.0
        ED1(I) = C1 * C2 * DI * CEXP(-J * ARG)

*   Compute diffracted field at Edge 2
        IF (I .LT. (-90)) THEN
          ANG = -3 * PI / 2.0 - TH
        ELSE
          ANG = PI / 2.0 - TH
        END IF
        CALL CF(RCF, UCF, D2, ANG, FN)
        DI = CMPLX(RCF, UCF)
        ARG = 2 * PI * (D2 + D2 * DSIN(TH)) + PI / 4.0
        ED2(I) = C1 * C3 * DI * CEXP(-J * ARG)

*   Total Diffracted Field
        ED(I) = ED1(I) + ED2(I)

      END DO

      FMAX = 0.0
      DO I = -180, 180, 3
        ET(I) = EGO(I) + ED(I)
        ETP(I) = CABS(ET(I))
        IF (FMAX .LT. ETP(I)) FMAX = ETP(I)
      END DO

      DO I = -180, 180, 3
        ETP(I) = 20 * ALOG10(ETP(I) / FMAX)
        WRITE (50,10) I, ETP(I), EGO(I), CABS(ED(I)), CABS(ET(I))
10      FORMAT (1X, 15, 3X, 4(F9.3, 2X))
      END DO

      CLOSE (50)
      STOP
      END

```

```
SUBROUTINE CF (RCF, UCF, R, ANG, FN)
```

```

* -----
* Purpose: This subroutine computes the CF product used to
*          determine the edge diffraction coefficient
*
* Source:  "Antenna Theory Analysis and Design", by C.A. Balanis.
*          Harper and Row, New York, 1982, pp 529-531
* -----

```

```

DOUBLE PRECISION RCF, UCF, R, ANG, FN
DOUBLE PRECISION RAG, DP, TSIN, PI
DOUBLE PRECISION C, S, BOTL
COMPLEX*16 TOP, EXP, UPPI, UNPI, J

PI = DACOS(-1.0D0)
J = (0.0, 1.0)
TPI = 2.0 * PI
TOP = J * 2.0 * DSQRT(TPI * R)
N = IFIX((PI + ANG) / (2.0 * FN * PI) + 0.5)
DN = FLOAT(N)
A = 1.0 + DCOS(ANG - 2.0 * FN * PI * DN)
BOTL = DSQRT(TPI * R * A)
EXP = CEXP(J * TPI * R * A)
CALL FRNELS(C, S, BOTL)
C = DSQRT(PI / 2.0) * (0.5 - C)
S = DSQRT(PI / 2.0) * (S - 0.5)
RAG = (PI + ANG) / (2.0 * FN)
TSIN = DSIN(RAG)
TS = ABS(SNGL(TSIN))
X = 10.0
Y = 1.0 / X ** 5
IF (TS .GT. Y) THEN
    DP = DSQRT(A) * DCOS(RAG) / TSIN
    COMP = SNGL(DP)
ELSE
    COMP = -DSQRT(2.0) * FN * DSIN(ANG / 2.0 - FN * PI * DN)
    IF (DCOS(ANG/2.0-FN*PI*DN) .LT. 0.0) COMP = -COMP
END IF
UPPI = TOP * EXP * COMP * CMPLX(C,S)
N = IFIX((-PI + ANG) / (2.0 * FN * PI) + 0.5)
DN = FLOAT(N)
A = 1.0 + DCOS(ANG - 2.0 * FN * PI * DN)
BOTL = DSQRT(TPI * R * A)
EXP = CEXP(J * TPI * R * A)
CALL FRNELS(C, S, BOTL)
C = DSQRT(PI / 2.0) * (0.5 - C)
S = DSQRT(PI / 2.0) * (S - 0.5)
RAG = (PI - ANG) / (2.0 * FN)
TSIN = DSIN(RAG)
TS = ABS(SNGL(TSIN))
IF (TS .GT. Y) THEN
    DP = DSQRT(A) * DCOS(RAG) / TSIN
    COMP = SNGL(DP)
ELSE
    COMP = SQRT(2.0) * FN * DSIN(ANG / 2.0 - FN * PI * DN)
    IF (DCOS(ANG/2.0-FN*PI*DN) .LE. 0.0) COMP = -COMP
END IF
UNPI = TOP * EXP * COMP * CMPLX(C,S)
RCF = REAL(UPPI + UNPI)
UCF = AIMAG(UPPI + UNPI)
RETURN
END

```

SUBROUTINE FRNELS(C, S, XS)

```

*
* Purpose: This subroutine computes the Fresnel integral based on
* the algorithm by J. Boersma in "Computation of Fresnel
* Integrals", J. Math. Comp., vol 14, p. 380, 1960
*
* Source: "Antenna Theory Analysis and Design", by C.A. Balanis.
* Harper and Row, New York. 1982, pp 529-531
*

```

DOUBLE PRECISION C, S, XS
 DOUBLE PRECISION A(12), B(12), CC(12), D(12)

```

A(1)  =  1.595769140
A(2)  = -0.000001702
A(3)  = -6.808568854
A(4)  = -0.000576361
A(5)  =  6.920691902
A(6)  = -0.016898657
A(7)  = -3.050485660
A(8)  = -0.075752419
A(9)  =  0.850663781
A(10) = -0.025639041
A(11) = -0.150230960
A(12) =  0.034404779
B(1)  = -0.000000033
B(2)  =  4.255387524
B(3)  = -0.000092810
B(4)  = -7.780020400
B(5)  = -0.009520895
B(6)  =  5.075161298
B(7)  = -0.138341947
B(8)  = -1.363729124
B(9)  = -0.403349276
B(10) =  0.702222016
B(11) = -0.216195929
B(12) =  0.019547031
CC(1)  =  0.0
CC(2)  = -0.024933975
CC(3)  =  0.000003936
CC(4)  =  0.005770956
CC(5)  =  0.000689892
CC(6)  = -0.009497136
CC(7)  =  0.011948809
CC(8)  = -0.006748873
CC(9)  =  0.000246420
CC(10) =  0.002102967
CC(11) = -0.001217930
CC(12) =  0.000233939
D(1)  =  0.199471140
D(2)  =  0.000000023
D(3)  = -0.009351341
D(4)  =  0.000023006
D(5)  =  0.004851466
D(6)  =  0.001903218
D(7)  = -0.017122914
D(8)  =  0.029064067
D(9)  = -0.027928955
D(10) =  0.016497308
D(11) = -0.005598515
D(12) =  0.000838386

```



```

IF (XS .LE. 0.0) THEN
  C = -0.0
  S = -0.0
ELSE
  X = XS
  X = X * X
  FR = 0.0
  FI = 0.0
  IF (X .LT. 4.0) THEN
    Y = X / 4.0
    DO K = 12, 2, -1
      FR = (FR + A(K)) * Y
      FI = (FI + B(K)) * Y
    END DO
    FR = FR + A(1)
    FI = FI + B(1)
    C = (FR * DCOS(X) + FI * DSIN(X)) * DSQRT(Y)
    S = (FR * DSIN(X) - FI * DCOS(X)) * DSQRT(Y)
  ELSE
    Y = 4.0 / X
    DO K = 12, 2, -1
      FR = (FR + CC(K)) * Y
      FI = (FI + D(K)) * Y
    END DO
    FR = FR + CC(1)
    FI = FI + D(1)
    C = 0.5 + (FR * DCOS(X) + FI * DSIN(X)) * DSQRT(Y)
    S = 0.5 + (FR * DSIN(X) - FI * COS(X)) * SQRT(Y)
  END IF
END IF

RETURN
END

```

SUBROUTINE BESJ(X, N, BJ, D, IER)

```

*-----*
*
*               SUBROUTINE BESJ
*
* PURPOSE:
*   Computes the J Bessel Function for a given argument and order.
*
* USAGE:
*   CALL BFSJ(X,N,BJ,D,IER)
*
* PARAMETERS:
*   X      Argument of the J Bessel Function
*   N      Order of the J Bessel Function
*   BJ     Resultant
*   D      Required Accuracy
*   IER    Error Code :      IER = 0      No Error
*                           = 1      N is negative
*                           = 2      X is negative or zero
*                           = 3      Range not correct
*
* REMARKS:
*   Conditions on N:
*
*               N >= 0 and
*               N < 20 + 10 * X - X**2 / 3   (for X <=15)
*               N < 90 - X / 2              (for X > 15)
*
*   Source: "Antenna Theory Analysis and Design", by C.A. Balanis.
*           Harper and Row, New York. 1982, pp 202
*-----*

```

```

DOUBLE PRECISION X, BJ
REAL  BPREV, FM1, FM, ALPHA, BMK, S
INTEGER NTEST, N1, MA, MB, MZERO, MMAX, M
INTEGER JT, M2, K, MK
*
* Check the limits of N and X
*
  BJ = 0.0

  IF (N .LT. 0) THEN
    IER = 1
    RETURN
  END IF

  IF (X .LE. 0) THEN
    IER = 2
    RETURN
  END IF

  IF ((X - 15.0) .LE. 0) THEN
    NTEST = 20.0 + 10.0 * X - X ** 2 / 3
  ELSE
    NTEST = 90.0 + X / 2.0
  END IF

  IF ((N - NTEST) .GE. 0) THEN
    IER = 3
    RETURN
  END IF

```

```

IER = 0
N1 = N + 1
BPREV = 0.0
.
.
Compute starting value of M
.
IF ((X - 5) .LT. 0) THEN
    MA = X + 6.0
ELSE
    MA = 1.4 * X + 60.0 / X
END IF

MB = N + IFIX(X) / 4 + 2
MZERO = MAX0(MA, MB)
.
.
Set upper limit of M
.
MMAX = NTEST

DO M=MZERO, MMAX, 3
    FM1 = 1.0E-28
    FM = 0.0
    ALPHA = 0.0

    IF ((M-(M / 2) * 2) .EQ. 0) THEN
        JT = -1
    ELSE
        JT = 1
    END IF

    M2 = M - 2

    DO K = 1, M2
        MK = M - K
        BMK = 2.0 * FLOAT(MK) * FM1 / X - FM
        FM = FM1
        FM1 = BMK
        IF ((MK-N-1) .EQ. 0) BJ = BMK
        JT = - JT
        S = 1 + JT
        ALPHA = ALPHA + BMK * S
    END DO

    BMK = 2.0 * FM1 / X - FM
    IF (N .EQ. 0.0) BJ = BMK
    ALPHA = ALPHA + BMK
    BJ = BJ / ALPHA

    IF ((ABS(BJ - BPREV) - ABS(D * BJ)) .GT. 0) THEN
        BPREV = BJ
    ELSE
        RETURN
    END IF

END DO

IER = 3
END

```

APPENDIX 2Cross Polarization to Axial Ratio Conversion

$$AR = 20 \log((1 + \chi) / (1 - \chi))$$

where: $\chi = 10^{(x/20)}$

x = Cross Polarization Level (dB)

AR = Axial Ratio (dB)

Sample Calculation:

Cross Polarization Level = x = -21 dB

$$\chi = 10^{(x/20)} = 0.0891$$

$$\begin{aligned} AR &= 20 \log((1 + \chi) / (1 - \chi)) \\ &= 20 \log((1 + 0.0891) / (1 - 0.0891)) \\ &= 1.55 \text{ dB} \end{aligned}$$

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END

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FIN