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**A STACKED MICROSTRIP ANTENNA
FOR MOBILE SATELLITE COMMUNICATION.**

by

Kevin Roy McKay, B.Sc.E.

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

Master of Engineering

Department of Electronics

Carleton University

Ottawa, Ontario

October 25, 1993

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"A Stacked Microstrip Antenna for Mobile Satellite Communication"

submitted by Kevin Roy McKay, B.Sc.E.

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

Thesis Supervisor

Chair, Department of Electronics

Carleton University

November 23, 1993

ABSTRACT

This thesis investigates the technique of stacking rectangular microstrip antennas to increase their impedance bandwidth. In the particular configurations studied, the lower patch is fed directly while the upper patch is allowed to act as a parasitic element. A novel stacked configuration consisting of end-shortened microstrip patches is examined in terms of the input impedance and coupling between the driven and parasitic patches. In addition, the effect of the relative sizes and offset between stacked open patches is presented.

A circularly-polarized radiating element consisting of an offset stack of open microstrip patches has been optimized for use as a land-mobile satellite communication antenna. This element is used in a phased array in which the beam is fixed at the midpoint of the required coverage region. The array can be conformally mounted on the roof of vehicles and is designed to be mechanically rotated in azimuth only.

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LIST OF PRINCIPAL SYMBOLS

SYMBOL	DESCRIPTION
α_c	attenuation constant for conductor loss
α_d	attenuation constant for dielectric loss
β	electrical phase shift between elements in an array
ϵ_e	effective dielectric constant
ϵ_r	relative dielectric constant
f_r	resonant frequency
η_0	intrinsic impedance of free space ($\approx 377 \Omega$)
I_n	current amplitude of the n th array element
k	wave number
L	length of a microstrip line
λ_0	wavelength in free space
λ_g	wavelength in guided transmission medium
ω	angular frequency
σ	conductivity
Q	quality factor
W	width of a microstrip line
Z_m	characteristic impedance of a microstrip transmission line
h	dielectric substrate thickness
$\tan\delta$	dielectric loss tangent
VSWR	voltage standing wave ratio

1. INTRODUCTION

1.1 Introduction

The use of microstrip antennas has grown rapidly in the past decade. They have been successfully employed in applications including cellular and satellite communication, microwave landing systems, radar, and biomedical radiators. The growth in their application can be attributed to the low profile, light weight, simplicity of construction, and low cost offered by microstrip technology. For certain applications the antenna and feed can be etched on a single layer resulting in a compact and easily fabricated structure. The antenna can also be made conformal to mounting structures such as aircraft fuselage to yield an aerodynamic design or to buildings to provide an aesthetic appearance within the surrounding environment. However, microstrip antennas have several important performance limitations that restrict their application. Patch antennas have a narrow bandwidth of operation limited to a few percent (typically 1-5) [1]. The bandwidth can be increased by choosing a thicker substrate or by selecting a substrate with a lower dielectric constant. This requires that the element dimensions be increased for a given resonant frequency causing a narrowing of the antenna beamwidth. Also, the antenna gain is limited by losses and there is some difficulty in isolating feed lines from the radiating patch.

This thesis describes work directed toward increasing the bandwidth of microstrip antennas without significantly increasing the element dimensions or decreasing the pattern beamwidth. A stacked configuration with increased bandwidth has been investigated and is demonstrated in an antenna array for mobile satellite communication from ground vehicles.

1.2 Objectives

This thesis investigates a technique for increasing the bandwidth of microstrip antennas. Rectangular microstrip elements are stacked by placing a parasitic patch above a standard microstrip antenna. The lower driven element is fed at its edge with a microstrip transmission line. The upper patch is not directly connected, but rather, acts as a parasitic element by coupling energy from the radiating edges of the lower element. This technique has been applied to the novel configuration of end-shortened as well as open microstrip patches.

A study has been conducted to determine the effect of varying several physical parameters of the stacked structure including the relative sizes and offset between the two patches. This serves to illuminate the nature of the coupling and the impedance behaviour of such antennas. These observations were used to develop a stacked, circularly-polarized antenna with increased impedance bandwidth. This antenna was then used as an element in a planar phased array antenna for mobile satellite communication.

1.3 Organization

This thesis is divided into three major sections. The first section provides background information on mobile satellite communication and discusses the bandwidth limitations which currently prevent low-profile microstrip technology from being employed in the application. Chapter 2 provides an overview of the requirements on antennas for use in land-mobile satellite communication. Some of the antennas recently developed for this application are discussed. Chapter 3 concentrates on the impedance bandwidth limitation suffered by microstrip antennas and discusses some of the conventional techniques that have been employed to enhance the range of operation. A

review of work involved with stacking microstrip antennas is also presented and the major findings are outlined.

The second section is involved with an experimental investigation of stacking applied to end-shortened and open microstrip antennas. Chapter 4 looks at end-shortened microstrip patches as a potential antenna for mobile satellite communication because they exhibit an increased bandwidth in the E-plane. A stacked arrangement is examined and potential drawbacks to this configuration are discussed. Chapter 5 describes the impedance behaviour of stacked, half-wavelength microstrip antennas and looks at the effect of patch offset and relative patch sizes on coupling between the layers. A circularly polarized antenna is optimized for use as an element in an array for mobile satellite communication.

The third section covered in Chapter 6 outlines the design and testing of a phased array. The antenna is enclosed in a disk structure and provides a fixed elevation beam and can be rotated to provide azimuthal coverage.

Chapter 7 provides a summary of the work undertaken in this thesis and draws conclusions and recommendations for future work.

2. MOBILE SATELLITE COMMUNICATION

2.1 Introduction

In recent years there has been a significant effort undertaken to develop a communications system that is accessible regardless of location. While international telephone networks provide access to many regions there are still many areas without telephone service. Although cellular networks can provide service in cities where systems have been implemented, they cannot meet communication requirements where such systems are either not feasible or have not yet been developed. In addition, some businesses such as trucking companies must maintain contact between the home office and drivers who cover large distances. As a result, there is a large potential for satellite links to fill gaps in a worldwide mobile communication system.

2.2 Antenna Requirements

Mobile satellite communication systems are currently being developed to provide access from landborne, airborne and maritime platforms. Each platform has its own unique requirements for an antenna. Aeronautical systems require either a conformal antenna to meet aerodynamic requirements or an antenna which can be installed inside the aircraft fuselage. Mobile antennas developed for ships must be capable of coping with significant platform motion in the elevation plane. Antennas intended for application on ground vehicles, which are the focus of this work, must primarily be compact and conformal to the vehicle to meet both aesthetic and aerodynamic requirements.

The antenna can provide complete azimuth coverage through either mechanical or electrical beam steering. A motor and drive system can be used to rotate the antenna

about its centre on a rotary joint through which the RF connection is made. Pancake drive motors are available which offer high reliability and are relatively inexpensive. The alternative is to employ an electronically-steered array which may use phase scanning or beam switching and eliminates the need for an external drive system. Such antennas are more complex than mechanically-steered antennas because they must incorporate a feed structure in which various phase delays can be switched in and out. However, since there are no moving parts they tend to be highly reliable. Thus the selected method of azimuth steering is a trade-off between mechanical complexity for mechanically-steered antennas and feed network complexity for electronically-steered antennas.

Land-mobile antennas are typically required to provide coverage from 20 to 60 degrees above the horizon in order to allow operation at most latitudes. This requirement favours the use of an antenna which has a broad elevation beamwidth. Alternatively, the feed network can be designed to provide multiple beams in the elevation plane using variable phases between elements which can be turned on and off. The beam that provides the highest gain for a particular latitude is switched on.

Finally, the antenna must have a wide bandwidth to allow operation in the frequency bands 1530-1559 MHz for receive and 1626.5-1660.5 MHz for transmit. This bandwidth requirement limits the element types that are suitable for the application. For example, microstrip antennas appear to be well suited for ground-mobile communication because their low-profile allows them to be mounted conformally with vehicle rooftops. However, their use is limited by a narrow bandwidth of operation. The remaining chapters of this thesis discuss techniques for overcoming this bandwidth limitation to develop a low-profile antenna for the application. This chapter summarizes other antenna configurations which have been developed for land-mobile satellite communication.

2.3 Current Technology

Chung *et al.* [2] developed a low profile electronically-steered array using a cavity-backed printed crossed-slot radiating element. The crossed-slot was etched on a single-sided copper substrate and plated-through holes were used to form a cavity wall as shown in Figure 2.1. Circular polarization was obtained by feeding the crossed-slot in phase quadrature and three-bit diode phase shifters were used to steer the antenna beam. The array consisted of 19 elements arranged in a triangular lattice with a diameter of 55.9 cm and a total thickness of 1.9 cm excluding the RF connector. A total of four layers were used to construct the array including: (1) the top crossed-slot board, (2) the stripline feeder board to excite the radiators, (3) the corporate feed network board, and (4) the bottom ground plane. The measured gain ranged from 8.3 dBic at 1660 MHz to 9.3 dBic at 1545 MHz at an elevation angle 20 degrees above the horizon.

Schmidt [3,4] developed an electronically-steered array consisting of electromagnetically-coupled circular disk patches. Each element is similar to the structure shown in Figure 3.8b which employs an air gap between the two layers resulting in a thicker antenna. The feed network, which was implemented using stripline on a layer below the radiating elements, and the large number of PIN diodes resulted in a very expensive design. Nineteen patches were arranged in a triangular lattice structure with three-bit phase shifters used to achieve beam steering. The antenna scans a pattern of 48 beam positions spaced 15 degrees apart in azimuth with two elevation positions. The gain ranged from 13.4 dBic at 60 degrees elevation to 8-9 dBic at 20 degrees elevation. The array had a diameter of 61 cm and a height of 3.3 cm.

Densmore and Huang [5-7] developed an array of microstrip patches that were mechanically rotated using a belt drive. The patches were utilized in the same manner in which dipole elements are used in the classical Yagi-Uda array to obtain an increased gain.

Figure 2.1: Crossed-slot radiating element. After [2].

Four sets of Yagi antennas are used as shown in Figure 2.2a to reduce the azimuth beamwidth. The reflector element has a lower resonant frequency than the two director elements. Variation in the size of the parasitic reflector and director patches from the resonant dimension add a reactance component to their impedances and hence introduce an additional phase component to the radiation field. The combination of this phase with the coupling phase serve to tilt the antenna beam from its broadside direction to obtain better coverage at lower elevation angles. However, the elevation radiation pattern suffers from a high backlobe as shown in Figure 2.2b. In order to obtain a sufficient degree of coupling the elements are spaced by less than one substrate thickness in the E-plane. In addition, the driven element is fed in two locations to provide circular polarization. Elevation coverage extended from 20 to 60 degrees above the horizon with a gain of more than 10 dBic. The entire structure has an outer diameter of 56 cm and a height of 3.8 cm and is intended to be mounted conformally to the rooftop of cars and trucks.

Petosa [8] developed a cylindrical patch array as shown in Figure 2.3 consisting of an array of 18 circularly polarized microstrip elements. In this case sufficient bandwidth for the microstrip elements was obtained by using an air dielectric (honeycomb substrate) with a thickness of 1.25 cm. The antenna beam was fixed in the elevation plane to 30 degrees above the horizon by mounting the elements on the surface of a cylindrical cone. The beam is steered electronically in azimuth in 10 degree increments by turning on either 4 or 5 of the elements using PIN diodes. In addition, a one-bit phase shifter provides a phase shift to the central 2 or 3 elements that are turned on in order to narrow the beamwidth and reduce sidelobe levels. The array has a height of 11 cm and the diameter at the mounting surface is 61 cm. An excellent axial ratio of better than 1.5 dB was achieved for the array while the peak gain ranged from 11.3 dBic to 12.3 dBic.

(a)

(b)

Figure 2.2: (a) Planar microstrip Yagi array. (b) Measured elevation pattern at 1654 MHz. From [7].

Figure 2.3: Cylindrical patch array. From [8].

2.4 Conclusions

Effort is still being directed towards improving the state-of-the-art despite the recent progress made in the development of antennas for land-mobile satellite communication. Lower profile antennas are required for installation on typical cars and trucks while the gain of larger volume designs must be maintained or improved. The next chapter will describe the limitations which prevent the application of ordinary microstrip technology to the problem and investigates electromagnetic coupling between stacked patches as a potential solution.

3. BANDWIDTH ENHANCEMENT

3.1 Radiation from a Rectangular Microstrip Antenna

The radiation mechanism of a microstrip antenna is well documented in the literature [e.g., 1] and will only be briefly discussed here. A microstrip antenna consists of a dielectric slab that has a radiating conductor on one side and a ground plane on the other. The upper patch is usually fed by either a microstrip line as shown in Figure 3.1 or by a coaxial connector in which the centre conductor protrudes through the dielectric material and is soldered to the patch. The fields are mostly confined between the ground plane and the resonant conductor along its length. However, at the ends of the patch a non-ideal open circuit exists from which not all of the energy is reflected. Instead, fringing fields occur at the ends causing some of the energy to be radiated. Along the resonant length (corresponding to approximately one-half wavelength) the electric field will vary as shown in Figure 3.2. The fields at the end of the patch can be decomposed into normal and tangential components. For a half-wavelength patch, the components that are normal to the plane of the conductors cancel while the tangential components add in phase to give a maximum in the plane of the conductors.

3.2 Microstrip Antenna Analysis Techniques

Numerous models have been developed for the analysis and synthesis of microstrip radiators. Numerical techniques employing the moment method offer a high degree of accuracy. These techniques are highly beneficial for applications utilizing expensive materials or processing or when tuning of the structure is difficult. Other techniques, such as the transmission-line model or the modal-expansion cavity model, are based on various

Figure 3.1: Rectangular microstrip antenna fed with a microstrip transmission line.

Figure 3.2: Rectangular microstrip patch antenna showing approximate electric field distribution: (a) side view (b) top view. After [1, p.6].

assumptions which serve to simplify the required calculations at the expense of less accurate solutions.

3.2.1 Transmission-Line Model

The transmission-line model [9] provides the simplest description for patch antennas. This analysis method is largely restricted to the case of rectangular patches. Here the structure is viewed as a transmission line with a radiating slot at each end. For a typical patch the slots are spaced 180 degrees apart electrically at resonance. This is achieved by adjusting the resonant length L of the patch to be slightly less than half of the guided wavelength λ_g to account for the fringing fields at the radiating edges. Thus, the reactive components of each slot will cancel when transformed to one edge of the patch as shown in Figure 3.3. The input impedance at resonance is determined by simply doubling the radiation conductance of each slot.

3.2.2 Modal Expansion Cavity Model

The modal expansion technique, as described in [9] has been shown to provide a reasonably accurate prediction of the input impedance and resonant frequency for patch antennas on electrically thin substrates. This technique views the patch radiator as a thin cavity enclosed by leaky magnetic walls. An impedance boundary condition is imposed at the radiating walls by introducing a complex wall admittance which accounts for the externally radiated and stored energy. Provided the substrate is electrically thin, it can be assumed that the electric field exists parallel to the walls (i.e. z -directed) with no components tangential to the ground plane and is given by [1, p.45]

Figure 3.3: Transmission-line model of rectangular microstrip antenna. After [9].

$$E_z(x, y) = \sum_m \sum_n \frac{A_{mn} \chi_{mn}}{\sqrt{\epsilon W L h}} \cos(k_x x) \cos(k_y y)$$

$$\chi_{mn} = \begin{cases} 1, & m = 0 \text{ and } n = 0 \\ \sqrt{2}, & m = 0 \text{ or } n = 0 \\ 2, & m \neq 0 \text{ and } n \neq 0 \end{cases} \quad (\text{Eqn. 3.1})$$

where A_{mn} are the amplitude coefficients for each mode. For the TM_{mn} mode, the resonant wave number k_{mn} is given by

$$k_{mn}^2 = \omega_{mn}^2 \mu \epsilon = k_x^2 + k_y^2 \quad (\text{Eqn. 3.2})$$

The most commonly excited mode is TM_{10} as shown in Figure 3.4 in which the electric field varies only along the length of the patch so that $k_y = 0$. In order to find the resonant frequency the value for k_x must be determined. The complex wave number is found by lumping the losses into the conductance of the radiating walls and then imposing the boundary conditions. For thin substrates the complex transcendental equation (Eqn. 3.3) [9] can be used to extract the wave number k_{10} .

$$\tan(k_{10} L) = \frac{2 k_{10} \alpha_{10}}{k_{10}^2 - \alpha_{10}^2} \quad (\text{Eqn. 3.3})$$

where

$$\alpha_{10} = j \frac{2 \pi Z_o}{\lambda_o} \frac{h}{W} Y_w F_y \quad (\text{Eqn. 3.4})$$

in which $Y_w = G_w + jB_w$ is the admittance of the radiating walls with

$$G_w = \frac{\pi}{Z_o} \frac{W}{\lambda_o} \quad (\text{Eqn. 3.5})$$

and

Figure 3.4: Electric field and magnetic surface current distributions in walls for various modes of a rectangular microstrip patch antenna. From [1, p.43].

$$B_w = 0.01668 \frac{\Delta l}{h} \frac{W}{\lambda_0} \epsilon_e \quad (\text{Eqn. 3.6})$$

where ϵ_e is an effective dielectric constant that has a value between that of the substrate material ϵ_r and the dielectric constant of the surrounding medium which is typically air ($\epsilon_r = 1$) and is given by [1, p. 46]

$$\epsilon_e = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left(1 + \frac{12h}{W} \right)^{-1/2}. \quad (\text{Eqn. 3.7})$$

$\Delta l/h$ is a normalized line extension which compensates for the apparent increase in the resonant length of a microstrip antenna caused by the fringing fields at the radiation edges with [1, p. 46]

$$\frac{\Delta l}{h} = 0.412 \frac{(\epsilon_e + 0.3)(W/h + 0.264)}{(\epsilon_e - 0.258)(W/h + 0.8)} \quad (\text{Eqn. 3.8})$$

The expression for Y_w is based on a wave normally incident on a parallel-plate TEM waveguide slot radiating into a half-space. F_y is an empirically derived aspect ratio factor which accounts for the fact that the admittance of the walls are also dependent on the angle of incidence. Carver [9] derived this expression by near-field probing close to the wall.

$$F_y(W/L) = 0.7747 + 0.5977(W/L - 1) - 0.1638(W/L - 1)^2 \quad (\text{Eqn. 3.9})$$

An iterative algorithm was developed [9] to approximate the solution to the transcendental equation 3.3 to find k_{10} . It is obtained by expanding the first two terms of the Taylor series for $\tan(k_{10}L)$ about π such that

$$k_{10} = (\pi / L) - (\Delta_4 / L) \quad (\text{Eqn. 3.10})$$

where

$$\Delta_{p+1} = \frac{2(\alpha_{10}L)(\pi - \Delta_p)}{(\alpha_{10}L)^2 + 2\Delta_p\pi - \Delta_p^2 - \pi^2} - \frac{\Delta_p^3}{3} \quad (\text{Eqn. 3.11})$$

with $\Delta_0 = 0$ for an initial value.

Once the complex eigenvalue k_{10} has been determined the resonant frequency can be found from the relation [1, p. 46]

$$\omega = \omega_r + j\omega_i = \frac{c}{\sqrt{\epsilon_r(1 - j \tan \delta)}} k_{10} \quad (\text{Eqn. 3.12})$$

where $\tan \delta$ is the loss tangent of the dielectric.

The radiation resistance can be found from the radiation quality factor which is given by [9]

$$Q_r = \frac{\text{Re}(k_{10})}{2 \text{Im}(k_{10})} \quad (\text{Eqn. 3.13})$$

The radiation resistance at resonance referred to the input is [9]

$$R_r = \frac{Q_r}{\omega_r C_{10}} = \frac{2Q_r}{\omega_r C_{dc}} = \frac{2Q_r}{\omega_r (\epsilon L W / h)} \quad (\text{Eqn. 3.14})$$

where C_{dc} is the dc patch capacitance and C_{10} is the capacitance at resonance referred to the patch edge.

The FORTRAN program MICROANT.FOR is listed in Appendix 1. It uses the modal expansion technique described here to determine the resonant frequency, radiation resistance, quality factors, and the bandwidth for a microstrip antenna. The program is based largely on a program in [10] with modifications including the use of more recent empirical expressions and minor corrections.

The model for microstrip antennas that are fed using a coaxial probe must add an additional series reactive component to account for the inductance introduced by the probe. This inductive reactance can be approximated by [1, p. 55]

$$X_{probe} = \frac{\eta_o}{\sqrt{\epsilon_r}} \tan\left(\frac{2\pi h}{\lambda_o}\right). \quad (\text{Eqn. 3.15})$$

For thin substrates in which $h \ll \lambda_o$ the $\tan$ function in Eqn. 3.15 can be approximated by its argument. This leads to the realization that X_{probe} increases linearly with frequency and thus the probe can be accurately accounted for by an inductance with a value given by

$$L_{probe} = \frac{\mu_o h}{\sqrt{\epsilon_r}}. \quad (\text{Eqn. 3.16})$$

A sample comparison between measured and predicted performance was undertaken for the case of a square microstrip antenna with $L = W = 58.6$ mm, $\epsilon_r = 2.32$, and $h = 3.15$ mm. The predicted values were obtained using the MICROANT.FOR program and are listed in Table 3.1. The antenna was fed in the centre of a radiating edge using an SMA-type coaxial probe. The probe feed will introduce an additional inductive reactance in series with the antenna. For this case the inductor has a value of 2.61 nH. The structure, including the feed, can be effectively modeled as an inductance in series with a parallel RLC network which models the behaviour of the microstrip radiator near its resonant frequency [9] as shown in Figure 3.5. A Smith chart showing the comparison between predicted and measured impedance loci is provided in Figure 3.6. The correlation between the data points is quite good with the most significant discrepancy resulting from a low prediction for the resonant resistance.

$L_{\text{probe}} = 2.61 \text{ nH}$
 $R = 201 \text{ ohms}$
 $L = 0.878 \text{ nH}$
 $C = 11.2 \text{ pF}$

Figure 3.5: A probe fed microstrip antenna and its lumped-element circuit model derived from the modal-expansion cavity model.

Table 3.1: Predicted parameters for a microstrip antenna.

Parameter	Value
Resonant Frequency	1.606 GHz
Bandwidth (for VSWR < 2)	3.2 %
Resistance	201 Ω
Inductance	0.878 nH
Capacitance	11.2 pF

3.3 Bandwidth of a Rectangular Microstrip Antenna

Antenna performance can be described by a number of bandwidths which refer to various performance characteristics. There is no global definition for bandwidth since antenna characteristics (such as input impedance, polarization, and directivity) do not vary in the same way with frequency. The bandwidth of operation is often separated into the two main categories of impedance bandwidth and pattern bandwidth [11, p.47]. Pattern bandwidth refers to the range of frequencies over which the antenna radiation pattern meets certain criteria including sidelobe levels, beam pointing direction, gain, and polarization. The impedance bandwidth refers to the frequency range over which a suitable impedance match can be attained. The application of microstrip antennas is primarily limited by their narrow impedance bandwidth.

The impedance bandwidth is determined by the overall quality factor (Q) of the patch antenna. In fact, a microstrip antenna can be modeled, near its resonant frequency, as a parallel RLC network with a high quality factor [1, p.46]. This factor accounts for various losses including conductor, dielectric, and radiation.

The non-zero conductivity of the antenna substrate material will result in a small dielectric loss that increases with frequency. The attenuation constant due to the presence of the dielectric material is given by [12, p.26]

$$\alpha_d = 27.3 \frac{\epsilon_r (\epsilon_r - 1) \tan \delta}{\epsilon_r (\epsilon_r - 1) \lambda_g} \text{ [dB/unit length]} \quad (\text{Eqn. 3.17})$$

The Q-factor associated with this loss is given by [12, p. 76]

$$Q_d = \frac{\omega \epsilon}{\sigma_d} = \frac{1}{\tan \delta} \quad (\text{Eqn. 3.18})$$

Conductance loss arises from heating in the radiating patch and the ground plane. This is a more significant factor with an attenuation constant given by,

$$\alpha_c \propto \frac{R_s}{WZ_0} \quad (\text{Eqn. 3.19})$$

where R_s is the surface resistivity with

$$R_s = \sqrt{\frac{\omega \mu_0}{2 \sigma_c}} \text{ [\Omega per square]} \quad (\text{Eqn. 3.20})$$

The Q-factor which accounts for the conductance loss is expressed as [12, p.77]

$$Q_c = h \sqrt{\mu_0 \eta f \sigma_c} \quad (\text{Eqn. 3.21})$$

The total quality factor Q_t is given as a combination of the individual Q-factors through the relation

$$\frac{1}{Q_t} = \frac{1}{Q_r} + \frac{1}{Q_d} + \frac{1}{Q_c} \quad (\text{Eqn. 3.22})$$

where Q_c , Q_d , and Q_r are the Q-factors corresponding to conductor, dielectric, and radiation losses respectively. It is clear that for an antenna structure it is desirable to maximize the radiated power while minimizing the conductance and dielectric losses.

An impedance matching network is often implemented between the antenna and the input port. Assuming that a perfect match (i.e., VSWR = 1) is achieved at the resonant frequency, the bandwidth of an antenna with a feed line VSWR < s can be related to the total quality factor Q_t [1, p.62]

$$BW = \frac{s-1}{Q_t \sqrt{s}}. \quad (\text{Eqn. 3.23})$$

Of course if a poorer match is achieved the resulting bandwidth of operation will be degraded. The bandwidth of microstrip antennas is typically 1 - 5 percent [1, p. 2].

3.4 Conventional Techniques for Bandwidth Enhancement

The bandwidth of a microstrip antenna can be increased by increasing the substrate thickness since the Q-factor is inversely proportional to h . However, this is limited by the excitation of undesirable surface waves [1, pp. 26-27]. These waves propagate in the substrate outside the conducting microstrip patch. Coupling between the quasi-TEM wave propagating along the radiating patch and surface waves is strongest when their phase velocities are close. Thus, the upper frequency limit for the antenna is determined by the lowest frequency for surface wave coupling. Although the lowest order TM-type mode surface wave has no low order cutoff frequency (f_c), for the higher-order TM_n and TE_n modes f_c is inversely proportional to the thickness of the substrate and the square root of the dielectric constant. However, the cutoff frequencies for surface waves are sufficiently high for most configurations that they can often be neglected at L-band.

The bandwidth can be increased by decreasing the dielectric constant of the substrate material since Q_t is proportional to the square root of ϵ_r . This requires an increase in the resonant patch dimensions since the resonant length is inversely proportional to the square root of ϵ_r . Thus the volume occupied by the antenna will be increased and the beamwidth of the antenna will be reduced.

In addition, it is possible to modify the match between the antenna and its feed, thereby increasing the impedance bandwidth seen at the input to the matching network. The most common technique involves the use of impedance transformers and Tchebychev matching networks. In cases where there is a large impedance mismatch, higher-order matching networks must be employed. The stubs and transformers used can take up a significant amount of real estate on the circuit board, even exceeding the area occupied by the radiator itself.

Increased impedance bandwidth has been achieved with the use of a gap-coupled feed in which the feedline and patch radiator are separated by a small gap as shown in Figure 3.7a. The effect of this gap is to add a capacitive component which is used to cancel the reactance at the edge of the patch. However, because the required capacitance is usually very small, design is difficult and repeatability in the fabrication process is unreliable [12, pp. 70-71].

Pozar and Kaufman [13] developed a configuration in which proximity coupling was found to increase the bandwidth of a rectangular microstrip antenna. The feedline and patch are etched on different substrate layers with the feedline extending under the patch radiator to facilitate indirect coupling as shown in Figure 3.7b. Maximum coupling between the feedline and the patch was found to occur when the feedline was inset to half

Figure 3.7: Techniques for increasing the impedance bandwidth of microstrip antennas using: (a) capacitive gap-coupling, (b) proximity coupling through an inset feedline under the patch, and (c) gap-coupled microstrip antennas.

the length of the radiator. This configuration was modeled as a capacitor in series with a parallel RLC resonator. The RLC circuit models the behaviour of the patch resonator while the series capacitor accounts for the coupling between the feedline and the radiator.

Strickland [14] investigated two approaches for obtaining broadband radiating elements using transmission line coupled patches. In the first approach, filter theory used for the synthesis of high impedance - low impedance microstrip filters was applied to series-fed microstrip patch arrays of directly coupled rectangular resonators. A short section of a long series-fed array can be synthesized using filter theory to obtain the desired bandpass characteristic. Applying this to each section of an array ensures a desirable return loss characteristic for the complete array. The second approach for producing broadband radiating elements in a series-fed array involved making the radiator non-resonant by giving it the same characteristic impedance as the feedline. Since the radiators are non-resonant they are not required to be a half-wavelength long and the only discontinuities are the radiation conductances. The characteristic impedance of each section can be made equal through periodic loading using narrow capacitive stubs or by varying the substrate thickness.

Kumar and Gupta [15,16] developed a technique which employed the use of additional resonators that are gap-coupled to the edges of a rectangular patch. The first configuration used two resonators gap-coupled to the non-radiating edges of the centre element. The second configuration had four additional resonators at each of the four sides of the microstrip element as illustrated in Figure 3.7c. The resonant length of the coupled resonators is staggered with the result that the overall structure has increased bandwidth. Similar performance was also obtained for directly-coupled multiple resonator antennas in which a narrow microstrip line rather than a gap was used to connect the central radiator with the surrounding patches [17]. However, the radiation pattern characteristics for

these antennas vary significantly with frequency in terms of the direction of beam peak and half-power beamwidth.

Another technique which has been used for dual-frequency operation, instead of a single wide band, involves stacking resonators in a piggyback configuration. One such configuration consists of two microstrip antennas that are stacked above each other on a common ground plane [1, p.74]. The lower element is larger than the top element allowing it to serve as a ground plane. The antennas are fed separately to provide independent radiation.

3.5 Bandwidth Enhancement by Stacking Elements

Stacked microstrip antennas have been investigated in which only one of the elements is directly fed while the other acts as a parasitic element. This configuration holds several advantages over other techniques.

- The feeding mechanism is simpler than the piggyback configuration in which each element must be fed independently.
- Each element can utilize thin substrate layers to maintain a low-profile design and to minimize the excitation of surface waves.
- The configuration does not require an increase in the surface area occupied by the antenna as in the case of gap-coupled resonators.
- The designer is provided with several additional degrees of freedom including the relative sizes of the stacked patches, the thickness of the two substrate layers, and the offset between the element layers.

The next section presents a brief summary of work carried out on stacked radiators and some of the key findings. Experimental investigations by others have been carried out for patches of various shapes including rectangular [18-20], circular [21-24], triangular

[25], annular rings [26], and with numerous substrate topologies. It has been generally concluded that stacking radiators results in an increased bandwidth of operation over that of single-layer antennas.

3.5.1 Summary of Studies on Stacked Microstrip Antennas

Early work by Long and Walton [23] involved an experimental investigation of stacked circular-disc antennas in the frequency range 2-4 GHz. The intended purpose was to develop a microstrip antenna which would support operation at two distinct frequencies, each with a narrow bandwidth, rather than operating over one broad frequency band. Two feed configurations were studied.

In the first configuration, the top disc was fed by passing the inner conductor of a coaxial probe through a clearance hole in the lower element. The bottom disc radius was fixed while that of the upper disc was varied over the range 0.93 to 1.06 of the fixed value. This arrangement resulted in separations of about 10 percent in the resonant frequencies. The actual impedance bandwidth about each resonant frequency was only about 1 percent, similar to that of a single disc. It was determined that the lower resonant frequency was predominantly determined by the parameters of the lower disc. This allowed the upper frequency to be selected by adjusting the parameters of the upper disc, without significantly altering the lower frequency. Pattern measurements showed that no significant pattern deterioration results from the technique of stacking.

In the second configuration, Long and Walton [23] investigated the effect of feeding the bottom disc and allowing the upper patch to act as the parasitic element. The performance of this configuration as a dual-frequency antenna was poor. Only one resonant frequency was achieved with a slightly increased impedance bandwidth. However, the work of other researchers, summarized in the remainder of this section,

demonstrates that increased impedance bandwidth can be achieved by feeding the lower patch. In the particular configuration studied by Long and Walton it is possible that the diameter of the parasitic patch was too close to the diameter of the driven patch to allow sufficient coupling to the upper element.

Chen *et al.* [21] studied a stacked configuration operating in C-band and consisting of circular patches as shown in Figure 3.8a. The bottom patch was fed by a coaxial probe while the upper patch was electromagnetically coupled. This configuration was later called "inverted" by Dahele *et al.* [22] because the upper patch is etched on the bottom side of the upper substrate. The two discs were separated by a gap filled with foam to achieve the equivalent of an air dielectric. Experimental investigation showed that the upper resonant frequency decreases while the lower resonant frequency increases as the separation between the discs was increased. However, the lower resonant frequency remained relatively constant and was close to that predicted for the bottom disc alone. An impedance bandwidth (for 10 dB return loss or VSWR of 1.92) of 20.4 percent centred about 4.255 GHz was reported. The patches used in the investigation had varying diameters. It was found that by fixing the diameter of the bottom (fed) patch and the spacing between the two, an optimum diameter for the upper patch could be found empirically. The degree of circular polarity that could be achieved with the element was also studied by implementing several quadrature feed arrangements. Measurements showed that a high degree of purity in circular polarization was possible over a large bandwidth.

Two configurations of stacked circular discs were compared by Dahele *et al.* [22] in the frequency range 3-4 GHz. The first, inverted configuration, was similar to that studied by Chen *et al.* [21] in which the upper patch was etched on the bottom of the top substrate (Figure 3.8a). The second "normal" configuration was modeled after the work

Figure 3.8: Stacked microstrip antennas with an air gap separation between the elements. (a) inverted configuration (b) normal configuration. After [22].

of Sabban [24] with the upper patch etched on the top side of the upper substrate (Figure 3.8b). The empirical study used discs on 0.159 cm thick RT/Duroid substrate ($\epsilon_r = 2.3$). The upper patch was larger with a diameter of 1.85 cm while that of the lower fed patch was 1.65 cm. The air gap spacing between the patches was varied. Measurements showed that the resonant frequencies for the inverted configuration are higher than those of the normal configuration. Also, the impedance bandwidth for the inverted configuration was larger and had a larger radiation resistance.

Barlatey *et al.* [27,28] present a full-wave analysis of stacked microstrip patches using a mixed potential integral equation. The analysis is performed in the space domain and applies the method of moments. It requires complicated numerical techniques and mathematical formulations and only the results will be discussed here. The analysis was applied to two stacked microstrip configurations. The first structure consisted of stacked rectangular microstrip elements in the normal configuration with no air gap separation between the layers. The patches had the same width but differed slightly in their resonant length. The structure was found to behave as a dual frequency antenna with a frequency separation of 16 percent. The second configuration consisted of identically-sized rectangular microstrip antennas in the normal configuration with no air separation between the layers. The upper patch was slightly offset in one dimension from the bottom patch. The resulting structure was found to have an impedance bandwidth of 15 percent for a VSWR less than 2. Good agreement between measured and calculated performance was achieved.

Recent work by Lee and Lee [20] examined a different area in the behaviour of stacked structures. Here the concentration was on increasing the gain of microstrip antennas by stacking elements. The experimental configuration consisted of two rectangular patches. The bottom patch was driven from a coaxial probe while the upper

patch was electromagnetically coupled. The patches were fabricated on a Cufion substrate ($\epsilon_r = 2.17$). Both air ($\epsilon_r = 1$) and teflon ($\epsilon_r = 2.1$) dielectric separations (s) were investigated. Their experiment studied the effect of varying the dielectric separation between the patches. Three distinct regions of operation were found to correspond with various separations. The first region is characterized by a relatively large bandwidth and occurs for $s < 0.15\lambda_g$. This was primarily the region investigated earlier by other researchers. The second region has abnormal patterns, with a dip at broadside and occurs for $0.15 < s < 0.30\lambda_g$. The third region has high gain but narrow bandwidth, and exists for $s > 0.30\lambda_g$.

The research to date has shown a great deal of promise for the technique of stacking resonators to achieve an increased bandwidth of operation. Much of the work has been focused on configurations in which there is an air separation between the resonators. These have been found to yield much wider impedance bandwidths than can be achieved with ordinary microstrip technology.

3.5.2 Application to Mobile Satellite Communication

Mobile satellite communication from ground vehicles in L-band operates in two distinct bands (1.530-1.559 GHz for receive and 1.6265-1.6605 GHz for transmit). The separation between the bands is such that a thicker substrate microstrip patch with a low dielectric constant is required if an ordinary antenna is to be employed. However, the antenna thickness must be minimized to allow for conformal mounting on the roof of cars and trucks. In addition, increasing the substrate thickness and lowering the dielectric constant require that the resonant patch dimensions be increased causing the antenna beamwidth to narrow [1]. Since conformal mounting is required, the antenna beam must

be scanned off broadside to provide coverage at low elevation angles. A narrow beamwidth is undesirable since it limits the maximum scan angle that can be achieved.

Many of the stacked structures appearing in the literature have utilized a dielectric separation between the layers. While this gap tends to achieve the desired bandwidth behaviour, the overall thickness of the structure has to be significantly increased. As a result, a promising microstrip antenna for the land-mobile application is a stacked patch arrangement which maintains a low substrate thickness and minimizes the antenna dimensions, by decreasing or eliminating separations between the patches.

4. END-SHORTED MICROSTRIP ANTENNAS

4.1 Characteristics of End-Shorted Microstrip Antennas

End-shortened microstrip antennas provide an effective means of reducing the size of the patch while maintaining the same resonant frequency. An electric wall can be implemented along the centre line of a half wavelength patch without altering the boundary conditions. Thus, the length of the patch can be made approximately one quarter wavelength with a short between the patch and the ground plane on one edge (see Figure 4.1). The short circuit can be achieved with a continuous sheet or a number of shorting pins. The radiation mechanism is similar to that for an open microstrip antenna, except in the case of an end-shortened antenna radiation occurs at only one end of the patch. Having a single radiation zone results in two important differences between an end-shortened and an open microstrip antenna. One distinction is that end-shortened antennas have a larger E-plane beamwidth. Secondly, the radiation resistance for an end-shortened patch is significantly larger than that for an open microstrip radiator.

The mobile satellite communication application requires coverage over a range of elevation angles (typically 20 to 60 degrees above the horizon). This requirement favours the use of a radiation element which has a broad beamwidth in the elevation plane, such as the end-shortened microstrip antenna. As a result, the end-shortened patch was initially selected as an appropriate element for a low-profile antenna.

4.2 Elevation Beamwidth

Shorting one end of the patch results in a larger beamwidth than that of an opened $\lambda/2$ microstrip antenna. An open patch has radiation occurring at both ends and

Figure 4.1: (a) Side view of an end-shortened microstrip antenna. (b) Shorting posts used to create an effective short circuit on one edge of the radiating patch.

form an interferometer with reduced beamwidth. For the end-shortened $\lambda/4$ patch, radiation occurs at only one end. As a result, no interferometric-type reduction in beamwidth occurs and the E-plane beamwidth becomes more omnidirectional [12, p.103]. Figure 4.2 shows a comparison of elevation patterns for the open and end-shortened configurations.

4.3 Effect of Shorting Pins

Using a continuous sheet to achieve the end short requires a cut in the substrate material. This adds machining complexity and may weaken the rigidity of the antenna's mechanical structure. A suitable alternative is to use shorting pins. However, the spacing between the pins creates an inductance rather than a perfect short at the antenna edge. The effect of the edge inductance has been accounted for by assuming that there will be minimal fringing of the electric field at the short because of the low electric field intensity in this region. The shorted end of the element is then modeled as a series of posts across a parallel plate transmission line with an open circuit on one side. This added inductance requires that the patch resonant length L be shortened to achieve a particular resonant frequency. The effective position of the short extends beyond the pins by an amount [12, p.104]

$$\Delta L = \frac{a}{2\pi} \left\{ \ln \left(\frac{a}{2\pi r} \right) - \frac{4\pi^2 r^2}{a^2} + 0.601 \frac{a^2 \epsilon_r}{\lambda_o^2} \right\} \quad (\text{Eqn. 4.1})$$

where a is the spacing between shorting posts and r is the post radius as shown in Figure 4.1b. Since the effective short position is sensitive to pin spacing and diameter, these parameters affect the resonant length and hence frequency for the antenna. The detuning can be minimized by decreasing the spacing between the posts.

Figure 4.2: A comparison of the theoretical elevation plane radiation patterns for an open (---) and an end-shortened (—) rectangular microstrip antenna. In both cases the dielectric constant ϵ_r is 1 (air) and the patch width is $\lambda_0/2$.

4.4 Impedance Bandwidth for End-Shorted Microstrip Antennas

The bandwidth of operation for an end-shortened patch is one of the most critical considerations for its potential application as an element in a low-profile mobile antenna array. A useful comparison can be found between the bandwidths for open and end-shortened rectangular microstrip antennas by following the reasoning of James *et al.* [12]. The self-conductance of one end of an open-circuited microstrip antenna is given by [12,p.87],

$$G_r = \frac{1}{120\pi^2} F_2 \left(\frac{2\pi}{\lambda_o} w_e \right) \quad (\text{Eqn. 4.2})$$

where

$$F_2 = xS_1(x) - 2\sin^2(x/2) - 1 + \sin(x)/x. \quad (\text{Eqn. 4.3})$$

The radiation resistance at resonance of an open patch is then equal to the equivalent resistance of two single radiating slots in parallel because the radiating ends are electrically spaced by 180 degrees. As a result, the radiation resistance for the open patch is half that for a single open end slot. In addition, due to the close spacing between the slots (typically $\lambda_g/2$) the radiation resistance is reduced further due to mutual coupling between the slots. The mutual conductance term is given by the following expression [12, p.88],

$$G_m = \frac{1}{120\pi^2} \int_0^\pi \frac{\sin^2\left(\frac{\pi W_e}{\lambda_o} \cos\theta\right)}{\cos^2\theta} \sin^3(\theta) J_o\left(\frac{2\pi L \sin\theta}{\lambda_o}\right) d\theta \quad (\text{Eqn. 4.4})$$

It is obtained by integrating the interference component of the far-field radiation patterns of two magnetic current sources. These parallel sources have length W_e and are separated by a distance L . This expression for mutual conductance has been evaluated numerically and is plotted in Figure 4.3.

Figure 4.3: Mutual conductance between the radiating ends of a rectangular microstrip antenna for $W_e/\lambda_0 = 0.5$ (—) and $W_e/\lambda_0 = 1.0$ (---).

For an end-shortened radiator, the radiation conductance is simply given by the self-conductance for one end of an open-end microstrip radiator. To a first approximation, without considering mutual coupling effects, the radiation resistance of an end-shortened radiator is twice that for an open patch since there is only one slot. The absence of mutual coupling for the shortened patch increases the relative radiation resistance further over that for the open patch. For a half wavelength microstrip antenna the ratio of mutual conductance to self-conductance is typically 1:4 [12]. In total, the radiation resistance for an end-shortened microstrip radiator is approximately 2.5 times larger than for an open patch of twice the length.

When these factors are considered in evaluating the quality factor Q for a microstrip patch, it is possible to compare the bandwidths for the opened and shortened cases. Since the end-shortened patch has only one radiating end the stored energy will be half that of an open patch. The end-shortened patch has a radiation resistance 2.5 times larger than the open case causing the radiated power to be only 40 percent of the power radiated by an equivalent open patch, for a given applied voltage, since radiated power is proportional to V^2/R_r . The result is that the Q for an open radiator is 0.8 times that of an end-shortened patch as shown in Eqn. 4.5. Equivalently, the bandwidth of the shortened patch is reduced to 80 percent of that of the open patch.

$$\frac{Q_{open}}{Q_{short}} = \frac{2\pi f_r W_{open} / P_{open}}{2\pi f_r (0.5W_{open}) / (0.4P_{open})} = 0.8 \quad (\text{Eqn. 4.5})$$

4.5 Stacking End-Shorted Patches

The bandwidth that can be achieved with a normal end-shortened radiator on a thin substrate is much less than that required for the mobile satellite communication application. As a result, the technique of stacking such elements to increase the overall bandwidth was investigated.

Satellite communication systems typically use circular polarization and a method for producing circular polarization by combining end-shortened radiators is depicted in Figure 4.4. Here two elements are employed with each fed from a common feedpoint. The elements are positioned such that their radiating edges are orthogonal to each other. Circular polarization is then achieved by introducing a delay corresponding to 90 electrical degrees in the feed line to one of the elements.

4.5.1 Measured Results

A study was carried out to examine the effect of varying the size of the parasitic patch while maintaining the resonant dimension of the lower, driven element. These measurements indicate the extent to which the performance of the driven element is altered by the presence of the parasitic element.

The first configuration of the stacked elements is shown in Figure 4.5. Initially only the bottom element was constructed and measurements were made on its resonant frequency in a stand-alone configuration. This information was then used to determine the effect of the upper element on the behaviour of the lower element. The resonant frequency of the lower element at the coaxial probe was measured to be 1.647 GHz with a resonant resistance of 40.9Ω and a bandwidth (for $VSWR < 2$) of 1.6 percent. The shorting pins had a radius of 0.645 mm and a uniform spacing of 7.7 mm. Using Eqn. 4.1 the change in the resonant length ΔL is 0.44 mm. The lower patch was fed at its edge by a

Figure 4.4: A technique for obtaining circular polarization by combining two end-shortened microstrip antennas. After [12, p.196].

TOP VIEW

SIDE VIEW

Dimensions are in mm.

Figure 4.5: A stacked microstrip antenna composed of end-shortened patches.

narrow microstrip line used as a quarter-wavelength transformer with a width of 0.47 mm and a length of 36.4 mm to give a characteristic impedance of 175 Ω .

The upper and lower elements in the stack used polyolefin as the substrate material. This material has a dielectric constant of 2.32 and a loss tangent of 0.0003 at 1.6 GHz. While the dielectric constant is similar to that of RT Duroid 5870, the loss tangent for polyolefin is superior and the lamination is such that the copper can be peeled off the substrate, bypassing the need for etching the material during early prototype development.

The top patch was bonded to the lower patch and the length of the upper patch was varied to examine the resulting return loss behaviour. The data is plotted in Figure 4.6 where it is seen that variations in the length of the parasitic patch have minimal effect on the upper resonant frequency which is controlled by the lower, driven element. Therefore changes in the resonant length of the parasitic element primarily affect only its own resonant frequency. It is also apparent that when the length of the parasitic patch approaches that of the driven element, there appears to be minimal coupling between the elements since the behaviour of the lower element is dominant. However, the interaction tends to increase as the length of the parasitic patch is increased beyond that of the lower element. These observations are reasonable when one considers that the electric field lines from the driven element largely remain between the patch and the larger ground plane below. As the relative size of the upper patch increases, the field lines will continue to fringe toward the ground plane but also toward the extended upper element.

A second configuration was constructed in which the bottom patch in the stack was fed directly at its edge by a coaxial probe. The dimensions of the antenna are listed in Table 4.1. The measured resonant frequency for the bottom patch alone was 1.436 GHz with a radiation resistance of 939 Ω . The shorting pins had a radius of 0.645 mm and a

Figure 4.6: Variation of the frequencies of best return loss for the lower (■) and upper (●) end-shortened patches in a stacked configuration.

uniform spacing of 6.9 mm resulting in an increase in the resonant length of 0.21 mm at the resonant frequency. The shorting pins were soldered at the ground plane, at the bottom patch and at the top patch. The solder applied between the driven and parasitic patches introduced a small gap between the layers at the shorted end of the patches where the electric field intensity is minimum. It should be noted that this gap may have had some impact on the measured results by changing the effective dielectric constant along the length of the upper patch and indicates a drawback to stacking end-shortened radiators.

Table 4.1: Parameters of end-shortened elements in a stacked configuration.

Parameter	Top Patch	Bottom Patch
Substrate Thickness, h	$h_t = 1.57$ mm	$h_b = 3.15$ mm
Dielectric Constant, ϵ_r	2.32	2.32
Width, W	$W_t = 50.0$ mm	$W_b = 50.0$ mm
Length, L	$L_t = \text{variable}$	$L_b = 30.4$ mm

It is seen in Figure 4.7 that stacking the patches forms a loop in the impedance locus. Such impedance loops are characteristic of broadband structures provided the diameter is sufficiently narrow to allow matching about the characteristic impedance of the system to which the antenna is to be connected (typically 50 Ω). The diameter varies with the relative resonant lengths of the driven and parasitic patches and the bandwidth around the loop for each case is listed in Table 4.2. However, for the end-shortened stack the loop tends to have a large diameter which makes matching difficult over a large bandwidth. For example, the impedance loop shown in Figure 4.7c was centred about the 50 Ω point on the Smith chart using sections of transmission line and the optimization feature of the Touchstone software program. The return loss achieved was poor ranging between 5.5 dB and 9.5 dB over the bandwidth of the original loop. Thus although the bandwidth

(a) $L_t/L_b = 1.07$ (b) $L_t/L_b = 1.00$ (c) $L_t/L_b = 0.93$ (d) $L_t/L_b = 0.87$

Figure 4.7: Variation of input impedance for various ratios of length of the top and bottom patches in an end-shorted stack.

of the loops listed in Table 4.2 may be large, the frequency range over which an acceptable impedance match can be achieved is limited.

Finally, it is seen in Figure 4.7d that when the length of the top patch becomes increasingly shorter than the driven patch, the impedance loop is no longer present and the structure returns to the impedance behaviour of the isolated bottom patch.

Table 4.2: Bandwidth of loop in impedance locus for various ratios of patch size.

L_t/L_b	Frequency Range of Loop (MHz)	Bandwidth of Loop (MHz)
0.87	no loop	0
0.93	1525-1669	144
1.00	1425-1605	180
1.07	1298-1555	257

4.5.2 Conclusions

The investigation of end-shortened microstrip antennas in a stacked configuration has demonstrated that dual-frequency operation can be achieved. Measurements have been completed for a configuration in which the bottom radiator is fed while the top patch acts as a parasitic element. The resonant frequency of the fed element undergoes minimal variation when the resonant dimension of the parasitic patch is altered.

However, the design of these antennas suffer from several important limitations. Once the shorting pins have been connected, the bottom radiator becomes inaccessible for further tuning. The physical difficulty in achieving an end-short for a microstrip patch also poses certain fabrication difficulties. A small portion of the substrate for the parasitic radiator must be removed to allow space for the solder used to obtain a connection between the pins and the bottom patch. This causes a discontinuity in the substrate of the

top radiator. The alternative of using a continuous end-short, requires a cut in the substrate material and is difficult to implement when the radiators and the feed network are to be printed on the same substrate layer.

Open microstrip antennas appear to offer more promise for the mobile satellite communication application since they have an inherently wider impedance bandwidth than end-shortened patches and also provide a more flexible geometry for empirical optimization. Investigations on this configuration are discussed in Chapter 5.

5. STACKED MICROSTRIP ANTENNAS

5.1 Introduction

This chapter describes the development and characterization of a stacked antenna configuration consisting of open microstrip patches. The open microstrip element was selected to avoid the fabrication difficulties encountered with the end-shortened configuration and to increase the bandwidth of operation. The antenna was optimized for application as an element to be used in a planar array for mobile satellite communication from ground vehicles.

Initially the impedance bandwidth of thin, stacked microstrip antennas were investigated to assess their potential. Design tools for stacked elements are not currently available and the application of numerical techniques to the problem is very complex for most designs. As a result, a parametric study was carried out to illuminate the nature of interaction between elements in a stacked configuration.

5.2 Selected Geometry

There are numerous possible configurations in which microstrip patches can be stacked. Either the top or the bottom patch may be directly fed allowing the other patch to act as a parasitic element. The advantage of feeding the lower element is that once its size has been nearly optimized the upper element can be adjusted for position and size relatively easily. This allows the antenna to be tuned without physically disturbing the driven structure.

Air separations between the patches have been used in several studies described in the literature as discussed in Section 3.5.1. The separation is varied to provide an

additional tuning parameter. However, in this work no separation has been employed since it adds additional height to the structure. In addition, the separation adds a degree of mechanical complexity to obtain a rigid structure that will not suffer from temperature variations.

A square microstrip element was selected for investigation since it can be easily made to radiate circular polarization by implementing a quadrature feed and the radiation pattern has a peak at broadside which readily allows it to be employed in a phased array.

5.3 Impedance Characteristics of Stacked Microstrip Antennas

In the initial design of an antenna one of the first considerations is the impedance bandwidth that can be achieved. As a result, the impedance behaviour of stacked microstrip patches was experimentally investigated. Measurements were taken to examine the effects of patch size and offset.

5.3.1 Effect of Relative Patch Dimensions

In the first set of measurements, the size of an upper parasitic patch was varied while the lower element was fixed in size and driven at its edge by a coaxial probe. Both layers had a substrate consisting of polyolefin with a dielectric constant (ϵ_r) of 2.32, a loss tangent ($\tan\delta$) of 0.0003, and a thickness (h) of 3.15 mm. Since the intent was to provide circular polarization, the patches were square to provide geometrical symmetry and their centres were aligned as shown in Figure 5.1. The lower element had a fixed length $L = 58.6$ mm and width $W = 58.6$ mm. For this bottom patch the predicted resonant frequency using the program MICROANT.FOR in Appendix 1 is 1.605 GHz which is 0.8 percent higher than the measured value of 1.593 GHz. The predicted radiation resistance of 201 Ω is less than the measured value of 299 Ω . The resulting mismatch to a 50 Ω system would require the implementation of an impedance transformer. However, it will be

shown that stacking elements offers the additional benefit of lowering the impedance seen at the edge of the fed element.

Figure 5.2 shows the measured input impedance behaviour for various sizes of the upper parasitic element when the lower driven element is fixed. It is seen that when the upper patch is smaller than the driven element the input impedance curve is similar to that for a single element. In fact the impedance locus for the structure approaches that of the driven element in isolation. In Figure 5.2a the resonant frequency is 1552 MHz with a measured resistance of 231Ω , similar to the values for the lower element described earlier. This implies that there is minimal coupling between the patches through fringing E-fields. This is because the majority of the electric field is concentrated between the driven element and the ground plane. However, as the size of the parasitic element is increased the degree of coupling also increases and a broadband impedance loop appears. The increased coupling also tends to increase the radius and frequency range covered by the loop as shown in Table 5.1.

Since coupling between the patches occurs primarily through the fringing of electric field lines between layers it can be assumed that this coupling is capacitive in nature. A simple circuit model can be used to describe the measured results. By way of example, consider the case where the top and bottom patches are identical and each element can be considered as a parallel RLC-type resonator. The RLC parameters were obtained from the MICROANT.FOR program for the bottom patch in the configuration used for the measurements described above. Each RLC network is then coupled with a capacitance as shown in Figure 5.3. The Smith charts in Figure 5.4 show that as the capacitance is increased the impedance loop follows the same trend as when the size of the upper patch in a stacked configuration is increased thereby increasing the coupling between the layers.

Figure 5.1: A stacked microstrip antenna with patches aligned along common axes.

(a) $L_t/L_b = 0.90$ (b) $L_t/L_b = 0.93$ (c) $L_t/L_b = 0.96$ (d) $L_t/L_b = 0.99$

Figure 5.2: Input impedance over the frequency range 1-2 GHz for various ratios of top and bottom patch dimensions in a stacked configuration of square radiators. Continued...

(e) $L_t/L_b = 1.03$ (f) $L_t/L_b = 1.06$ (g) $L_t/L_b = 1.09$ (h) $L_t/L_b = 1.13$

Figure 5.2: Concluded.

Figure 5.3: A simple lumped-element circuit model used to study the coupling between stacked patches.

(a) $C_c = 0.5$ pF

(b) $C_c = 1.0$ pF

(c) $C_c = 2.0$ pF

Figure 5.4: Input impedance of stacked patches over the frequency range 1-2 GHz for various values of coupling capacitance C_c .

Table 5.1: Bandwidth of loop in impedance locus for various ratios of patch size.

L_t/L_b	Frequency Range of Loop (MHz)	Bandwidth of Loop (MHz)
0.90	1830-1830	0
0.93	1717-1804	87
0.96	1637-1768	131
0.99	1555-1725	170
1.03	1465-1685	220
1.06	1395-1655	260
1.09	1345-1625	280
1.13	1265-1585	320

5.3.2 Effect of Patch Offset

Stacking elements provides the designer with several additional degrees of freedom over conventional microstrip antennas. In addition to selecting the patch dimensions and dielectric constants for each element, the relative position of the radiators can be varied since the parasitic element is simply bonded to the surface of the driven element. Results presented by Barlatey [28] for various degrees of offset in a configuration of stacked rectangular patches showed a variation in the size of the impedance locus. A set of impedance measurements were made on the configuration shown in Figure 5.5. Since it is intended to develop a stacked element that radiates circular polarization, the offset was varied along the diagonal of the lower element to preserve a plane of symmetry.

Figure 5.5: A microstrip patch arrangement used to study the effects of offset between layers. Square patches are used with the same dimensions for each patch.

Impedance variations for several degrees of offset are graphed in Figure 5.6. Table 5.2 lists the frequency bandwidth covered by the perimeter of the loop in the impedance locus. The effect of offsetting the parasitic element has been to increase both the diameter of the impedance loop, as well as the frequency range covered by it. It was found previously, for the case in which the relative patch sizes were varied, that this impedance behaviour is consistent with increased coupling between the patches. As the diagonal offset is made larger it is expected that the degree of coupling would tend to increase on the radiating edge for which the parasitic patch extends beyond the driven element. Conversely, the degree of coupling would tend to be minimized at the opposite radiating edge. In addition, there is little rotation of the impedance loop around the Smith chart. This observation leads to the potential of utilizing offset as a technique for impedance matching stacked patch configurations. When smaller bandwidths of operation are required it is desirable to minimize patch offset to decrease the radius of the impedance loop. The loop can then be centred about the $50\ \Omega$ point on the Smith chart using tuning stubs in conjunction with an impedance transformer to obtain good return loss. When the antenna must provide dual-frequency operation, as in the case of mobile satellite communication, the offset technique may be employed to adjust the diameter of the impedance loop in order to obtain improved return loss in the required bands.

Figure 5.6: Input impedance variations over the frequency range 1-2 GHz for patch offsets of (A) 0 mm, (B) 2 mm, (C) 4 mm, and (D) 6 mm.

Table 5.2: Bandwidth of loop in impedance locus for offset patches.

Offset (mm)	Frequency Range of Loop (MHz)	Bandwidth of Loop (MHz)
0.0	1595-1680	85
2.0	1590-1700	110
4.0	1580-1710	130
6.0	1565-1730	165

5.3.3 Optimized Element

A stacked patch was empirically optimized to obtain an impedance bandwidth near that required for mobile communication. The element uses square patches on a polyolefin substrate ($\epsilon_r = 2.32$) as shown in Figure 5.7. The substrate thickness of the bottom patch is 1.57 mm while that of the upper patch is 3.15 mm. It is desirable to minimize the thickness of the lower patch because this allows the feed network to be etched on a thinner substrate and minimizes the line width required for a given characteristic impedance. The bottom patch is fed by a microstrip transmission line with a characteristic impedance of 50Ω . A tuning stub is located at the feedpoint in order to position the impedance locus about the centre of the Smith chart. Offset between the patch layers was also used for this purpose. The impedance curve is plotted on the Smith chart in Figure 5.8 and the corresponding return loss is given in Figure 5.9. It is clear that one of the advantages of using a stacked arrangement is that the introduction of the parasitic element decreases the resonant resistance to a value that is more easily matched to a typical 50Ω system. Ordinary microstrip antennas typically have a resonant resistance near 300Ω and require the use of impedance transformers in the feed network.

Figure 5.7: A stacked microstrip antenna.

Figure 5.8: Input impedance for an empirically optimized stacked antenna.

Figure 5.9: Return loss for an empirically optimized stacked antenna.

5.4 Circular Polarization

Circular polarization can be achieved in several ways by modifying the feeding mechanism employed to excite a microstrip antenna. One of the simplest and least sensitive methods is to feed the element on adjacent sides with a 90 degree electrical phase difference between the two feed lines as shown in Figure 5.10.

A Wilkinson splitter/combiner shown in Figure 5.11 can be employed to balance the relative power levels appearing at each of the two fed patch edges. A balanced power division is required in order to achieve a high degree of circular polarity and to minimize cross polarization loss. Examination of the equations that describe axial ratio demonstrate the dependence of polarization loss on the balance that can be achieved with the power division/combination network. The instantaneous field for a plane wave propagating in the negative z direction is given by [11, pp.48-51],

$$\vec{E}(z,t) = E_x \cos(\omega t + kz + \phi_x) \vec{a}_x + E_y \sin(\omega t + kz + \phi_y) \vec{a}_y. \quad (\text{Eqn. 5.1})$$

The axial ratio is the ratio of the major axis to the minor axis of the ellipse traced at a given position with time. For simplicity, it can be assumed that the major and minor axes of the ellipse are aligned with the x and y axes of the coordinate frame, giving

$$AR = \frac{\max(E_x, E_y)}{\min(E_x, E_y)}. \quad (\text{Eqn. 5.2})$$

Figure 5.12 shows the relationship between the axial ratio and the balance achieved with the Wilkinson power splitter. Figure 5.13 plots the variation of axial ratio with phase deviation from the required value of 90 degrees for circular polarization.

Minor modifications were made to the element described in Section 5.3.3 to develop a circularly polarized element for use in the mobile satellite communication

Figure 5.10: Feed configuration used to radiate circular polarization from a square microstrip antenna.

Figure 5.11: A two-way Wilkinson splitter/combiner implemented in microstrip transmission line.

Figure 5.12 : A graph of the variation of axial ratio with the percentage of total output power appearing at one output port of a two-way Wilkinson splitter ($P_2/(P_2+P_3) \times 100$).

Figure 5.13: A graph of the variation of axial ratio with the phase error from 90 degrees.

bands. The finalized dimensions of the element are provided in Figure 5.14. The photo mask used for etching the lower layer of the element, including the Wilkinson power splitter/combiner is shown in Figure 5.15. The return loss measured at the input to the Wilkinson splitter is graphed in Figure 5.16. A set of measured radiation patterns, using a circularly-polarized source, are given in Figures 5.17 and 5.18 for orthogonal cuts through the beam peak. It is seen that stacking the patches has not led to any degradation in the radiation patterns since they are symmetrical about a well-defined beam peak.

The axial ratio for a single element was measured at the edges of the transmit and receive bands on an antenna range using a spinning dipole source as shown in Figure 5.19. A sample radiation pattern is shown in Figure 5.20. The results of the axial ratio measurements are summarized in Table 5.3 as well as the corresponding cross-polarization loss. It is clear that stacked microstrip antennas are well suited to applications requiring circularly polarized radiation elements.

Table 5.3: Axial ratio and cross-polarization loss for a single stacked element.

Frequency (MHz)	Axial Ratio (dB)	Cross-Polarization Loss (dB)
1530	2.7	0.10
1559	3.5	0.18
1627	2.3	0.08
1661	2.1	0.06

Figure 5.14: Dimensions of the circularly polarized stacked microstrip antenna (a) top view (b) side view.

Figure 5.15: Photo mask for the bottom layer of the circularly polarized antenna.

Figure 5.16: Return loss for a single element measured at input to Wilkinson splitter.

Figure 5.17: Measured elevation radiation pattern at $\phi=45$ degrees for a stacked microstrip antenna at 1627 MHz.

Figure 5.18: Measured elevation radiation pattern at $\phi=135$ degrees for a stacked microstrip antenna at 1627 MHz.

Figure 5.19: Simplified diagram of the apparatus used to measure radiation patterns and axial ratio.

Figure 5.20: Measured axial ratio at $\phi=45$ degrees for a stacked element at 1627 MHz.

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5.5 Conclusions

This chapter has presented the characteristics for a particular geometry of stacked microstrip antenna. Specifically, a configuration was examined in which the bottom patch is driven at its edge while the top patch acts as a parasitic element. The only separation between the stacked radiators is the substrate required for the upper patch. This arrangement serves to take advantage of the bandwidth increase offered by electromagnetic coupling between patches while minimizing the increase in the volume occupied by the antenna. The antenna that has been optimized for satellite communication has shown no degradation in either its radiation pattern or degree of circular polarity over that of conventional microstrip antennas.

6. ARRAY SYNTHESIS AND DESIGN

6.1 Introduction

This chapter describes the development of a phased array which is mechanically-steered in azimuth for mobile satellite communication. The element is a stacked microstrip antenna as described in Chapter 5 and the array is intended for use over the bands 1530-1559 MHz and 1626.5-1660.5 MHz. It radiates right-handed circular polarization to conform to existing system protocols and can track the communications satellite as the vehicle moves by mechanically rotating the antenna to provide omnidirectional coverage in azimuth while elevation coverage is provided by a fixed beam located at the midpoint of the required coverage region.

One of the major achievements of the design developed in this thesis is its low profile allowing conformal mounting on the roof of motor vehicles. Ordinary microstrip patch technology cannot satisfy both the bandwidth and size requirements simultaneously, thereby driving the need for a stacked element. Application of the stacking technique has resulted in a low profile antenna with increased bandwidth. The resonant dimensions of the antenna are minimized to give a broader beamwidth allowing the beam to be scanned from broadside to provide coverage at lower elevation angles. Antennas that are currently available for the application occupy a large volume and suffer from a high production cost.

The prototype array developed here is designed to provide a fixed elevation beam with the peak set at approximately 40 degrees above the horizon which is the midpoint of the typical coverage region. Complete azimuth coverage is then achieved by rotating the array with a thin pancake motor. Such motors are available at a relatively low cost and

are sufficiently thin to maintain the shallow depth occupied by the array structure. Although it is possible to develop an array with a beam that is electronically steered in azimuth and elevation, the number of elements and adjustable phase shifters required would occupy a surface area larger than is practical for motor vehicles. The cost involved with a large number of phase shifters is also significant due to the relatively high cost of PIN diodes and their implementation.

6.2 Array Synthesis

6.2.1 Geometrical Constraints

Since the array is intended to be rotated to provide complete azimuth coverage, efficient utilization of space can be achieved by implementing an array in the form of a circular disk. The maximum diameter is restricted by the limited surface area available for mounting on mobile platforms. A realistic upper limit is a diameter of 50 cm.

In order to develop a relatively high gain array the beamwidth must be sufficiently narrow to provide the required directivity. Either the elevation beamwidth should be large enough to illuminate the entire coverage region or multiple switched beams must be provided. Since the antenna can be rotated to provide complete azimuthal coverage the beamwidth in this plane should to be kept narrow to maximize the directivity and to minimize possible interference from unwanted sources. The beam from the stacked microstrip element has a maximum in the direction perpendicular to the surface of the patch. This means that for a flat-roof mounting, the antenna will have a beam peak at 90 degrees above the horizon. This requires the beam to be scanned approximately 50 degrees to provide an array with maximum gain at 40 degrees (mid value of the desired elevation coverage region) above the horizon. These factors favour an element configuration in which there is a maximum number of elements in the elevation plane to

provide the necessary flexibility for scanning the beam peak, while keeping the sidelobes sufficiently low (approximately 13 dB below the peak array gain). In addition, the circular geometry allows several elements to be combined in the orthogonal plane to minimize the azimuthal beamwidth, thereby maximizing the array's directivity.

6.2.2 Pattern Synthesis

An array composed of ten stacked microstrip elements is appropriate for the disk structure of the array. The feed lines can be etched on the same layer as the lower microstrip elements to minimize the thickness of the array by eliminating a separate layer for the feed board. The cost to produce the array is also minimized by simplifying the manufacturing process. The basic layout shown in Figure 6.1 demonstrates that there are five pairs of elements in the elevation plane in which the beam is to be scanned and four sets in the orthogonal plane, providing the necessary degree of control to scan the beam in elevation while minimizing sidelobe levels. The elements have been arranged with collinear diagonals to minimize mutual coupling between elements.

Standard array techniques were employed to determine the required amplitude (I_n), phase (β_n), and position for each element. The array factor was combined, through pattern multiplication, with a measured element pattern to obtain the radiation pattern for the array. The array factor for a planar array lying in the x - y plane with N elements is given by the expression [11, p.261],

$$AF = \sum_{n=1}^N I_n \exp(jk(x_n \sin \theta \cos \phi + y_n \sin \theta \sin \phi)) \exp(j\beta_n) \quad (\text{Eqn. 6.1})$$

The layout is symmetrical about the centre elements (5 and 6) which are given the largest amplitude weighting. Thus, $I_1=I_2=I_9=I_{10}$, $I_3=I_4=I_7=I_8$, and $I_5=I_6$ reducing the number of amplitude factors to three. Similar symmetrical considerations can be used to select the

Figure 6.1: Diagram showing the element layout within the circular disk area of the array.

required position of each element. In addition, the phase shift between elements (β_n) is chosen to provide a beam peak at approximately 42 degrees above the horizon. Table 6.1 lists the amplitude weighting and phase factor of each element corresponding to the numbering system given in Figure 6.1.

Table 6.1: Optimized amplitude, phase and positions for each element in the planar array.

Element n	I_n (A)	β_n (degrees)
1	0.49	0
2	0.49	0
3	0.81	-129
4	0.81	-129
5	1.00	-258
6	1.00	-258
7	0.81	-387
8	0.81	-387
9	0.49	-516
10	0.49	-516

Using these values, radiation patterns were simulated in the azimuth and elevation planes intersecting the main beam at the extreme lower and upper frequencies and at the centre frequency. The normalized results have been graphed in Figure 6.2 for the azimuth cut through the main beam and in Figure 6.3 for the elevation plane. It can be seen that the beam peak location is located near the desired location of 42 degrees above the horizon in the elevation plane throughout the band. In both sets of pattern simulations the sidelobe levels are kept at least 13 dB below the main beam.

Figure 6.2: Simulated radiation patterns of the azimuth cut through the main beam.

Figure 6.3: Simulated radiation patterns in the principal elevation plane. Zenith corresponds to 0 degrees.

6.3 Mutual Coupling

The degree of mutual coupling between elements in an array is an important consideration in determining the minimum spacing between elements within the array. Nearby radiating elements can interfere with the field radiated by an antenna and alter its current distribution thereby causing a change in the input impedance to the antenna. However, the effects of mutual coupling are not always detrimental. In fact, this effect has been exploited for the stacked antenna by introducing a parasitic antenna with no current excitation of its own in the vicinity of a fed microstrip patch. The input impedance to the lower driven patch was found to decrease significantly from its isolated value. Similar use of parasitic elements has been employed in the development of the well known Yagi-Uda antenna.

An obvious technique for minimizing mutual coupling is to increase the inter-element spacing within the array. The extent to which this method can be utilized is of course limited by the maximum allowable size of the array. The degree of mutual coupling is also influenced by the relative orientation of the elements. Maximum coupling occurs for parallel radiating edges with a significant reduction for configurations in which radiating edges are orthogonal to each other. For this reason, the land-mobile array was designed so that elements with the closest spacing have perpendicular radiating edges.

Lee *et al.* [19] studied the degree of mutual coupling between stacked antenna as a function of inter-element spacing. In addition, the coupling between two single-layer patch antennas was measured for comparison. Their investigation was carried out at X-band near 10 GHz and used rectangular fed and parasitic elements of the same dimensions ($W=15$ mm and $L=10$ mm) for the stacked elements. The measurements were performed for coupling between both radiating (E-plane) and non-radiating (H-plane) edges for patches excited in the TM_{01} mode. The results shown in Figure 6.4 indicate that stacking

Figure 6.4: Mutual coupling between single layer (O) and stacked (Δ) microstrip antennas for (a) radiating edge coupling and (b) non-radiating edge coupling. From [19].

patches does not significantly change the degree of mutual coupling between elements.

A model of a portion of the land-mobile array was built in order to test the degree of mutual coupling. Three of the elements were selected to provide all of the possible inter-element spacings. Figure 6.5 shows the element configuration used for the measurements and provides element labels that are referred to in Figures 6.6 - 6.8 which plot the insertion loss between various elements. The unused element in each measurement set was terminated in a $50\ \Omega$ load to eliminate any effects from its presence. The return loss of each element has also been subtracted in the insertion loss graphs. The maximum coupling levels are approximately -18 dB, indicating minimal effects on the array performance.

6.4 Feed Design

The next step in the array development was the design of the feed network. Touchstone was used as the development tool to model the frequency-dependent behaviour and Academy was used to produce the mask layout for photo etching. Although these software tools provide accurate electrical modeling, physical layout of the microstrip lines required significant effort to ensure correct placement of the radiating elements. The photo mask for the bottom layer which includes the feed network and the lower radiators is shown in Figure 6.9. Appendix 2 gives the Touchstone code used to produce the circuit layout for the bottom layer of the array.

The feed network was designed at the centre frequency of 1.595 GHz. The deviation from desired values at the extreme band edges was simulated using Touchstone to ensure that the degradation in predicted performance would be minimized. Combinations of quarter wavelength transformer sections were used to obtain the required

Figure 6.5: Element configuration used to measure mutual coupling.

Figure 6.6: Coupling between adjacent antenna elements 1 and 2.

Figure 6.7: Coupling between adjacent antenna elements 2 and 3.

Figure 6.8: Coupling between antenna elements 1 and 3.

Figure 6.9: Photo mask used for etching feed network and bottom patches.

input impedance to each branch. The characteristic impedances of each line were kept between 50 Ω and 100 Ω to keep the line widths sufficiently narrow to minimize the space required without using lines too narrow for etching tolerances.

6.5 Array Assembly

Following the design procedure outlined in section 6.4, layout files were generated to produce the photo masks used for etching the mobile array on two layers. The first layer consisted of the feed network and the bottom radiators as shown in Figure 6.9. This sheet was cut into the shape of a circular disk. The disk was then bonded into an aluminum support tray using conductive epoxy for support. This was done to alleviate the warping undergone by the polyolefin dielectric when copper has been removed from one side. The resistors used in the Wilkinson splitters were also connected to the board using conductive epoxy.

The upper layer contained a simple grid of patches for the parasitic elements of the stacked configuration as shown in Figure 6.10. These patches were sheared into individual elements and bonded to the surface of the driven patches on the bottom layer. The array has an outer diameter of 47.3 cm and a very thin profile with a total thickness, excluding the connector, of 0.79 cm as shown in Figure 6.11. Figure 6.12 shows a photo of the completed assembly.

Figure 6.10: Photo mask used for etching a grid of parasitic patches.

Figure 6.11: Side view of assembled mobile array in support tray.

Figure 6.12: Antenna array.

6.6 Measured Performance

6.6.1 Radiation Patterns

A probing technique was used to measure the relative excitation amplitude and phase at each element. A dipole antenna was placed near each element with a fixed air separation and orientation, using honeycomb material between the patch and the dipole. A sample measured radiation pattern through the main beam at 1627 MHz is shown for azimuth in Figure 6.13 and for elevation in Figure 6.14. Calculated patterns using the results of the dipole probe measurements have been superimposed for comparison on the corresponding plots. The calculated patterns were obtained through pattern multiplication of the array factor and a measured element radiation pattern. It can be seen that there is a close correspondence between the measured and calculated patterns indicating that the dipole probing technique is a useful tool for analyzing array performance in L-band.

Sidelobe levels can be reduced by maximizing the relative excitation of the central elements of the array with respect to those of the peripheral elements. However, this comes at the expense of reducing the gain of the array. This becomes clear by considering the extreme case in which the excitation to all elements except the central element is set to zero. In this case the sidelobe levels are reduced, but only a small portion of the available aperture is utilized. This low aperture efficiency yields a reduced gain, and in the extreme case provides only the gain of a single element. On the other hand, if the sidelobe levels become too large then the gain will also be reduced because of the decreased directivity. Thus, there is a tradeoff between sidelobe level reduction through tapering the excitation across the aperture and the gain that can be achieved.

The sidelobe levels of the radiation patterns were lowered by adjusting the amplitude weighting on the array. The relative power to the central elements was increased by decreasing the input impedance seen looking into these elements. This was

Figure 6.13: Comparison of measured and calculated radiation patterns at 1626.5 MHz in the azimuth plane of the main beam. The peak gain is 11.4 dBic.

Figure 6.14: Comparison of measured and calculated radiation patterns at 1626.5 MHz in the elevation plane where 0 degrees corresponds to zenith. The peak gain is 11.4 dBic.

accomplished by increasing the line width of a quarter wavelength transformer by soldering a conductive strip over the existing track. In addition, the power level to the outer elements was decreased slightly using the same technique. The measured radiation patterns after these modifications are shown in Figures 6.15 - 6.18. It is seen that the sidelobe levels have been improved using the information supplied by the dipole probing measurements. The requirement has been met to keep sidelobe levels at least 13 dB below the level of the main beam at 40 degrees from boresight in azimuth and 80 degrees from boresight in elevation, for the principal cuts through the main beam. The measured beamwidths for the two principal planes are listed in Table 6.2.

Table 6.2: 3 dB beamwidths for the principal azimuth and elevation patterns.

Frequency (GHz)	Azimuth 3dB Beamwidth (degrees)	Elevation 3dB Beamwidth (degrees)
1.5300	24	36
1.5590	24	35
1.6265	23	33
1.6605	21	33

6.6.2 Antenna Noise Temperature

Noise is produced by the movement of electrons in any component, such as a resistor, that has a temperature above absolute zero or 0 K (-273°C) [29, ch.12]. The warm electrons move about randomly and collide with ions that are present in the material. Although there is no net average current produced in the medium, there are a series of random pulses which produce thermal agitation noise or Johnson noise. The resulting noise power is directly proportional to the temperature of the medium with

$$P_{noise} = kTB \quad (\text{Eqn. 6.2})$$

where k is Boltzman's constant (1.38×10^{-23} joules per K), T is the temperature of the device, and B is the bandwidth.

Figure 6.15: Measured azimuth pattern through main beam at 1530 MHz.

Figure 6.16: Measured azimuth pattern through main beam at 1661 MHz.

Figure 6.17: Measured elevation pattern at 1530 MHz. 0 degrees corresponds to zenith.

Figure 6.18: Measured elevation pattern at 1661 MHz. 0 degrees corresponds to zenith.

Antenna noise temperature provides a measure of the antenna's contribution to the signal-to-noise ratio (SNR) degradation of the receiver/transmitter system. It is well known that the noise performance of the receiver front-end electronics predominantly determine the SNR ratio of the overall system. The antenna temperature is a combination of noise contributions from external sources, due to the sky and ground temperature in the illumination region of the antenna, and internal sources from ohmic losses.

The illumination temperature (T_a^i) is a function of the distribution of brightness temperature (T_b) viewed by the gain of the antenna radiation pattern as given in the following equation [11, p.68],

$$T_a^i = \frac{\iint G(\theta, \phi) T_b(\theta, \phi) \sin \theta d\theta d\phi}{\iint G(\theta, \phi) \sin \theta d\theta d\phi} \quad (\text{Eqn. 6.3})$$

This equation applies the simplifying assumption that the polarization of the antenna and the noise are the same. In reality, the sky noise is essentially randomly polarized with the result that Eqn. 6.3 yields higher temperatures. A more rigorous treatment of the effects of polarization on antenna temperature can be found in [30, ch.31]. The temperature distribution of the sky varies from about 5 K toward zenith to 100-150 K toward the horizon [11, p.68]. This is significantly below that of the ground which is at approximately 300 K. If permissible for the application, it is desirable to limit the amount of noise introduced into the antenna by reducing sidelobe levels at low elevation angles.

The second contributing factor to antenna temperature is due to ohmic losses within the antenna itself and its feed network. Thus the antenna temperature may provide an indication of the antenna efficiency. These losses will cause a decrease in the illumination temperature through attenuation but will cause a more significant additional noise term.

The antenna temperature for the mobile array was measured in the receive band using the equipment setup shown in Figure 6.19. The measurements were made in an obstruction-free area of a large roof. In addition, a large copper-clad sheet, simulating a vehicular roof, was placed under the antenna to minimize contributions from the high temperature spillover energy of the ground.

The first stage in the determination of antenna temperature is to calculate the noise temperature introduced by the receiving circuitry. This is accomplished by measuring the noise temperature of the low noise amplifier (LNA) within the setup shown in Figure 6.19. This approximates the noise temperature for the receiving system since the noise source, load and antenna were directly connected to the input of the LNA without any cable loss. A calibrated noise source was connected to the input of the LNA and the resulting output power was measured on a spectrum analyzer. A second power measurement was made with the noise source turned off, causing it to act as a $50\ \Omega$ termination. The ratio of these two power levels is known as a Y-factor. Combining this ratio with the calibration chart of excess noise ratio (ENR) for the noise source yields the noise figure (F) of the LNA using [29, ch.12],

$$F = ENR - 10\log(Y - 1) \quad (\text{Eqn. 6.4})$$

where,

$$Y = \frac{N_{on}}{N_{off}}. \quad (\text{Eqn. 6.5})$$

The noise figure of the LNA can then be converted to a noise temperature through the relation,

$$T_{LNA} = T_o(F_{LNA} - 1) \quad (\text{Eqn. 6.6})$$

in which T_o is 290 K. Table 6.3 summarizes the results for frequencies at the extrema of the receive band.

Figure 6.19: Test equipment used to measure antenna noise temperature.

Table 6.3: Calculation of the noise temperature of the LNA.

Freq. (GHz)	N _{on} (dBm)	N _{off} (dBm)	Y (dB)	ENR (dB)	F _{LNA} (dB)	T _{LNA} (K)
1.530	-38.0	-51.8	13.8	15.09	1.53	121.8
1.559	-38.8	-52.5	13.7	15.08	1.52	121.8

The antenna temperature can be found using the ratio of the output power measured with the antenna at the input, P_{ant} , and the power level measured when the input is terminated in the system impedance of $50\ \Omega$, P_{load} .

$$T_a = \frac{P_{ant}}{P_{load}} (T_{sys} + T_{load}) - T_{sys} \quad (\text{Eqn. 6.7})$$

$$T_{sys} \cong T_{LNA}$$

Table 6.4 summarizes the calculations used to determine the antenna noise temperature in the receive band.

Table 6.4: Calculation of the antenna noise temperature.

Freq. (GHz)	P _{ant} (dBm)	P _{load} (dBm)	T _{load} (K)	T _{sys} (K)	T _a (K)
1.530	-54.5	-51.8	290	121.8	96.8
1.559	-56.0	-52.5	290	121.8	62.1

The antenna was mounted parallel to the ground with the beam peak directed to approximately 40 degrees above the horizon to give an illumination temperature of roughly 40 K. Since the radiation patterns at the two frequencies are quite similar, it can be concluded that there is more ohmic loss at the low end of the frequency band. This is most likely due to a favourable resonance in the feed network toward the centre of the frequency band (1.595 GHz) where the design was optimized.

6.6.3 Antenna Gain

The required antenna gain is determined from the ratio of gain to noise temperature (G/T) which should be greater than -12 dB/K. Assuming that the system utilizes a low noise amplifier with a noise figure of 0.5 dB ($F_{LNA}=1.12$) and the combined diplexer, cable and connector losses are 0.9 dB ($F_{LOSS}=1.23$) the receiver temperature is found to be 110 K using Eqn. 6.8.

$$\begin{aligned} T_{REC} &= (F_{REC} - 1)T_o \\ F_{REC} &= F_{LOSS} + (F_{LNA} - 1) / G_{LOSS} \end{aligned} \quad (\text{Eqn. 6.8})$$

Adding the receiver temperature to the measured antenna temperature from Table 6.4 requires that the antenna gain be at least 11.1 dBic at 1.530 GHz and 10.4 dBic at 1.559 GHz in order to have a minimum G/T of -12 dB/K.

The gain of the array was measured using a circularly-polarized helical source and a calibrated gain standard antenna. Table 6.5 lists the measured gain and axial ratio at the edges of the transmit and receive bands. The directivity values were approximated using the expression [11, p.33]

$$D \cong \frac{\pi^2}{\Omega_A} \approx \frac{\pi^2}{\Theta_{HP1} \Theta_{HP2}} \quad (\text{Eqn. 6.9})$$

in which Ω_A is the beam solid angle which would enclose all of the radiated power if the radiation intensity was set constant to its maximum value. For radiation patterns with only one major lobe and low intensity sidelobes Ω_A can be approximated by the product of half-power beamwidths in orthogonal planes which cut through the main lobe. Improved accuracy was achieved by using effective half-power beamwidths (Θ_{eff}) which were calculated by integrating the area under each orthogonal cut using

$$\Theta_{\text{eff}} = \frac{\int_0^{2\pi} G(\theta) d\theta}{G_{\text{max}}}. \quad (\text{Eqn. 6.10})$$

The difference between the calculated directivity and measured gain indicate significantly higher losses than the anticipated losses listed in Table 6.6 which include such factors as the return loss plotted in Figure 6.20. The loss values are much higher than anticipated based on the noise temperature measurements which did not indicate substantial feed losses.

Table 6.5: Axial ratio, directivity, gain, and losses.

Frequency (GHz)	Axial Ratio (dB)	Θ_{eff} in Azimuth (deg.)	Θ_{eff} in Elevation (deg.)	Directivity (dB)	Measured Gain (dBic)	Loss (dB)
1.530	4.7	25.7	45.1	14.5	10.2	4.3
1.559	4.1	25.4	46.7	14.4	11.2	3.2
1.627	3.0	25.2	45.1	14.5	11.4	3.1
1.661	3.4	25.5	46.7	14.3	11.3	3.0

Table 6.6: Estimated losses.

Loss Factor	Estimated Loss (dB)
Mismatch (Return Loss = 13 dB)	0.22
Connector	0.05
Element (96% Efficiency)	0.18
Feed Network	0.30
Cross-Polarization (Average)	0.22
Total	0.97

Figure 6.20: Return loss of mobile array.

Pattern measurements at additional azimuth angles (not intersecting the main beam) revealed the presence of two lobes in addition to the main beam resulting in the reduced antenna gain. The directivity values shown in Table 6.5 did not account for these additional lobes since the calculation was based on the two principal cuts through the main beam. These unwanted lobes were located at an azimuth angle ± 120 degrees from the main beam and at an elevation angle approximately 57 degrees above the horizon with a peak level nearly equal to that of the main beam.

Analysis was carried out to determine if these unwanted lobes were grating lobes resulting from a larger than allowable inter-element spacing. The maximum allowable inter-element spacing (d) to avoid grating lobes is [30, pp. 20-6]

$$d = \frac{\lambda}{1 + \sin|\theta_o|} \quad (\text{Eqn. 6.11})$$

where θ_o is the scan angle measured from broadside. The array was designed with a phase taper to give a squint angle of $\theta_o = 48$ degrees for the pattern cut at $\phi = 0$ degrees. For the pattern cut at $\phi = 90$ degrees, the elements are excited in phase and so the beam is not scanned in this plane. Pattern cuts at intermediate azimuth angles will have a beam squint between these two values (i.e., $0 \leq \theta_o \leq 48$ degrees) and can be found from Eqn. 6.12 which has been derived from simple geometry and where $\theta_{\max}$ is the scan angle in the plane of maximum scan ($\phi=0$)

$$\theta_o = \tan^{-1}(\tan \theta_{\max} \cdot \cos \phi) \quad (\text{Eqn. 6.12})$$

The spacing between elements was determined for various azimuth angles by projecting a perpendicular line from each element onto the plane of the cut as shown in Figure 6.21. The maximum spacings are plotted in Figure 6.22 along with the maximum allowable spacing to avoid grating lobes using Eqn. 6.11. It is seen that the inter-element spacing is not large enough to produce grating lobes.

Since the unwanted sidelobes were not found to be grating lobes, the amplitudes and phases were examined for each element when projected onto the cut for $\phi = \pm 120$ degrees as shown in Figure 6.23. It was determined that the non-uniform amplitude, phase and element spacings in this plane combine to form a linear array which produces the radiation pattern shown in Figure 6.24. The pattern has a maximum directed 33 degrees off boresight giving a high sidelobe and reducing the gain of the array.

6.7 Optimized Array to Suppress Sidelobes

This section proposes a re-optimized design to suppress sidelobes in the non-principal planes. The design presented here will maintain the same overall size of the original array but incorporates two additional elements within the array.

The only way to suppress the sidelobes using the current number and arrangement of elements is by decreasing the phase taper to allow the beam peak to move further above the horizon. However, the radiation pattern rolls off to such an extent that the gain for low elevation angles would be impractical. As a result, a re-optimized design was carried out in which pattern cuts at all azimuth angles, in addition to those through the main beam were considered. The diameter of the array was kept within the bounds of the previous design and the scan angle was also maintained. Adjustable parameters for optimization include the number of elements, their location and the amplitude and phase taper across the array.

Figure 6.21: Diagram showing the projection of the element positions onto an arbitrary azimuth cut.

Figure 6.22: A graph showing that the inter-element spacing avoids grating lobes by not exceeding the maximum allowable spacing.

(a)

(b)

Figure 6.23: Graphs showing the (a) amplitude and (b) phase fluctuations for $\phi = \pm 120$ degrees.

Figure 6.24: Simulated radiation pattern for the cut at $\phi=120$ degrees showing sidelobe at approximately 33 degrees off boresight. 0 degrees corresponds to zenith.

Filling the aperture of the array with additional elements, within the available space, will provide additional parameters for optimization and pattern control. Additional space could be obtained by implementing the feed network in stripline on a substrate layer below the radiating elements. This option was not selected because it increases the thickness of the array and results in a more expensive structure to fabricate. Using a microstrip feed network will allow the addition of two more elements within the confines of the 47 cm diameter disk.

After numerous iterations using Eqn. 6.1, the layout shown in Figure 6.25 was selected. This design suppresses unwanted sidelobes for all azimuth angles while maintaining the beam scan and diameter of the previous design. The simulated principal pattern cuts through the main beam plotted in Figures 6.26-6.27 show that sidelobe levels are kept at least 13 dB below the level of the main beam.

Table 6.7: Optimized amplitude, phase and position for a 12-element planar array.

Element n	I_n (A)	β_n (degrees)
1	0.50	0
2	0.50	0
3	0.58	-129
4	0.58	-129
5	0.58	-129
6	1.00	-258
7	1.00	-258
8	0.58	-387
9	0.58	-387
10	0.58	-387
11	0.50	-516
12	0.50	-516

All dimensions in cm.

Figure 6.25: Optimized layout for 12-element array.

Figure 6.26: Simulated elevation pattern for 12-element phased array. 0 degrees corresponds to zenith.

Figure 6.27: Simulated azimuth pattern through the beam peak of the 12-element phased array.

6.8 Summary

This chapter has provided a complete characterization of a land-mobile array composed of stacked microstrip elements with a significantly lower profile than other antennas developed to date. Although the principal radiation patterns for the main beam, the antenna noise temperature, polarization and axial ratio were satisfactory, the gain was decreased due to the presence of unwanted sidelobes. However, a re-optimization of the design has been proposed which will eliminate these sidelobes and should meet the requirements for land-mobile communication from ground vehicles. In addition, although the array has a fixed beam in the elevation plane, the addition of phase shifters could be implemented in a future design to produce multiple beams and provide increased gain within the coverage envelope.

7. SUMMARY AND RECOMMENDATIONS

7.1 Summary

This thesis has investigated the technique of stacking microstrip antennas to increase their impedance bandwidth. In the particular configurations studied, the lower patch is fed directly either by a coaxial probe or using a microstrip transmission line. The upper patch acts as a parasitic element through electromagnetic coupling with the lower element. The impedance behaviour for two types of stacked microstrip radiating elements has been studied.

The first configuration involved a novel arrangement consisting of end-shortened microstrip patches. The large beamwidth of end-shortened antennas is an attractive feature for mobile satellite communication since it allows the beam to be scanned from broadside to provide coverage at lower elevation angles. However, the mechanical fabrication difficulties of stacked configurations pose a serious drawback to their application.

The second stacked configuration consisted of open microstrip patches. The impedance behaviour and the degree of coupling between the layers was investigated for this arrangement. It has been found that the coupling can be increased by increasing the size of the upper patch relative to the bottom patch and can be modified by introducing an offset between the stacked patches.

A circularly polarized radiating element consisting of open stacked patches has been optimized for use as a land-mobile satellite communication antenna. The antenna exhibits an increased impedance bandwidth without adding significantly to the overall volume occupied by it. The resonant dimensions are much smaller than would be required

if the conventional technique of using a thick substrate with a low dielectric constant was used to satisfy the bandwidth requirements. As a result the bandwidth is increased and a larger beamwidth is maintained which is desirable for scanning the beam from broadside in a phased array. There is no degradation to the antenna's performance in terms of radiation patterns or the degree of circular polarity that can be achieved by employing a standard quadrature feed arrangement.

An array of stacked elements has been designed and tested. The elements are placed within the confines of a disk structure allowing the array to be mechanically-rotated to provide azimuthal coverage. The phase and amplitude distribution across to each element has been optimized to produce an elevation beam peak directed to approximately 42 degrees above the horizon. This location is the midpoint within the required elevation coverage envelope. The array has a diameter of only 47 cm with a substrate thickness of 0.47 cm. The prototype has been mounted in an aluminum support tray giving an overall diameter of 47.3 cm and a height of 0.79 cm (excluding the RF connector). These dimensions represent a vast improvement over previous antennas developed for the application.

7.2 Recommendations for Future Work

Stacked microstrip configurations have a large range of possible applications and further effort could be directed toward investigating their behaviour. In particular, the following areas could be pursued:

- Investigate feed arrangements to improve the axial ratio that can be achieved and determine the effect of dielectric separation and offset on polarization purity.
- Engineering design models could be developed, perhaps based on empirical measurements to produce design curves or lumped-element circuit models.

For the land-mobile array, the optimized design incorporating 12 elements to eliminate sidelobes in the non-principal planes can be implemented. In addition, future designs could be developed which incorporate phase shifters to provide multiple beam positions and enhanced coverage.

APPENDIX 1

```

*-----
*
* Program:      MICROANT.FOR
* Purpose:      Uses the cavity model to determine the resonant
*               frequency, radiation resistance, and bandwidth
*               of a rectangular microstrip antenna.
* References:   K. Carver and J.W. Mink, "Microstrip antenna
*               technology," IEEE Trans. Antennas Propagat.,
*               vol. AP-29, pp. 2-24, Jan. 1981.
*               D.M. Pozar, Antenna Design Using Personal Computers,
*               Dedham, MA: Artech House, 1985, pp. 121-126.
*-----

* Variable definitions
*      Eo        - permittivity of free space
*      uo        - permeability of free space
*      Zo        - characteristic impedance of free space
*      SIGMA     - conductivity of copper
*      Ee        - effective permittivity for non-air dielectrics
*      Er        - dielectric constant
*      tanD      - loss tangent for substrate
*      Gw        - wall conductance
*      Bw        - wall susceptance
*      Yw        - wall admittance
*      Qr,Qc,Qd- radiation, conductance, and dielectric loss Q-factors
*      Qt        - total quality factor
*      Rrad      - radiation resistance at resonance
*      Rin       - input resistance including copper and dielectric
*                 losses
*      dell      - effective length extension
*      cmplxw    - complex resonant frequency
*      fr        - resonant frequency
*      BandW     - impedance bandwidth
*      AntEff    - antenna efficiency
*      K10       - eigenvalue corresponding to TM10 mode

* Variable declarations
*      COMPLEX ALPHA, DELTA, K10, Yw, j, cmplxw
*      REAL    PI, c, Eo, uo, Zo, SIGMA
*      REAL    WLo
*      REAL    dell
*      REAL    Gw, Bw
*      REAL    fr
*      REAL    W, L, h, Er, Ee, tanD
*      REAL    Qr, Qc, Qd, Qt, BandW, AntEff
*      REAL    Rrad, Rin, Cdc, CAP, IND
*      INTEGER I

```

* Constants

```

PI = 3.1415926
c = 3E8
Zo = 377
Eo = 8.854E-12
uo = PI * 4E-7
SIGMA = 5.80E7
j = (0, 1)

```

* Enter the patch parameters

```

PRINT*, 'Resonant Length (m)'
READ*, L
PRINT*, 'Width (m)'
READ*, W
PRINT*, 'Dielectric Constant'
READ*, Er
PRINT*, 'Substrate Thickness (m)'
READ*, h
PRINT*, 'Loss Tangent'
READ*, tanD
IF (tanD .LT. 1E-8) THEN
    tanD = 1E-8
ENDIF

```

* Effective dielectric constant, Ee

```

Ee = (Er+1)/2 + (Er-1)/(2*SQRT(1+12*h/W))

```

* Approximate free space wavelength, WLo

```

WLo = 2*L*SQRT(Ee)

```

* Effective length extension, dell

```

dell = 0.412*h*(Ee+0.3)*(W/h+0.264)/(Ee-0.258)/(W/h+0.8)

```

* Wall conductance

```

Gw = 0.00836 * W / WLo

```

* Wall susceptance

```

Bw = 0.01668 * dell / h * W / WLo * Ee

```

* Wall admittance

```

Yw = CMPLX(Gw, Bw)

```

* Aspect ratio factor, F(W/L)

```

F = 0.7747 + 0.5977 * (W/L - 1) - 0.1638 * (W/L - 1)**2

```

```

ALPHA = j * 2 * PI * Zo * h * Yw * F / WLo / W

```

* Solve transcendental equation for complex eigenvalue, K10

```

* Set seed value for DELTA = 0 + j0

```

```

      DELTA = (0,0)

      DO 10 I = 1, 4
        DELTA = (2*ALPHA*L*(PI - DELTA)) /
*              ((ALPHA*L)**2 + 2*DELTA*PI - DELTA**2 - PI**2)
*              - DELTA**3/3
10      CONTINUE

      K10 = (PI - DELTA) / L

* Resonant frequency, fr
      cmplxw = c / SQRT(Er * (1 - j * tanD)) * K10
      fr = REAL(cmplxw) / 2 / PI

* Radiation quality factor, Qr
      Qr = REAL(K10) / 2 / IMAG(K10)

* Patch capacitance, CAP
      Cdc = Eo * Er * W * L / h
      CAP = Cdc / 2

* Patch inductance, IND for RLC model
      IND = 1 / CAP / (REAL(cmplxw))**2

* Radiation resistance at resonance, Rrad
      Rrad = Qr / REAL(cmplxw) / CAP

* Dielectric loss Q-factor, Qd
      Qd = 1 / tanD

* Conductor loss Q-factor, Qc
      Qc = h * SQRT(uo * PI * fr * SIGMA)

* Total quality factor, Qt
      Qt = 1 / (1/Qr + 1/Qd + 1/Qc)

* Input resistance, Rin
      Rin = Qt / REAL(cmplxw) / CAP

* Bandwidth for VSWR < 2
      BandW = 1 / Qt / SQRT(2)

* Antenna efficiency, AE
      AntEff = Qt / Qr * 100

* Report results
      PRINT*, '
      PRINT*, 'Resonant frequency      =', fr, ' Hz'
      PRINT*, 'Radiation resistance    =', Rrad, ' ohms'

```

```
PRINT*, 'Input resistance      =', Rin, ' ohms'
PRINT*, 'Q-factor            =', Qt
PRINT*, 'Bandwidth (for VSWR<2) =', BandW
PRINT*, 'Antenna efficiency    =', AntEff, ' %'
PRINT*, 'Element values for parallel RLC model'
PRINT*, 'Capacitance          =', CAP, ' F'
PRINT*, 'Inductance           =', IND, ' H'

STOP
END
```


APPENDIX 2

This appendix gives the Touchstone code used to generate a mask of the lower layer of the ground-mobile array including the bottom radiators and the feed network.

Figure A1.1: Node numbers referred to in Touchstone code used to generate mask for array feed network.

! Element and feed layout.

DIM

FREQ	MHZ
LNG	CM

VAR

! Table of line widths and quarter wavelengths for various
! characteristic impedances.

! format: W12345 = width of line for Zo=123.45 ohms

! L12345 = quarter wavelength for Zo=123.45 ohms

W05000	=0.4637
L05000	=3.3567
W05384	=0.4137
L05384	=3.3694
W05670	=0.3810
L05670	=3.3785
W06173	=0.3314
L06173	=3.3937
W07321	=0.2456
L07321	=3.4252
W07363	=0.2431
L07363	=3.4261
W07621	=0.2278
L07621	=3.4327
W08265	=0.1945
L08265	=3.4482
W08826	=0.1699
L08826	=3.4608
W09100	=0.1593
L09100	=3.4666
W10000	=0.1288
L10000	=3.4845

CKT

MSUB ER=2.32 H=0.15748 T=0.003556 RHO=0.71 RGH=0

!-----Radiating Element and Power Divider

! Hybrid power splitter/combiner

MTEE 1 2 3 W1=0.262 W2=0.262 W3=0.464

! Reference feed path with no delay

MLIN 1 200 W=0.262 L=0.990

MCURVE 201 200 W=0.262 ANG=180 RAD=0.367

MLIN 201 202 W=0.262 L=0.591

MBEND2 202 203 W=0.262

MLIN 203 204 W=0.262 L=0.453

MSTEP 204 205 W1=0.262 W2=0.464

! Tuning stub

```

MTEE      210 205 220 W1=0.464 W2=0.464 W3=0.464
MLIN      220 225 W=0.464 L=1.24
MLIN      210 211 W=0.464 L=2.5363
MBEND     211 212 W=0.464 ANG=-42 M=0
MLIN      212 213 W=0.464 L=0.288

!      Orthogonal feed path with 90 degree delay
MLIN      2 300 W=0.262 L=0.990
MCURVE    300 301 W=0.262 ANG=180 RAD=0.367
MLIN      301 302 W=0.262 L=0.591
MBEND2    303 302 W=0.262
MLIN      303 304 W=0.262 L=0.453
MSTEP     304 305 W1=0.262 W2=0.464
MBEND2    306 305 W=0.464
MLIN      306 307 W=0.464 L=3.330
MBEND     307 308 W=0.464 ANG=45 M=0.146

!      Tuning stub
MTEE      320 308 327 W1=0.429 W2=0.429 W3=0.001
MLIN      327 328 W=0.464 L=1.271
MBEND     320 321 W=0.464 ANG=45 M=0.146
MLIN      321 322 W=0.464 L=2.220
MBEND     322 323 W=0.464 ANG=45 M=0
MLIN      323 324 W=0.464 L=0.413

!      Radiating microstrip patch
MLIN      324 325 W=5.974 L=5.974

DEF1P 3 PATCH

!-----LINE FROM SPLIT S13 TO ELEMENT E3

PATCH    300
MCURVE    300 303 W^W05000 ANG=90 RAD=0.732
MLIN      303 304 W^W05000 L=2.6260
MSTEP     304 305 W1^W05000 W2^W08265
MLIN      305 310 W^W08265 L=1.3320
MCURVE    310 320 W^W08265 ANG=60.95 RAD=2
MSTEP     320 325 W1^W08265 W2^W05000
MCURVE    325 330 W^W05000 ANG=29.05 RAD=2
MLIN      330 340 W^W05000 L=2.32

DEF1P     340 S13_E3

!-----LINE FROM SPLIT S13 TO ELEMENT E1

MLIN      350 360 W^W05000 L=1.4163
MCURVE    370 360 W^W05000 ANG=90 RAD=1
MLIN      370 380 W^W05000 L=1.568
MCURVE    380 390 W^W05000 ANG=90 RAD=0.732
PATCH    390

```

DEF1P 350 S13_E1

!-----LINE FROM SPLIT S57 TO ELEMENT E5

MLIN 350 360 W^W05000 L=1.4163
 MCURVE 370 360 W^W05000 ANG=90 RAD=1
 MLIN 370 380 W^W05000 L=1.568
 MCURVE 380 390 W^W05000 ANG=90 RAD=0.732
 PATCH 390

DEF1P 350 S57_E5

!-----SPLIT S13

MTEE 100 200 500 W1^W05000 W2^W05000 W3^W05000
 S13_E3 100
 S13_E1 200

DEF1P 500 S13

!-----LINE FROM SPLIT S24 TO ELEMENT E2

PATCH 200
 MCURVE 200 210 W^W05000 ANG=90 RAD=0.732
 MLIN 210 220 W^W05000 L=4.6

DEF1P 220 S24_E2

!-----LINE FROM SPLIT S68 TO ELEMENT E6

PATCH 200
 MCURVE 200 210 W^W05000 ANG=90 RAD=0.732
 MLIN 210 220 W^W05000 L=4.6

DEF1P 220 S68_E6

!-----LINE FROM SPLIT S24 TO ELEMENT E4

MLIN 230 240 W^W05000 L=0.354
 MCURVE 250 240 W^W05000 ANG=90 RAD=1
 MLIN 250 261 W^W05000 L=1.375
 MSTEP 261 262 W1^W05000 W2^W08265
 MLIN 262 263 W^W08265 L=2.825
 MCURVE 263 264 W^W08265 ANG=17.00 RAD=2
 MSTEP 264 265 W1^W08265 W2^W05000
 MCURVE 265 270 W^W05000 ANG=73.00 RAD=2
 MLIN 270 280 W^W05000 L=0.118
 MCURVE 280 290 W^W05000 ANG=90 RAD=0.732
 PATCH 290

```

DEF1P    230 S24_E4

!-----SPLIT S24

MTEE     100 200 500 W1^W05000 W2^W05000 W3^W05000
S24_E2   100
S24_E4   200

DEF1P     500 S24

!-----LINE FROM JUNCTION J1 TO SPLIT S24

MLIN     120 130 W^W07321 L^L07321
MSTEP    130 140 W1^W07321 W2^W10000
MLIN     140 150 W^W10000 L^L10000
MSTEP    150 160 W1^W10000 W2^W05000
MLIN     160 170 W^W05000 L=2.251
MCURVE   170 180 W^W05000 ANG=90 RAD=0.704
S24      180

DEF1P     120 J1_S24

!-----LINE FROM JUNCTION J1 TO SPLIT S13

MLIN     120 130 W^W07321 L^L07321
MSTEP    130 140 W1^W07321 W2^W10000
MLIN     140 150 W^W10000 L^L10000
MSTEP    150 160 W1^W10000 W2^W05000
MLIN     160 170 W^W05000 L=0.6675
MCURVE   170 180 W^W05000 ANG=90 RAD=0.856
MCURVE   190 180 W^W05000 ANG=90 RAD=0.856
S13      190

DEF1P     120 J1_S13

!-----LINE FROM SPLIT S57 TO ELEMENT E7

PATCH   300
MCURVE   300 303 W^W05000 ANG=90 RAD=0.732
MLIN     303 304 W^W05000 L=2.6934
MSTEP    304 305 W1^W05000 W2^W06173
MLIN     305 310 W^W06173 L=1.2646
MCURVE   310 320 W^W06173 ANG=60.95 RAD=2
MSTEP    320 325 W1^W06173 W2^W07621
MCURVE   325 330 W^W07621 ANG=29.05 RAD=2
MLIN     330 340 W^W07621 L=2.32

DEF1P     340 S57_E7

```

!-----LINE FROM SPLIT S68 TO ELEMENT E8

```
MLIN      230 240 W^W07621 L=0.354
MCURVE    250 240 W^W07621 ANG=90 RAD=1
MLIN      250 261 W^W07621 L=1.375
MSTEP     261 262 W1^W07621 W2^W06173
MLIN      262 263 W^W06173 L=2.825
MCURVE    263 264 W^W06173 ANG=17.00 RAD=2
MSTEP     264 265 W1^W06173 W2^W05000
MCURVE    265 270 W^W05000 ANG=73.00 RAD=2
MLIN      270 280 W^W05000 L=0.118
MCURVE    280 290 W^W05000 ANG=90 RAD=0.732
PATCH    290
```

```
DEF1P     230 S68_E8
```

!-----SPLIT S57

```
MTEE      100 200 500 W1^W07621 W2^W05000 W3^W05000
S57_E7    100
S57_E5    200
```

```
DEF1P     500 S57
```

!-----SPLIT S68

```
MTEE      100 200 500 W1^W05000 W2^W07621 W3^W05000
S68_E8    200
S68_E6    100
```

```
DEF1P     500 S68
```

!-----LINE FROM JUNCTION J2 TO SPLIT S57

```
MLIN      120 130 W^W10000 L=3.3919
MSTEP     130 140 W1^W10000 W2^W09100
MLIN      140 150 W^W09100 L=3.4515
MSTEP     150 160 W1^W09100 W2^W05000
MLIN      160 170 W^W05000 L=0.6342
MCURVE    170 180 W^W05000 ANG=90 RAD=0.856
MCURVE    190 180 W^W05000 ANG=90 RAD=0.856
S57       190
```

```
DEF1P     120 J2_S57
```

!-----LINE FROM JUNCTION J2 TO SPLIT S68

```
MLIN      120 130 W^W10000 L=3.3919
MSTEP     130 140 W1^W10000 W2^W09100
MLIN      140 150 W^W09100 L=3.4515
```

```

MSTEP  150 160 W1^W09100 W2^W05000
MLIN    160 170 W^W05000 L=2.2177
MCURVE  170 180 W^W05000 ANG=90 RAD=0.704
S68     180

```

```

DEF1P   120 J2_S68

```

```

!-----LINE BETWEEN ELEMENT 9 AND ELEMENT 10

```

```

PATCH  100
MCURVE  100 110 W^W05000 ANG=90 RAD=1
MLIN    110 120 W^W05000 L=0.8011
MSTEP   120 130 W1^W05000 W2^W10000
MLIN    130 140 W^W10000 L^L10000
MTEE    200 140 300 W1^W10000 W2^W10000 W3^W10000
MLIN    200 210 W^W10000 L^L10000
MSTEP   210 220 W1^W10000 W2^W05000
MLIN    220 230 W^W05000 L=0.8011
MCURVE  230 240 W^W05000 ANG=90 RAD=1
PATCH  240

```

```

DEF1P   300 J3

```

```

!-----LINE BETWEEN J2 AND J3

```

```

MLIN    100 110 W^W10000 L=1
MCURVE  110 120 W^W10000 ANG=90 RAD=0.6
MLIN    120 121 W^W10000 L=0.6968
MCURVE  122 121 W^W10000 ANG=47.25 RAD=1
MSTEP   122 123 W1^W10000 W2^W05384
MCURVE  124 123 W^W05384 ANG=132.75 RAD=1
MCURVE  124 125 W^W05384 ANG=90 RAD=0.6
MLIN    125 126 W^W05384 L=0.0350
MSTEP   126 127 W1^W05384 W2^W10000
MLIN    127 172 W^W10000 L=4.4568
MCURVE  174 172 W^W10000 ANG=90 RAD=0.8
MCURVE  174 176 W^W10000 ANG=90 RAD=0.8
MCURVE  176 177 W^W10000 ANG=90 RAD=0.8
MLIN    177 178 W^W10000 L=0.755
MCURVE  180 178 W^W10000 ANG=90 RAD=0.8
MCURVE  188 180 W^W10000 ANG=90 RAD=0.82645
MLIN    188 189 W^W10000 L=0.755
MCURVE  189 200 W^W10000 ANG=90 RAD=0.82645
J3      200

```

```

DEF1P   100 J2_J3

```

```

!-----JUNCTION J2

```

```

MCROS   100 200 300 400 W1^W10000 W2^W05670 W3^W10000 W4^W10000

```

J2_S68 300
 J2_S57 100
 J2_J3 400

DEF1P 200 J2

!-----LINE FROM JUNCTION J0 TO JUNCTION J1

MBEND 100 110 W^W08826 ANG=7.86 M=0
 MLIN 110 120 W^W08826 L^L08826
 MBEND 120 130 W^W08826 ANG=-7.86 M=0
 MTEE 200 300 130 W1^W07321 W2^W07321 W3^W08826
 J1_S13 200
 J1_S24 300

DEF1P 100 J0_J1

!-----LINE FROM JUNCTION J0 TO JUNCTION J2

MCURVE 310 300 W^W07363 ANG=90 RAD=0.6972
 MLIN 310 340 W^W07363 L=3.3654
 MCURVE 340 350 W^W07363 ANG=90 RAD=0.3000
 MCURVE 360 350 W^W07363 ANG=180 RAD=0.7700
 MCURVE 360 370 W^W07363 ANG=77.13 RAD=1.7032
 MSTEP 370 371 W1^W07363 W2^W05670
 MCURVE 371 375 W^W05670 ANG=12.87 RAD=1.7032
 MLIN 375 380 W^W05670 L=2.9960
 J2 380

DEF1P 300 J0_J2

!-----CENTRE JUNCTION JC

MTEE 100 200 300 W1^W05000 W2^W08826 W3^W07363
 J0_J2 300
 J0_J1 200

! LINE FROM CENTRAL FEED POINT JC TO MAIN POWER SPLIT J0

MLIN 100 110 W^W05000 L=4.4188
 MCURVE 110 120 W^W05000 ANG=90 RAD=0.3937

! GUIDE FOR CUTTING DISK

MCROS 120 125 130 140 W1^W05000 W2=0.001 W3^W05000 W4=0.001
 MRSTUB 130 W1^W05000 L=0.01 ANG=120
 MGAP 125 126 W=0.001 S=23.268
 MCORN 126 127 W=0.001
 MCURVE 127 128 W=0.04 ANG=360 RAD=23.5

DEF1P 140 JC

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END

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