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Novel High-Efficiency Planar Antenna Array Techniques

J.G. Philippe Lafleur

A thesis submitted in the partial fulfillment of the requirements
for the degree of Master of Electrical Engineering.

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

September 1998

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submitted by
J.G. Philippe Lafleur, B.Eng.

in partial fulfillment of the requirements for the
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Chair, Department of Electronics

Thesis Supervisor

Thesis Co-supervisor

External Examiner

Carleton University

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Abstract

Novel planar array techniques, namely multilayer parasitically-fed arrays, will be introduced in this work that permit small arrays (<32 patches) to be constructed without the use of a feed network, the principal source of loss in planar arrays. These novel structures provided improved bandwidth and efficiency over conventional arrays.

Large bandwidths of 37% were achieved with 12 dB gain for a corporate parasitically-fed structure, and 17.6 dB gain with narrowband performance for a series parasitically-fed configuration, each from a single feed point.

Novel dual-frequency / dual-polarization arrays were built based on the multilayer parasitically-fed structure. The measured isolation between ports exceeded 40 dB.

A 36 dB directivity planar array was constructed at 28 GHz to verify the performance of the multilayer parasitically-fed arrays themselves placed in an array environment. Large arrays of multilayer parasitically-fed array elements reduce feed loss and complexity when compared to similar sized arrays of smaller elements.

These novel antennas are promising low profile alternatives to conventional reflector antennas for emerging broadband multimedia applications.

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Table of Contents

1. INTRODUCTION	1
1.1 BACKGROUND	1
<i>1.1.1 Antennas for Emerging Broadband Multimedia.....</i>	<i>2</i>
1.2 OBJECTIVES AND METHODOLOGY	2
1.3 THESIS ORGANIZATION	3
2. EVOLUTION OF TERMINALS FOR BROADBAND WIRELESS COMMUNICATION SYSTEMS	4
2.1 CONVENTIONAL SYSTEMS	4
2.2 ALTERNATIVE ARCHITECTURES	5
3. APPLICATIONS.....	8
3.1 LMCS USER TERMINALS	8
3.2 USER TERMINALS FOR K-KA BAND SATELLITE CONSTELLATIONS	9
4. BROADBAND PRINTED ANTENNAS: OVERVIEW	11
4.1 TRAVELLING WAVE PRINTED ANTENNAS	11
4.2 BROADBAND RESONANT ANTENNAS.....	13
<i>4.2.1 Stacked Patch.....</i>	<i>13</i>
<i>4.2.2 Quad Patch.....</i>	<i>13</i>
5. THE PARASITICALLY-FED PRINTED ARRAY	15
5.1 THE CORPORATE PARASITICALLY-FED ARRAY	15
<i>5.1.1 Outline of Concept.....</i>	<i>16</i>
<i>5.1.2 Explanation of Operation.....</i>	<i>17</i>
<i>5.1.3 Advantages</i>	<i>20</i>
<i>5.1.4 Disadvantages</i>	<i>21</i>

5.2 THE SERIES PARASITICALLY-FED ARRAY	21
5.2.1 <i>Outline of Concept</i>	22
5.2.2 <i>Explanation of Operation</i>	23
5.2.3 <i>Advantages</i>	25
5.2.4 <i>Disadvantages</i>	25
6. MULTILAYER PARASITICALLY-FED PRINTED ARRAYS	26
6.1 INTRODUCTION.....	26
6.2 MULTILAYER PARASITICALLY-FED PRINTED ANTENNA DESIGN CONSIDERATIONS	26
6.3 DIRECT COUPLING TECHNIQUES	28
6.4 OBTAINING DIFFERENT POLARIZATIONS AND POLARIZATION DIVERSITY	30
6.5 OBTAINING MULTIPLE BEAMS.....	35
6.6 SUMMARY OF NOVEL TECHNIQUES.....	38
6.7 CORPORATE PARASITICALLY-FED DESIGNS	38
6.7.1 <i>Three-layer Slot-Coupled Linearly-Polarized Antenna</i>	39
6.7.2 <i>Four-layer Slot-Coupled Linearly-Polarized Antenna</i>	60
6.8 SERIES PARASITICALLY-FED DESIGNS.....	66
6.8.1 <i>Eight-Patch Slot-Coupled Linearly-Polarized Antennas</i>	67
6.8.2 <i>12-Patch Slot-Coupled Linearly-Polarized Antennas</i>	72
6.8.3 <i>32-Patch Slot-Coupled Linearly-Polarized Antennas</i>	79
6.9 ARRAY TECHNIQUES.....	87
6.9.1 <i>Configuration</i>	87
6.9.2 <i>Dual-Polarized / Dual-Frequency Arrays</i>	94
6.9.3 <i>Results</i>	99
7. SUMMARY AND FUTURE WORK.....	124

List of Figures

FIGURE 1: CONVENTIONAL SYSTEM DIAGRAM FOR FIXED TERMINALS	4
FIGURE 2: SPATIAL POWER COMBINING ILLUSTRATED.....	6
FIGURE 3: G/T IMPROVEMENT WITH LNAs	6
FIGURE 4: EXAMPLES OF TRAVELING WAVE PRINTED ANTENNAS.....	12
FIGURE 5: THE STACKED PATCH (SHOWN WITHOUT FEED POINT)	13
FIGURE 6: THE QUAD-PATCH ANTENNA (SHOWN WITHOUT FEED).....	14
FIGURE 7: FAILED EXTENSIONS OF THE QUAD-PATCH ANTENNA	16
FIGURE 8: THE CORPORATE PARASITICALLY-FED ANTENNA.....	17
FIGURE 9: THE PHASE DISTRIBUTION FOR A SIMPLE MICROSTRIP PATCH	18
FIGURE 10: SIMPLIFIED EXPLANATION OF THE COUPLING MECHANISM	19
FIGURE 11: PHASE ERRORS ON THE APERTURE FOR A NINE-PATCH ARRAY	20
FIGURE 12: THE SERIES PARASITICALLY-FED ARRAY	22
FIGURE 13: SIMPLIFIED EXPLANATION OF THE OPERATION OF THE SERIES-FED ARRAY	23
FIGURE 14: SLOT COUPLING TO ISOLATE THE FEED NETWORK AND IMPROVE CROSS-POLARIZATION PERFORMANCE	29
FIGURE 15: A LINE-FED ARRAY, SUITABLE FOR SIMPLE ARRAYS.....	29
FIGURE 16: PROBE COUPLING, THE SIMPLEST OPTION	30
FIGURE 17: FEEDING FOR LINEAR POLARIZATION.....	31
FIGURE 18: THREE POINT FEED SYSTEM FOR HIGH ISOLATION.....	32
FIGURE 19: HIGH ISOLATION TECHNIQUES FOR DUAL-LINEAR POLARIZATION SYSTEMS	33
FIGURE 20: FEEDING FOR CIRCULAR POLARIZATION	34
FIGURE 21: A BROADSIDE RADIATING SERIES PARASITICALLY-FED COLUMN ARRAY	35
FIGURE 22: AN OFFSET BEAM SERIES PARASITICALLY-FED COLUMN ARRAY	36
FIGURE 23: A MULTIPLE BEAM ARRAY	36
FIGURE 24: A MULTIPLE BEAM ARRAY USED AS AN LMCS REPEATER	37
FIGURE 25: THE THREE-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA	40
FIGURE 26: LAYER DEFINITION FOR THE THREE-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA.....	40
FIGURE 27: THE FREQUENCY RESPONSE OF THE THREE-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA ($ S_{11} $ AND GAIN).....	42
FIGURE 28: THE INPUT IMPEDANCE OF THE THREE-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA ..	43
FIGURE 29: RADIATION PATTERN AT THE CENTRE FREQUENCY FOR THE THREE-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA	44
FIGURE 30: RADIATION PATTERN AT THE LOWER BAND EDGE FOR THE THREE-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA	45
FIGURE 31: RADIATION PATTERN AT THE UPPER BAND EDGE FOR THE THREE-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA	46
FIGURE 32: THE MEASURED FREQUENCY RESPONSE FOR THE THREE-LAYER SLOT-COUPLED LINEARLY- POLARIZED ANTENNA	47
FIGURE 33: MEASURED E-PLANE RADIATION PATTERN AT THE CENTRE FREQUENCY FOR THE THREE-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA.....	48
FIGURE 34: MEASURED H-PLANE RADIATION PATTERN AT THE CENTRE FREQUENCY FOR THE THREE-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA.....	49
FIGURE 35: CURRENT DISTRIBUTION OF THE THREE-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA AT THE CENTRE FREQUENCY	50
FIGURE 36: CURRENT DISTRIBUTION OF THE THREE-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA AT THE UPPER BAND EDGE.....	52
FIGURE 37: A TOP VIEW OF THE REDUCED-SIZE-CENTRE-PATCH THREE-LAYER SLOT-COUPLED LINEARLY- POLARIZED ANTENNA	53
FIGURE 38: THE FREQUENCY RESPONSE OF THE THREE-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA WITH A SMALLER CENTRE PATCH ($ S_{11} $ AND GAIN).....	54

FIGURE 39: A TOP VIEW OF THE THREE-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA WHOSE CENTRE PATCH WAS REPLACED BY A MULTITUDE OF SMALL PATCHES.....	55
FIGURE 40: THE FREQUENCY RESPONSE OF THE THREE-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA ($ S_{11} $ AND GAIN) WHOSE CENTRE PATCH WAS REPLACED BY A MULTITUDE OF SMALL PATCHES.....	56
FIGURE 41: THE MEASURED FREQUENCY RESPONSE OF THE THREE-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA ($ S_{11} $, GAIN, AND CROSS-POL.) WHOSE CENTRE PATCH WAS REPLACED BY A MULTITUDE OF SMALL PATCHES	57
FIGURE 42: THE MEASURED $ S_{11} $ RESPONSE OF THE THREE-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA WHOSE CENTRE PATCH WAS REPLACED BY A MULTITUDE OF SMALL PATCHES OVER A LARGE BAND	58
FIGURE 43: A FOUR-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA.....	60
FIGURE 44: LAYER DEFINITION FOR THE FOUR-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA	61
FIGURE 45: THE FREQUENCY RESPONSE OF THE FOUR-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA.....	62
FIGURE 46: CURRENT DISTRIBUTION FOR THE FOUR-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA.....	63
FIGURE 47: THE RADIATION PATTERN OF THE FOUR-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA AT 18 AND 22 GHz	64
FIGURE 48: THE RADIATION PATTERN OF THE FOUR-LAYER SLOT-COUPLED LINEARLY-POLARIZED ANTENNA AT 24 AND 26.5 GHz	65
FIGURE 49: THE EIGHT-PATCH SLOT-COUPLED LINEARLY-POLARIZED ANTENNA	67
FIGURE 50: LAYER DEFINITION FOR THE EIGHT-PATCH SLOT-COUPLED LINEARLY-POLARIZED ANTENNA.....	68
FIGURE 51: THE FREQUENCY RESPONSE OF THE EIGHT-PATCH SLOT-COUPLED LINEARLY-POLARIZED ANTENNA.....	69
FIGURE 52: THE RADIATION PATTERN OF THE EIGHT-PATCH SLOT-COUPLED LINEARLY-POLARIZED ANTENNA AT THE CENTRE FREQUENCY	70
FIGURE 53: THE CURRENT DISTRIBUTION OF THE EIGHT-PATCH SLOT-COUPLED LINEARLY-POLARIZED ANTENNA AT THE CENTRE FREQUENCY	71
FIGURE 54: THE 12-PATCH SLOT-COUPLED LINEARLY-POLARIZED ANTENNA.....	73
FIGURE 55: THE FREQUENCY RESPONSE OF THE 12-PATCH SLOT-COUPLED LINEARLY-POLARIZED ANTENNA	74
FIGURE 56: THE RADIATION PATTERN OF THE 12-PATCH SLOT-COUPLED LINEARLY-POLARIZED ANTENNA AT THE CENTRE FREQUENCY	75
FIGURE 57: THE CURRENT DISTRIBUTION OF THE 12-PATCH SLOT-COUPLED LINEARLY-POLARIZED ANTENNA AT THE CENTRE FREQUENCY	76
FIGURE 58: MEASURED FREQUENCY RESPONSE FOR THE 12-PATCH SLOT-COUPLED LINEARLY-POLARIZED ANTENNA	77
FIGURE 59: MEASURED RADIATION PATTERN FOR THE 12-PATCH SLOT-COUPLED LINEARLY-POLARIZED ANTENNA AT THE CENTRE FREQUENCY	78
FIGURE 60: THE 32-PATCH SLOT-COUPLED LINEARLY-POLARIZED ANTENNA.....	80
FIGURE 61: THE FREQUENCY RESPONSE OF THE 32-PATCH SLOT-COUPLED LINEARLY-POLARIZED ANTENNA	81
FIGURE 62: THE RADIATION PATTERN OF THE 32-PATCH SLOT-COUPLED LINEARLY-POLARIZED ANTENNA AT THE CENTRE FREQUENCY	82
FIGURE 63: CURRENT DISTRIBUTION OF THE 32-PATCH SLOT-COUPLED LINEARLY-POLARIZED ANTENNA AT THE CENTRE FREQUENCY	83
FIGURE 64: THE MEASURED FREQUENCY RESPONSE OF THE 32-PATCH SLOT-COUPLED LINEARLY-POLARIZED ANTENNA ($ S_{11} $, GAIN)	84
FIGURE 65: THE MEASURED RADIATION PATTERN AT THE CENTRE FREQUENCY FOR THE 32-PATCH ANTENNA.....	85

1. Introduction

1.1 Background

In the last ten years there has been a dramatic increase in the use of wireless communications. Cellular phones, PCS (personal communication system), pagers, and identification tags are examples of the many wireless systems in use today. These systems operate at L-Band, and as the number of systems grow, the band becomes more congested. Having relatively narrow bandwidth requirements, these systems can afford to operate in this region of the spectrum.

For multimedia and other broadband applications, K- and Ka-bands have been allocated. Relatively broad bandwidths are available, with the principal challenge being the development of communications equipment capable of providing adequate performance at these frequencies, at a low cost. Systems such as LMCS/LMDS (local multipoint communication system/local multipoint distribution system), and Ka-band satcom (satellite communications) are slated to operate at these frequencies. These systems will revolutionize multimedia by providing broadband wireless communications to the home.

This new generation of broadband communications requires a compact, low cost terminal. Among the most costly portions of this terminal are the RF (radio frequency)

electronics and the antenna. The antenna portion is the focus of this thesis, with the integration of RF electronics on the antenna considered as well.

1.1.1 Antennas for Emerging Broadband Multimedia

Emerging broadband multimedia wireless services will require fixed beam antenna subsystems for their user terminals and base stations. These antennas will have to be low cost, efficient and high performance.

Low profile antennas are widely used as base station antennas, due to their ability to produce sectored beams. Their efficiency, although poor, is taken as a given and systems are designed to accommodate this shortcoming. This is possible at the base station level due to the relative cost insensitivity.

At the user terminal level, a simple pencil beam radiation pattern is required. Conventional low-profile antennas, although esthetically pleasing, have not been able to match the efficiency of the reflector based solution. Thus, for fixed point-to-point user terminals, the use of reflectors has generally been assumed.

1.2 Objectives and Methodology

The objective of this thesis was to make important advancements in low-profile antenna arrays to improve their efficiency. To make such a fundamental change to low

profile array technology, the investigation had to begin at the array element (or subarray) level. These findings would then be applied to an appropriate array architecture.

The first objective was to investigate high efficiency broadband and narrowband antennas of varying complexity. The second objective was to design a large array suitable for fixed terminal based applications using the aforementioned antennas.

1.3 Thesis Organization

The contents of this thesis is divided into the following chapters. Chapter 2 discusses the evolution of terminals for broadband wireless communication systems. Chapter 3 discusses the emerging systems and their specific user terminal requirements. Chapter 4 presents an overview of existing techniques for achieving broadband antenna performance. Chapter 5 provides an outline of the concept and description of the operation of the multilayer parasitically-fed printed antenna element. This is presented before the results of the design work and measurements to help the reader to understand all major aspects of these novel antenna structures before being asked to interpret the results. Chapter 6 presents the designs and results of the multilayer parasitically-fed arrays investigated. Finally, Chapter 7 concludes with a summary of the thesis work and suggestions for further studies.

2. Evolution of Terminals for Broadband Wireless Communication Systems

2.1 Conventional Systems

Despite important conceptual advancements in fixed terminal architectures in recent years, most fixed-beam commercial systems are, at the RF system level, the same as their predecessors from the 1970s. A typical system is presented below, and consists of a parabolic reflector connected to the RF electronics, via a diplexer or RF switch.

Figure 1: Conventional system diagram for fixed terminals

This antenna configuration offers several advantages such as low cost, high gain, and high efficiency. Until now, these requirements could not be met by low profile antenna arrays. This type of system, however, does not make use of the full potential of low profile printed array technology, and so alternative architectures should be considered.

Alternatives to reflectors do exist, such as reflect arrays and dielectric lens antennas. These configurations are not low profile, however, as each requires a focal point feed element. Slotted waveguide arrays are low profile, but are narrowband.

2.2 Alternative Architectures

The important conceptual advancements in fixed terminal architectures alluded to in section 2.1, are spatial power combining [1], and G/T (gain-to-noise-temperature ratio) improvement with multiple LNAs (low noise amplifiers) [2].

Spatial power combining is essentially the integration of parts of a transmit architecture, which involves spreading the high power amplification throughout the array. Replacing the single high power amplifier by several smaller amplifiers, eliminates the high power problems of a large high power amplifier at the antenna input, the combining losses incurred when many amplifiers are placed in parallel to provide higher output power, and the linearity problems of operating an amplifier near the compression point. This higher linearity can be achieved, since the amplifiers can be operated well below their compression point and the reduction in output power can be compensated for by using more amplifiers. This has the additional benefit of eliminating the impact of array feed network losses on the output power. Spatial power combining is explained graphically in Figure 2. Spatial power combining is a clear advantage of printed array technology over reflectors.

Figure 2: Spatial power combining illustrated

Another important technique is G/T improvement with LNAs. Although power level considerations are less important here, placing multiple LNAs near array elements masks the impact on noise temperature of the combining network losses, thereby improving G/T. This concept, illustrated in Figure 3, makes printed arrays more competitive with reflectors, which have no combining loss, for reception.

Figure 3: G/T improvement with LNAs

A component count problem is rapidly encountered with these techniques when large arrays are considered. This leads to compromises such as placing an amplifier near a group of elements, which leads to a degradation in performance due to the feed network

losses within the subarray degrading performance. Ideally, a subarray with little or no loss would eliminate this problem. This is exactly the need that this research will address.

3. Applications

Although there are a vast number of fixed terminal applications, there are few that have the far reaching impact and large scale of LMCS (LMDS also implied from here on) and the K-Ka band satellite constellations.

3.1 *LMCS User Terminals*

With all the frequencies allocated and the companies in place, LMCS is set to be deployed in North America. LMCS is slated to operate at approximately 28 GHz for both transmit and receive channels. Polarization diversity will be used to assist in channel separation for this full duplex system.

LMCS resembles cellular telephony in that it uses geographically distributed base stations to handle groups of customers within cells, approximately 4 km in diameter [3]. Frequency reuse between non-adjacent cells is also employed. LMCS differs from cellular telephony in that it is aimed at high-speed data services to fixed terminals, rather than mobile voice communications.

Proposed terminal configurations for LMCS closely follow the architecture illustrated in Figure 1. A parabolic reflector with about 35 dBi gain is connected to an OMT (ortho-mode transducer) which serves as the diplexer, and the transmit and receive electronics are connected to the OMT.

Requirements for LMCS are particularly stringent when it comes to isolation between the transmit and receive channels as well as cross-polarization levels. Isolation between channels must be high enough to prevent the output amplifiers from saturating the sensitive receive electronics.

These requirements are very difficult to achieve with printed technology, the most difficult being the isolation. This high degree of isolation can be achieved, however, by employing the novel array techniques discussed in this thesis. This places printed antenna technology as a potential front-running LMCS antenna solution for the first time.

3.2 User terminals for K-Ka band Satellite Constellations

Within a few years, we can expect the launching of a myriad of K-Ka band global satellite networks providing worldwide access to a multitude of communications services. In fact, there are over 55 systems proposed varying from one and two satellites, up to 840 satellites [4]. The majority of these new systems are geosynchronous, meaning that fixed-beam antennas are required. The low and medium earth orbit systems require not only many satellites to provide the desired capacity, due to their relatively close position to the earth, but also user terminal antennas to track the satellites in their polar orbits. This dramatically increases the cost and complexity of these terminals.

These satellite systems operate with a downlink at 20 GHz and uplink at 30 GHz. This large frequency separation makes isolation between channels relatively easy, but makes aperture collocation for transmit and receive somewhat more difficult. A final design challenge may be high antenna gain requirements, on the order of 40-50 dB.

Techniques described in this thesis address the collocation issue, and, with the advent of higher power satellites, the gain requirement may come down to a level realizable by a printed array.

4. Broadband Printed Antennas: Overview

The two key antenna performance parameters considered in this work were; efficiency, for large apertures, and bandwidth. At present, efficient printed structures capable of providing high gain do not exist. Therefore, the focus of this overview is shifted to point out some existing broadband techniques.

Historically, when considering printed broadband (>30% bandwidth) antennas, travelling wave, or non-resonant antennas, were the only solution. With recent developments, some resonant antenna structures are now among those to be considered. They are the stacked-patch and quad-patch antennas.

4.1 *Travelling Wave Printed Antennas*

Travelling wave antennas are best described as antennas whose current and voltage distributions are not based on standing waves. These antennas are designed to have voltage and current distributions consisting primarily of a single wave, not as a sum of incident and reflected waves. Because of this relatively uniform distribution, the frequency dependence of the phase is largely removed thereby permitting broadband performance.

There are two major types of travelling wave antennas, surface wave antennas and leaky wave antennas. Surface wave antennas are antennas which radiate power from

discontinuities in the structure that interrupt a bound wave on the antenna surface. Leaky wave antennas couple power to free space in small increments along their length in either a discrete or continuous manner [5]. Examples of surface wave and leaky wave antennas are rampart line antennas and a simple transmission line respectively. For the simple transmission line leaky wave antenna, the dominant “quasi” TEM mode does not radiate, but the higher order modes become leaky when the propagation constant is less than the free space wave number [6].

Figure 4: Examples of traveling wave printed antennas

Surface wave antennas generally radiate parallel to the structure, which for printed arrays, means end-fire radiation. This limits their applicability for the terminals described in section 3. Leaky wave antennas have beams that scan with frequency, from boresight to the ground plane. In addition, both of these techniques make use of transmission lines for their radiating structures, thereby making them inherently lossy due to the high concentration of fields around the transmission line structure.

4.2 Broadband Resonant Antennas

Typically, resonant printed antennas are relatively narrow band. This can be alleviated by structures with multiple resonances. The following techniques employ multiple resonances to broaden bandwidth.

4.2.1 Stacked Patch

The stacked microstrip [7] [8] patch alleviates the bandwidth limitation by providing a broad-bandwidth element. The idea behind the stacked patch is to have two or more patches stacked on top of each other. Each successively higher patch gets smaller. To a first order, each smaller patch uses the one beneath it as its ground plane, and radiates around the one above it. This technique broadens bandwidth, but does not increase gain, as the patches all have similar radiation characteristics. Bandwidths achieved with this technique can reach 40%. The stacked patch is shown in Figure 5.

Figure 5: The stacked patch (shown without feed point)

4.2.2 Quad Patch

The quad-patch antenna [9], consists of a four element subarray electromagnetically coupled to a driven patch on a lower substrate, as shown in Figure 6.

The four upper patches are fed through a coupling area where their corners are in close proximity to the corners of the lower patch.

Figure 6: The quad-patch antenna (shown without feed)

This technique is reported to have reached bandwidths up to 25% for bandwidth-optimized structures and directivities up to 12 dB for directivity-optimized structures. Typical results are 11 dB directivity with 15% bandwidth.

This scheme is a partial answer to the low feed loss subarray problem discussed at the end of Section 2.2. It would still be more desirable to provide broader impedance bandwidths and more directivity at the subarray level.

Fundamentally, this solution does not adequately increase the aperture size that can be fed from a single point.

5. The Parasitically-Fed Printed Array

Clearly, the quad-patch antenna made progress in increasing array efficiency and bandwidth, but to make planar arrays capable of competing with reflector based technology, another breakthrough was required.

With this as the research focus, two categories of antennas were developed. The first, broadband parasitically-fed architectures, are somewhat analogous to a corporate-fed array, while the second is a simpler parasitically-fed architecture, analogous to a series-fed array.

5.1 *The Corporate Parasitically-Fed Array*

Early in this work, simple extensions of the quad-patch antenna were tried, feeding a patch on a higher layer from one corner of a patch on a lower level. The antenna shown in Figure 7a failed due to aperture phase problems caused by the lower layer patches radiating through. The second extension, Figure 7b, failed due to the overcrowding of the patches on the top layer.

Figure 7: Failed extensions of the quad-patch antenna

5.1.1 Outline of Concept

The breakthrough that made the extension of the quad patch possible was a departure from the patch-to-patch single corner feed, to multiple patches coupling to feed a single patch.

As can be seen in Figure 8, the four outer patches on the top layer are each fed by two patches on the layer below it, with the centre patch on the top layer being fed by all four patches on the layer below it.

Figure 8: The corporate parasitically-fed antenna

5.1.2 Explanation of Operation

The coupling mechanism [10] is the fringing field. Coupling energy to nearby patches of different dimensions introduces additional resonances, which, when designed carefully, smooth out the impedance characteristic of the antenna and thereby provide a larger impedance bandwidth. As is the case with the stacked patch described in section 4.2.1, the resonances can be adjusted with the patch dimensions, but the mutual coupling inherent in this design makes the dimensions interdependent. A resonance cannot be associated with a single patch dimension.

To understand the way in which parasitically-fed arrays couple energy from one patch to another, the operation of a simple microstrip patch must first be considered. Using the simplest of the microstrip patch analogies, the open-circuited transmission line, the essence of the phase distribution of a microstrip patch can be described. As can be

seen in Figure 9, the patch is approximately half of a guided wavelength long in the resonant direction, thereby creating a 180 degree phase shift between the current on the top and bottom edges of the patch. The currents near the edges in the non-resonant direction are in phase. Although having the currents in the non-resonant direction in phase agrees with the transmission line analogy, it is actually due to the excitation coupling primarily to the lowest order mode of the patch.

Figure 9: The phase distribution for a simple microstrip patch

If another patch is brought near this patch, parasitic coupling from this patch to the next will occur. The phase of the driven patch nearest to the parasitically coupled patch determines the phase of the coupled patch. If four patches are placed near the top and bottom edges of the fed patch, as is the case for the quad-patch antenna, the patches near the top edge will be coupled to in-phase relative to each other and 180 degrees out of phase from the bottom edge. The 180-degree phase shift in the resonant direction is used to compensate for the spatial inversion of the four top patches relative to each other as illustrated in 10a) (patches 1 and 2 mirror 3 and 4). This leads to an in-phase aperture.

For the parasitically-coupled arrays discussed in this thesis, the energy is coupled to the radiators on the top layer in phase, as is the case with the quad-patch antenna, but, in contrast, some of the patches on the top layer are fed with a balanced (push-pull) feeding mechanism and some in phase. With a larger in-phase aperture, a larger directivity is obtained. This explanation, while useful for understanding how the coupling mechanism works, is not in itself complete. This is shown visually in Figure 10 b).

Figure 10: Simplified explanation of the coupling mechanism

As will be shown in section 6.7.1.2 of this thesis, some behaviour is not accounted for with the simplified model. For instance, Figure 35 shows that coupling to the centre patch is noticeably out of phase with respect to the other patches on the radiating layer. Also, if four more patches were to be added to the top layer to form a three-by-three array, these would be out of phase as well. This is summarized in Figure 11.

Figure 11: Phase errors on the aperture for a nine-patch array

A pattern in the phase distribution is easily found: the patches that are equidistant from the feed point are in phase. This would tend to suggest a radial-wave effect where the location relative to the wave's origin becomes important.

5.1.3 Advantages

As will be shown in section 6.7.1.2, the advantages of this scheme over the quad-patch antenna are:

- Larger bandwidth
- Higher directivity
- Higher efficiency (lower loss than feeding multiple quads).

This improved efficiency is due to decreased dielectric loss. Dielectric heating losses (substrate loss) are directly related to field concentrations. The lower losses inherent to all parasitically coupled arrays, as compared to conventional feeding schemes, are due to the fields being less confined.

5.1.4 Disadvantages

The primary disadvantage is cost. Multiple layers increase material and processing costs directly.

5.2 *The Series Parasitically-Fed Array*

With the corporately-fed configuration, continuing to increase the directivity requires more and more layers, which dramatically increases the cost and complexity of the antenna. In an effort to increase aperture area without increasing the number of layers, the series parasitically-fed array was devised.

Analogous to a series line-fed array, the first element radiates a portion of the energy incident on it and passes the remainder to the next element. To obtain a proper radiation pattern, the phases must be similar across the array, implying a phase shift that is an integer multiple of 360 degrees between elements.

As is the case with the series line-fed array, bandwidth is narrowed, since phase relationship between the radiators can only be maintained for a narrow range of frequencies.

5.2.1 Outline of Concept

Employing the same techniques as in the corporate parasitically-fed array, coupling from multiple patches is also used. The difference is that the structure is flattened out by coupling from a patch on the upper layer, down to a patch on the lower layer and back up to the next patch on the upper layer. This is shown in Figure 12.

Figure 12: The series parasitically-fed array

By forcing the energy to travel in lines as opposed to branching out, the series structure is created.

5.2.2 Explanation of Operation

As is the case for the corporate parasitically-fed arrays, the series concept uses a fringing field, corner-to-corner coupling mechanism. Again, like the corporate architecture, phase inversion across the patch in the resonant direction, and its equal phases in the non resonant direction are used to obtain an in-phase aperture distribution. This is explained visually in Figure 13.

Top Two Layers of 12 Patch "Series" Fed Array

Notes: - Shaded patches are on the top layer
 - Coupling arrows direction refers to the phase of the coupling, up is 0 degrees and down is 180 degrees

Figure 13: Simplified explanation of the operation of the series-fed array

This structure is narrower band than the corporate array because additional resonances are not introduced. Bandwidth is further limited by the requirement for the patches to provide a 360-degree phase shift in the resonant direction and 0 degrees in the non-resonant direction. As expected, E-plane coupling, because of the 360-degree phase shift required, is narrower band than H-plane coupling with its 0 degree phase shift.

As will be shown in section 6.8.1.2, the simplified model works very well for the series structures. The fundamental behaviour is well described without having to consider additional effects, as was the case with the corporate arrays. The absence of radial distance effects could be explained by the fact that the patches are located either at equal distances from the feed point or they are separated by an additional guided wavelength, thereby masking these effects.

Obtaining a reasonably uniform phase distribution is readily accomplished, but obtaining uniform amplitude is more difficult. Coupling can be increased from patch to patch by increasing the amount of overlap. This provides only a limited amount of tuning, which precludes obtaining a uniform distribution for very large apertures. The effect is that outer patches receive less power than inner patches. This limits the maximum array size, as outer elements will contribute less and less power as the array size is increased. This will be discussed in section 6.8.3.2.

5.2.3 Advantages

As will be shown in sections 6.8.1.2, 6.8.2.2 and 6.8.3.2, the advantages of this technique are that one can obtain:

- A large aperture with high efficiency
- Superior control of the phase relationships between the radiating elements
- Simpler structure and lower cost.

As is the case with the corporate parasitically-fed arrays, this technique benefits from the lower dielectric losses inherent to all parasitically-coupled arrays due to the fields being less confined.

5.2.4 Disadvantages

The bandwidth of this array configuration is narrower than that of the corporate configuration. The increase in the number of phase rotations narrows the range of frequencies over which the patches are in phase.

6. Multilayer Parasitically-Fed Printed Arrays

6.1 Introduction

With the background information regarding these novel antennas in place, and their operation explained, this chapter will present the key considerations and techniques to maximize the performance and versatility of the antennas. Beginning at section 6.7, seven different antenna designs will be presented, each revealing a different aspect of these antennas. It is these sections that support the claims made about antenna performance in the preceding chapters.

With the discovery of corporate and series multilayer parasitically-fed printed arrays, the challenge was to design antennas that would take full advantage of these characteristics, and to examine the extent to which these techniques could be taken. There would be limits as to how large the apertures could become, but these limits needed to be defined.

6.2 Multilayer Parasitically-Fed Printed Antenna Design Considerations

When designing a multilayer parasitically-fed printed antenna, there are several aspects to be considered, the most important of which are gain and bandwidth. From these, the choice between series and corporate parasitically-fed structures can be made. Unless the bandwidth requirement warrants it, the series technique should be used as it is

less expensive, and provides a more easily adjusted aperture distribution, can offer a larger aperture area, and has higher aperture efficiency.

The maximum aperture size is restricted by coupling limitations (see section 6.8.3). The aperture size that can be realized by a parasitically-fed printed array is limited to less than 32 patches. Large aperture (>20 dB directivity) areas can only be realized by placing the elements in an array. With this relatively high-gain and low-loss array element, however, achieving acceptable efficiency numbers is facilitated, as discussed at the end of section 2.2.

Other important considerations include the desired radiation pattern such as the shape of the main beam, sidelobe levels, the backlobe level and cross-polarization levels. For beam shapes, corporate parasitically-fed antennas can provide narrow or wide broadside beams, while series parasitically-fed antennas can provide broadside or offset, narrow or sectored beams. As will be seen in the sections following 6.7, due to inherent tapering in the aperture distribution, sidelobe levels are easily brought to below -15 dB, and careful design can further minimize these, at the expense of aperture efficiency. Clearly, the antennas can be used in arrays if larger arrays are required. Once in an array, tailoring the radiation pattern is relatively simple.

Finally, cost and losses have to be considered. These two factors are related because the primary source of loss is the dielectric material, which is also a significant

contributor to cost. Less expensive materials can be used for the patch layers since the lower field concentrations make them inherently less susceptible to loss, as discussed in section 5.1.2. If the antenna being designed is an array of subarrays, a low loss substrate must be used for the feed network layer because of the extensive use of loss-susceptible transmission lines.

6.3 Direct Coupling Techniques

Another design consideration is the type of direct coupling that is to be used to feed the multilayer parasitically-fed antennas. If a complex or active array of these antennas is to be considered, slot coupling [11] should be used to feed the elements. This reduces the number of feedback coupling paths by isolating the feed network and devices from the radiating elements, and simplifies manufacturing, since no probes are required. Slot coupling also ensures low cross-polarization components. Slot coupling is shown in Figure 14.

Figure 14: Slot coupling to isolate the feed network and improve cross-polarization performance

For simple arrays, when cross-polarization level specifications are not too stringent, line-fed solutions are the most attractive solution. An example of a line-fed array is shown in Figure 15.

Figure 15: A line-fed array, suitable for simple arrays

If a single multilayer parasitically-fed antenna is to be used, and cross-polarization requirements are not very demanding, probe feeding could be used. Probe coupling is not as simple for multilayer structures due to the blind connection that has to be made. A connector could be wedged in a large printed circuit via to make electrical contact, or a soldering access hole could be drilled in the top layer so that a conventional probe attachment could be used. Probe coupling is illustrated in Figure 16.

Figure 16: Probe coupling

6.4 Obtaining Different Polarizations and Polarization Diversity

A variety of polarization schemes is being employed in wireless communications systems. Most often, the polarization is selected to optimize propagation characteristics, such as the use of circular polarization to mitigate the effects of Faraday rotation in the ionosphere or horizontal polarization to minimize multipath from ground reflections.

Many choices are available to the system designer, such as linear, dual-linear, circular, or dual-circular. Linear polarization is the easiest to achieve, and can be achieved with a single slot, probe or line as shown in Figure 17. Slot coupling provides the best cross-polarization performance, with probe and line coupling being similar.

Figure 17: Feeding for linear polarization

Some full-duplex systems employ polarization diversity to help to isolate the two channels. LMCS uses dual-linear polarization. Polarization diversity requires extremely low cross-polarization levels, and a high degree of isolation between the transmit and receive channels to prevent unintentional coupling from the high-power transmitter to the sensitive receiver.

The best solution offered to this point was to use a three-point feed system as shown in Figure 18 [12]. Here a balanced, two point-feeding system is used for the

horizontal polarization and a single point feed for the vertical polarization. The balanced port causes equal and out of phase coupling to the vertical polarization and coupling from the vertical polarization is combined out of phase at the horizontal polarization's port. No more than 30 dB of isolation could be achieved. The feed points are simply too close to each other to achieve more than 30 dB isolation. This technique can be applied to any aperture-coupled patch or aperture-coupled parasitically-fed array.

Figure 18: Three-point feed system for high isolation

The series parasitically-fed array allows for an improvement to this technique. A higher degree of isolation can be achieved by using a three-point feed system and by placing the feed points on different drive patches. This can only be done if the series parasitically-fed antenna is used, because it allows for multiple-driven patches. The three-point feed system provides better than 40 dB isolation between the two ports, as is shown in Figure 19. An example can be found in section 6.9.2.

Figure 19: High isolation techniques for dual-linear polarization systems

Circular polarization has to be achieved by a two (or more) point feeding system in quadrature. The 90-degree phase difference can be realized in several ways, such as a hybrid coupler or a power divider and a 90-degree delay line. This is illustrated in Figure 20.

Figure 20: Feeding for circular polarization

Although not the only factor, isolation between polarizations impacts cross-polarization for circularly polarized antennas. Low cross-polarization would be critical in a dual-circularly polarized system.

6.5 Obtaining Multiple Beams

By taking advantage of the multiple feed point capability of the series configuration, and taking care to ensure the correct phase relationship between the radiators, multiple beams can be achieved.

To obtain a better understanding, consider the antenna shown in Figure 21. If the patches were designed to change phase by an integer multiple of 360 degrees relative to each other, then moving the feed point to the patch on the left or right of the centre patch should not affect the beam angle, which will stay at boresight.

Figure 21: A broadside radiating series parasitically-fed column array

If the phase relationship is not an integer multiple of 360 degrees, then changing the position of the feed point determines the phase progression across the array and hence the phase angle. This is shown in Figure 22.

Figure 22: An offset beam series parasitically-fed column array

If we now combine the properties of beam scanning and multiple feed points, we can obtain a multiple beam array, as illustrated in Figure 23. This is a narrowband technique, since the phase relationships would change with frequency.

Figure 23: A multiple beam array

It should be noted that the isolation between the two slots will not be very high, because the slots are of the same polarization. With orthogonal polarizations, isolation would be much improved.

If different polarizations were used, isolation would be sufficient for this structure to be used as a LMCS repeater as shown in Figure 24. This would allow increases of the signal level in faded areas, but two successive repeaters would have to be used to return the polarization to its initial value.

Figure 24: A multiple beam array used as an LMCS repeater

6.6 Summary of Novel Techniques

To restate which of the techniques described in sections 6.3 through 6.5 apply to which architectures, a summary is included in Table 1.

Technique	Applicability to corporate parasitically-fed arrays	Applicability to series parasitically-fed arrays
Slot, probe and line coupling	yes	yes
Multiple drive patches	no	yes
Beam squinting and multiple beams	no	yes

Table 1: Summary of Novel Techniques

6.7 Corporate Parasitically-Fed Designs

With the fundamentals of parasitically-fed arrays now introduced, a few examples of corporate parasitically-fed arrays will be shown. They will be presented in order of increasing complexity.

The antennas presented in this thesis were designed to have a centre frequency around 20 GHz, but these could be tailored to any frequency between 1 and 60 GHz. At lower frequencies, the feed network loss savings are not as significant, thereby making other solutions equally attractive. At frequencies above 60 GHz, manufacturability becomes a concern as dimensions and tolerances decrease.

Slot coupling was used in all the designs in this thesis. This is the method of choice, particularly if the intention is to eventually use the antennas to form an array, as discussed in section 6.3.

6.7.1 Three-layer Slot-Coupled Linearly-Polarized Antenna

The first example to be presented is the three-layer corporate parasitically-fed array. This was among the first antennas designed using the novel techniques discussed in this thesis. Understanding this antenna helped to develop conceptual models, which were key to the development of antennas of this type.

6.7.1.1 Configuration

A top view and a bottom view of this antenna are shown in Figure 25. This and other figures like it represent the antenna as it would appear with invisible dielectric substrates. The slot is black, the feed line is white, and the patches on the radiating (top) layer are dark grey, with lower patches being successively lighter.

Figure 25: The three-layer slot-coupled linearly-polarized antenna

There are three layers, not counting the feed network and slot layers. The first layer consists of the driven patch, there are four patches on the second layer, and five patches on the radiating layer. The ordering of the layers along with substrate selection is shown in Figure 26.

Figure 26: Layer definition for the three-layer slot-coupled linearly-polarized antenna

Although largely a matter of optimization, some reasoning for the choice of substrates exists. The 25 mil feed substrate was chosen to provide adequate thickness for carrier mounted active devices to be integrated at a later date. The high relative dielectric constant of 10 chosen for this layer would facilitate the compact integration of electronics. The substrate below the drive patch was optimized to be thick enough to provide an adequate impedance bandwidth while being thin enough to ensure proper coupling from the slot. A low relative dielectric constant of 2.94 was chosen for this layer to maintain a wide bandwidth. The layer beneath the quad-patch layer was chosen to be as electrically thin as possible to provide maximum coupling to the quad patches from the drive patch. Although more expensive, a material with a dielectric constant of 2.2 was chosen to achieve minimum electrical thickness. The final layer's thickness was chosen to optimize bandwidth while maintaining adequate coupling. Again, a low relative dielectric constant of 2.94 was chosen for this layer to maintain a wide bandwidth.

6.7.1.2 Results

The simulator used for throughout this work was *Ensemble* by Ansoft. This is a method of moments based code capable of multilayer mixed-dielectric simulation of microstrip antennas and circuits.

The frequency response for this antenna is shown in Figure 27, the input impedance is plotted on a Smith chart in Figure 28. It should be noted that mismatch loss was intentionally omitted from the gain calculation to better reveal the operating range of

the antenna. As seen in Figure 27, the return loss bandwidth exceeds the gain bandwidth, ensuring that the bandwidth of the structure is not being limited by the matching network.

Figure 27: The frequency response of the three layer slot-coupled linearly-polarized antenna ($|S_{11}|$ and Gain)

Figure 28: The input impedance of the three-layer slot-coupled linearly-polarized antenna

The impedance match is acceptable (VSWR 2:1) from 16.5 to 23.5 GHz. The spiral impedance locus is typical for all of the corporate antenna structures. The multiple resonances become a spiral when matched using the slot and stub as matching elements. Careful design allows for the use of a very simple matching network. In essence, the structure is optimized with itself as part of the matching network.

The radiation pattern stays relatively stable across this range as well, as can be seen from Figure 29, Figure 30 and Figure 31.

Figure 29: Radiation pattern at the centre frequency for the three-layer slot-coupled linearly-polarized antenna

At the centre frequency, a simulated gain of 12.7 dB and a 3 dB beamwidth of 43 degrees and 45 degrees for the E and H planes respectively is observed. The front-to-back ratio of 20 dB could be improved by blocking the back radiation of the slots with a metalized enclosure.

Figure 30: Radiation pattern at the lower band edge for the three-layer slot-coupled linearly-polarized antenna

At the lower band edge, a decrease in electrical aperture size and the corresponding decrease in gain is exhibited. The pattern symmetry is excellent, with 58 degree 3 dB beamwidths in both planes. At frequencies much below 17.5 GHz, the antenna is no longer near resonance, and performance degrades rapidly.

Figure 31: Radiation pattern at the upper band edge for the three-layer slot-coupled linearly-polarized antenna

At the upper band edge, the phase error between the patches increases, leading to large sidelobes in the E-plane and shoulders in the H-plane. These divert energy from the main beam, thereby reducing the gain. The E and H-plane 3 dB beamwidths are 31 degrees and 34 degrees respectively.

Throughout this work, simulated structures were built to verify simulated results. This allowed for fine tuning of the model, such as including glue sheet thickness, and ensured the relevance of future simulations.

This antenna was constructed and measured, and the frequency response is shown below, in Figure 32. The measured radiation pattern appears in Figure 33 and Figure 34.

Measured $|S_{11}|$ for the Three-Layer Slot-Coupled Linearly-Polarized Antenna

Figure 32: The measured frequency response for the three-layer slot-coupled linearly-polarized antenna

Although slightly mismatched at a few frequencies, these results confirm simulations. The mismatches were due to the omission of the glue used for substrate lamination in the model. The bandwidth is found to be between 16.2 GHz and 23.8 GHz. The 38% return loss bandwidth achieved here agrees well with the simulated 10 dB return

loss bandwidth of 34%. The centre frequency of the measured result, 20 GHz, was shifted down from the simulated centre frequency of 20.5 GHz. This is typical for Ensemble™, the moment-method based simulation package used in this work. To minimize design iterations, one typically tries to design for a simulated centre frequency a few percent above the target frequency.

Figure 33: Measured E-plane radiation pattern at the centre frequency for the three-layer slot-coupled linearly-polarized antenna

Figure 34: Measured H-plane radiation pattern at the centre frequency for the three-layer slot-coupled linearly-polarized antenna

The radiation patterns shown 11.5 dB gain and 42 degree beamwidths for both E and H planes. The additional length of feed line used was measured to have 1.0 dB of loss, making the element gain 12.5 dB. This agrees well with the simulated gain of 12.7 dB, as well as the 43 and 45 degree simulated beamwidths. Cross-polarization levels were approximately 35 dB down at boresight. Simulations do not reliably predict cross-polarization levels, so no comparison can be made.

Although the bandwidth achieved was large, even larger bandwidths were achieved in later iterations of this design. For this iteration, the limiting factor for lower band edge was the weak coupling to the array elements on the top layer. The limiting factor at the upper band edge was found to be excessive out of phase coupling to the centre patch on the radiating layer.

Figure 35 shows the current distribution of the patches on the upper layer at the centre frequency. This figure and others like it show the magnitude and phase of the current at each cell of the discretized structure. The magnitude is proportional to the length of the arrow, and the rotation of the arrow corresponds to phase. An arrow pointing to the right would be at 0 degrees, up would be 90 degrees and so on.

Figure 35: Current distribution of the three-layer slot-coupled linearly-polarized antenna at the centre frequency

Note the similar phase (rotation of arrows) and amplitude characteristics (length of arrows) of the currents on the outer four patches in contrast to the relatively weak currents on the centre patch. Similar phase is a clear indicator of broadside radiation.

At the upper band edge, there are far larger currents coupled to the centre patch and there is a large phase disparity between the centre and outer patches, as can be seen in Figure 36. This leads to poor broadside radiation, and a sudden change in the radiation resistance. These combined effects lead to a mismatched antenna with a poor radiation pattern.

Figure 36: Current distribution of the three-layer slot-coupled linearly-polarized antenna at the upper band edge

Clearly, suppression of the currents coupled to the centre patch would extend the bandwidth of this antenna. Eliminating the patch is not an option, as it suppresses the out-of-phase radiation from the four patches on the layer below.

Suppression was achieved by decreasing the size of the centre patch slightly, thereby increasing the frequency at which it was coupled to beyond the bandwidth of interest. This structure is shown in Figure 37.

Figure 37: A top view of the reduced size centre patch three-layer slot-coupled linearly-polarized antenna

The improved frequency response is shown in Figure 38.

Figure 38: The frequency response of the three-layer slot-coupled linearly-polarized antenna with a smaller centre patch ($|S_{11}|$ and Gain)

The impedance bandwidth is now improved to 37% by moving the upper band edge from 24 GHz to 24.2 GHz. More importantly, the gain over the upper portion of the band is improved. At 24 GHz, the gain of the antenna is improved by 2.6 dB over the equal sized top patch structure described in Figure 25.

This idea was taken to its full extent by placing a multitude of small patches in the centre area of the antenna. This resulted in having very little current coupled to each small

patch, while still blocking the radiation from the layer below. This design is shown in Figure 39.

Figure 39: A top view of the three-layer slot-coupled linearly-polarized antenna whose centre patch was replaced by a multitude of small patches

The frequency response is shown in Figure 40.

Figure 40: The frequency response of the three-layer slot-coupled linearly-polarized antenna ($|S_{11}|$ and Gain) whose centre patch was replaced by a multitude of small patches

At 24 GHz, the gain is increased to 9.6 dB, an improvement of 1 dB, which is only a slight improvement over the single small centre patch structure. Replacing the centre patch by a multitude of small patches yielded the highest performance of all the three-layer antennas.

This antenna was built, and its measured frequency response is shown in Figure 41, with the S_{11} across the band shown in Figure 42.

Figure 41: The measured frequency response of the three-layer slot-coupled linearly-polarized antenna ($|S_{11}|$, Gain, and Cross-pol.) whose centre patch was replaced by a multitude of small patches

Measured $|S_{11}|$ for the Three-Layer Slot-Coupled Linearly-Polarized Antenna with the Centre Patch Replaced by a Multitude of Patches

Figure 42: The measured $|S_{11}|$ response of the three-layer slot-coupled linearly-polarized antenna whose centre patch was replaced by a multitude of small patches over a large band

This confirms the bandwidth enhancement of current suppression for the centre patch. A bandwidth from 16.2 GHz to 23.7 GHz, or 37% was achieved. Gain was relatively constant across the band, with a peak value of 11 dB, which translates into 12 dB, when the 1 dB of line loss is removed.

6.7.1.3 Performance Summary for the Three-layer Slot-Coupled Linearly-Polarized Antenna

Gain at 19 GHz	10 dB Return Loss Bandwidth	3 dB Gain Bandwidth	Cross- Polarization	Efficiency* at 19 GHz
12 dB	37%	30%	-30 dB	99%

*Defined relative to the directivity of an ideal identically sized array of microstrip radiators

Table 2: Measured Performance of the three-layer slot-coupled linearly-polarized antenna whose centre patch was replaced by a multitude of small patches

Efficiency was calculated by subtracting the measured gain from the directivity resulting from the integration of the radiation pattern of an identical lossless array of microstrip patches (same size and relative position). This is a measure of aperture efficiency.

The three-layer corporate structure provides moderate gain, a large bandwidth, excellent cross-polarization isolation and high efficiency. Phase control problems were circumvented by minimizing coupling to the problem patches. Simulations have been shown to reflect actual performance, with a shift of a few percent in resonant frequency.

As a final note, dual-band performance is also possible with this structure. Wide band performance was achieved by bringing resonances together, but pushing the resonances apart could be just as easily done, leading to moderate bandwidths in separate bands.

6.7.2 Four-layer Slot-Coupled Linearly-Polarized Antenna

This is the second antenna configuration that will be discussed in this thesis. It is included to demonstrate the difficulties in extending the corporate feeding scheme beyond three layers. The phase problems alluded to in section 6.7.1.2 only become more prominent as the number of layers is increased.

6.7.2.1 Configuration

Top and bottom views of this antenna are shown in Figure 43.

Figure 43: A four-layer slot-coupled linearly-polarized antenna

There are four layers, not counting the feed network layer. The first layer consists of the driven patch, there are four patches on the second layer, five patches on the third layer, and twelve patches on the radiating layer. The ordering of the layers, along with

substrate selection, is shown in Figure 44. Layer selection considerations are similar to the three-layer case.

Figure 44: Layer definition for the four-layer slot-coupled linearly-polarized antenna

6.7.2.2 Results

The frequency response of the antenna is shown in Figure 45.

"Corporate" Parasitically-Fed Four-Layer Slot-Coupled Linearly-Polarized Antenna **Figure 45: The frequency response of the four-layer slot-coupled linearly-polarized antenna**

The useful bandwidth of this structure is considerably smaller than the three-layer antenna described in section 6.7.1. This is due to the instability of the radiation pattern over frequency, which is due to fundamental phase distribution problems occurring at the radiating layer. These problems cannot be eliminated by simple tuning. If designed for bandwidth alone, gain can be sacrificed at the upper band edge to smooth out the gain response. This would lead to very large bandwidths (approaching 50%). The control of the radiation pattern would be poor, however, and the aperture efficiency would be dismal.

Even at the frequencies where gain is at a maximum, there is a significant phase error between the inner patches and the outer patches as can be seen in Figure 46.

Figure 46: Current distribution for the four-layer slot-coupled linearly-polarized antenna

This phase error means that the outer patches are contributing less to the gain and adversely affect the radiation pattern. The radiation patterns at various frequency points are shown in Figure 47 and Figure 48.

"Corporate" Parasitically Fed Four Layer Slot-Coupled Linearly Polarized Antenna

Figure 47: The radiation pattern of the four-layer slot-coupled linearly-polarized antenna at 18 and 22 GHz

"Corporate" Parasitically Fed Four Layer Slot-Coupled Linearly Polarized Antenna

Figure 48: The radiation pattern of the four-layer slot-coupled linearly-polarized antenna at 24 and 26.5 GHz

Although the radiation pattern is passable at 24 GHz, the lack of phase control means that large bandwidths are difficult to maintain, while achieving good aperture efficiency.

6.7.2.3 Performance Summary for the Four-layer Slot-Coupled Linearly-Polarized Antenna

Gain at 24 GHz	10 dB Return Loss Bandwidth	3 dB Gain Bandwidth	Cross- Polarization	Efficiency* at 24 GHz
13.7 dB	54%	14%	not measured	47%

*Defined relative to the directivity of an ideal identically sized array of microstrip radiators

Table 3: Simulated Performance of the Four-layer Slot-Coupled Linearly-Polarized Antenna

The investigation of this structure has revealed that the nature of the corporate feed structure is not only a simple patch to patch coupling phenomenon, but also a relationship between phase and the radial distance from the feed point. It is this radial distance dependency, discussed earlier in section 5.1.2, that leads to the phase control problems encountered in this structure.

6.8 Series Parasitically-Fed Designs

Of the two parasitically-fed array architectures, the series parasitically-fed architecture, as mentioned in section 5.2, is capable of providing the largest aperture. Therefore, assuming bandwidth requirements permit, large apertures are best realized by using the series elements.

The first antenna shown is the eight-patch antenna. This is the simplest structure that illustrates the series concept. The second structure is larger and demonstrates both E- and H-plane coupling. The final design, shown in section 6.8.3, is the largest and most complex antenna structure in this thesis, a 32-patch antenna. This demonstrates the limit of this technique.

6.8.1 Eight-Patch Slot-Coupled Linearly-Polarized Antennas

The simplest of the series antennas, the eight-patch subarray, demonstrates the fundamental principles of the series array architecture.

6.8.1.1 Configuration

A top view of this antenna are shown in Figure 49.

Figure 49: The eight-patch slot-coupled linearly-polarized antenna

There are two layers, not counting the feed network layer. The first layer consists of the driven patch and two other feed patches, with eight radiating patches on the second layer. The ordering of the layers along with substrate selection is shown in Figure 50. Layer selection was similar to that of the three-layer antennas.

Figure 50: Layer definition for the eight-patch slot-coupled linearly-polarized antenna

6.8.1.2 Results

The frequency response of this antenna is shown in Figure 51.

Figure 51: The frequency response of the eight-patch slot-coupled linearly-polarized antenna

The truly remarkable characteristic of this antenna is not its gain or bandwidth, but its aperture efficiency. Referenced to a lossless microstrip patch array with perfectly uniform and in-phase excitation, this antenna made of lossy material, achieves nearly 100% efficiency. As discussed in section 5.1.2, this is due to the low losses in the feeding mechanism.

While this is indicative of very high efficiency, it raises questions as to the suitability of the lossless microstrip patch array as a reference. A possible explanation for this figure of merit is that the patches on the parasitically-fed array might be slightly more

directive than a single microstrip patch. Equally likely is the understatement of losses in simulation.

As any centre-fed series array, the beam remains at boresight over the band of interest and begins to split near the band edges. The radiation pattern at the centre frequency is shown in Figure 52.

Figure 52: The radiation pattern of the eight-patch slot-coupled linearly-polarized antenna at the centre frequency

A simulated gain of 15 dB with beamwidths of 30 degrees and 24 degrees for the E and H planes respectively is observed.

The amount of coupling can be controlled by the amount of overlap between the driver and radiating patch elements. The phase can be controlled by moving the patches closer to or farther from the centre. This kind of control is not critical for a small array, such as the eight-patch array, but becomes essential when designing larger arrays, squinted beam arrays, or multiple beam arrays.

The current distribution is shown in Figure 53.

Figure 53: The current distribution of the eight-patch slot-coupled linearly-polarized antenna at the centre frequency

From the current distribution, it is observed that there is good phase and amplitude uniformity between the radiating patches on the upper layer.

6.8.1.3 Performance Summary for the Eight-Patch Slot-Coupled Linearly-Polarized Antenna

Gain at 22 GHz	3 dB Gain Bandwidth	Cross- Polarization	Efficiency* at 19 GHz
15 dB	14%	not measured	100%

*Defined relative to the directivity of an ideal identically sized array of microstrip radiators

6.8.1.4 Table 4: Simulated performance of the eight-patch slot-coupled linearly-polarized antenna

With the high efficiency, phase and amplitude control, and structural simplicity (fewer layers), this design is an indication of the potential of the series parasitically-fed architecture.

6.8.2 12-Patch Slot-Coupled Linearly-Polarized Antennas

This antenna is included to demonstrate E-plane coupling and H-plane coupling simultaneously. Expansion in both directions allows for a larger aperture to be fed from a single point. This leads to increased efficiency once these antennas are placed in a large array, as discussed at the end of section 2.2.

6.8.2.1 Configuration

A top view of this antenna is shown in Figure 54. The substrate selection is identical to that of the 8-patch array shown in Figure 50.

Drawing to scale

Figure 54: The 12-patch slot-coupled linearly-polarized antenna

6.8.2.2 Results

The frequency response of this antenna is shown in Figure 55.

Figure 55: The frequency response of the 12-patch slot-coupled linearly-polarized antenna

Simulations show a 10 dB return loss between 22.08 GHz and 23.89 GHz.

With the H-plane coupled eight-patch antenna as a baseline, the E-plane portions were added and optimized to get the most uniform current distribution possible. Alternatively, the E-plane and H-plane coupled sections could be designed individually and then combined, since there is very little impact on the current distribution from one branch to another.

The gain was higher in proportion to the increase in aperture area relative to the eight-patch array. The radiation pattern is shown in Figure 56.

Figure 56: The radiation pattern of the 12-patch slot-coupled linearly-polarized antenna at the centre frequency

At the centre frequency of the 12-patch array, 16.81 dB of gain with an E-plane beamwidth of 28.7 degrees and a H-plane beamwidth of 23.3 degrees is observed.

The phase distribution on the radiating patches is still relatively uniform. This is shown in Figure 57.

Figure 57: The current distribution of the 12-patch slot-coupled linearly-polarized antenna at the centre frequency

This antenna was built and measured to confirm the results of simulation. The frequency response appears in Figure 58, and the measured radiation pattern in Figure 59.

Measured Frequency Response for the 12 Patch Array

Figure 58: Measured frequency response for the 12-patch slot-coupled linearly-polarized antenna

It is observed that the antenna is matched between 21.25 and 23.2 GHz, with a 3 dB gain bandwidth spanning 21.08 and 22.30 GHz. This agrees well with simulations, except for a downward shift in the bands.

Figure 59: Measured radiation pattern for the 12-patch slot-coupled linearly-polarized antenna at the centre frequency

A gain of 15.25 dB, which translates into 16.25 dB when the line loss is taken into account, is observed. The E- and H-plane beamwidths are 26.6 degrees and 25.4 degrees respectively. Again, good agreement with simulated results is apparent.

6.8.2.3 Performance Summary for the 12-Patch Slot-Coupled Linearly-Polarized Antenna

Gain at 22 GHz	10 dB Return Loss Bandwidth	3 dB Gain Bandwidth	Cross- Polarization	Efficiency* at 24 GHz
16.25 dB	8.8%	5.6%	not measured	84%

* Defined relative to the directivity of an ideal identically sized array of microstrip radiators

Table 5: Measured Performance of the 12-Patch Slot-Coupled Linearly-Polarized Antenna

This design has established the viability of E-plane coupling for parasitically-fed arrays. The efficiency is still very high, but the slight drop relative to the 8-patch array is due to the amplitude taper for the E-plane coupled patches. For feed-network-fed arrays, the larger the array, the more difficult it is to maintain high efficiency. With series parasitically-fed arrays, moderately large arrays can be built with a relatively small degradation in efficiency.

6.8.3 32-Patch Slot-Coupled Linearly-Polarized Antennas

The 32-patch slot-coupled linearly-polarized antenna is included in this thesis to push the limits of this technique. There are several limits that can be encountered, such as bandwidth reduction, aperture efficiency reduction due to weaker coupling to the outer elements, and computational hardware and software limitations for large structures. The 32-patch parasitically-fed array came close to all three limits.

6.8.3.1 Configuration

A bottom view of this antenna is shown in Figure 60. The substrate selection is identical to that of the 8-patch array shown in Figure 50.

Figure 60: The 32-patch slot-coupled linearly-polarized antenna

Here extensive use of both E and H-plane coupling is observed. The structure directs energy along four separate “T” shaped paths. This design does not have all of its

radiators at integer multiples of 360 degrees from the centre, so some additional phase error is to be expected at those patches due to the radial effect mentioned in section 5.1.2.

6.8.3.2 Results

The frequency response of this antenna is shown in Figure 61.

Figure 61: The frequency response of the 32-patch slot-coupled linearly-polarized antenna

With a 5% bandwidth, the 32-patch array is narrowband, but this is adequate for many applications. The radiation pattern is directive in both planes as shown in Figure 62.

Figure 62: The radiation pattern of the 32-patch slot-coupled linearly-polarized antenna at the centre frequency

The H-plane beamwidth is simulated to be 15.1 degrees, and the E-plane beamwidth is simulated to be 19.6 degrees. The E-plane pattern has significant shoulders due to the phase error in the corresponding plane. This can be seen in the current distribution shown in Figure 63.

Figure 63: Current distribution of the 32-patch slot-coupled linearly-polarized antenna at the centre frequency

One can clearly see the phase error and amplitude taper in the E-plane coupled patches, and the attenuation of the current on outer patches from inner patches. These factors place the limit on aperture size at 32 patches. Any further expansion would lead to only a small increase in gain, and would degrade aperture efficiency dramatically.

This antenna was built and measured, with the results shown in Figure 64.

Figure 64: The measured frequency response of the 32-patch slot-coupled linearly-polarized antenna ($|S_{11}|$, Gain)

The frequency response was as simulated, with a downward shift in frequency observed in the measurements when compared to the simulations. The antenna was well matched from 21.18 GHz to 23.34 GHz with a slight mismatch at 22.20 GHz. A 3 dB gain bandwidth from 21.3 GHz to 22.1 GHz was observed.

Figure 65: The measured radiation pattern at the centre frequency for the 32-patch antenna

From the radiation pattern, one can see an 18.3-degree H-plane beamwidth, and 17.3-degrees for the E-plane. This differs significantly from simulations. As well, the measured gain was 16.6 dB which, after adding 1 dB to account for the feed line loss, was 17.6 dB. This fell significantly short of the 19.3 dB predicted by simulation.

This discrepancy must be due to the limitations of the modelling. As the structure gets larger, the error caused by differences between the model and the real antenna

become more significant. For example, an error in the dielectric constant would lead to a significantly different phase for the outer patches.

6.8.3.3 Performance Summary for the 32-Patch Slot-Coupled Linearly-Polarized Antenna

Despite the discrepancy between measurements and simulations, the objective of determining the aperture size limits was still met. Performance for the 32-patch array is summarized in Table 6.

Gain at 21.6 GHz	10 dB Return Loss Bandwidth	3 dB Gain Bandwidth	Cross- Polarization	Efficiency* at 21.6 GHz
17.6 dB	9.7%	3.7%	-33 dB	53%

*Defined relative to the directivity of an ideal identically sized array of microstrip radiators

Table 6: Measured Performance of the 32-Patch Slot-Coupled Linearly-Polarized Antenna

6.9 Array Techniques

As mentioned in section 6.8.3.2, the maximum number of patches that can be placed in a parasitically-fed array cannot realistically exceed 32, since the amount of coupling to the outer patches becomes too small to make them contribute efficiently to the main beam. Thus, for directivities exceeding 20 dB, arrays of parasitically-fed arrays must be used.

6.9.1 Configuration

To design a printed array, important design choices need to be made. First, an appropriate array element, in this case a subarray, has to be chosen. Second, an array architecture has to be chosen. The term “array architecture” includes the way that the elements are positioned and the method by which the electromagnetic power is brought to them (feed network).

6.9.1.1 Array Element

The key attribute of an antenna when considering how it will perform in an array environment is its aperture efficiency. In other words, the array element has to contribute significantly to the main beam directivity for the amount of area that it occupies. If the aperture efficiency is too low, what is effectively realized is a sparse array. The sparse

array leads to large sidelobes commonly known as grating lobes. These degrade both the radiation pattern and the peak directivity.

Aperture efficiency will always be high for small structures like patch or slot antennas, and is therefore rarely considered in a design. For subarrays like those considered here, a high aperture efficiency for an element can no longer be assumed.

Given that the element has high aperture efficiency, we want to maximize its size, thereby minimizing the number of feed points, hence reducing the feed network loss and complexity.

After having considered the antenna designs in sections 6.7 and 6.8, an ideal candidate was not found. The corporate elements all had fairly small apertures and the eight and twelve-patch series elements were not much larger. The 32-patch array would have been an excellent candidate, but the aperture efficiency was too low. A design with more than 12 patches and less than 32 patches was required. As well, given that H-plane coupling is stronger than E-plane coupling, the array could be made larger in the H-plane direction than in the E-plane direction. The resulting design is shown Figure 66. Its performance along with the array performance will be discussed in section 6.9.3.1.2.

Figure 66: A 24-patch subarray well suited for arrays

It should be noted that this element was designed to operate at 28 GHz, to verify the suitability of parasitically-fed arrays for LMCS/LMDS. The substrate layers were 10 mil Duroid 5880 ($\epsilon_r=2.2$) for the feed layer, and low cost 8 mil 4003 material for the two patch layers. The low cost material was used to verify that the patch layers could tolerate higher loss material without degrading performance significantly, due to the parasitic feeding mechanism.

6.9.1.2 Array Architecture

There are three principal types of array architectures for printed arrays; series, corporate and corporate/series, as illustrated in Figure 67.

Figure 67: Series, Corporate and Corporate/Series Arrays

The series configuration minimizes the line loss to each element, but suffers from bandwidth limitations because the farthest element goes through several phase rotations. The corporate configuration has a broad bandwidth, as the elements are always in phase, but the line loss is higher due to the longer path length. For the corporate/series arrays,

one generally uses as much series feeding as the target bandwidth will allow, then use a corporate feed to complete the feed network. The corporate/series approach is preferred because it maximizes efficiency for a given bandwidth.

For arrays of electrically large subarrays, like the 24-patch subarrays considered here, the use of any series feed technique is not feasible. The phase at each feed point changes at a very rapid rate, with even the second element's phase ($\angle S_{31}$) changing too rapidly to achieve bandwidths useful for LMCS/LMDS. This rapid phase change is due to the electrically long separation between elements. A series array phase response on Rogers 3003 ($\epsilon_r=3.3$) substrate is shown below in Figure 68.

Figure 68: 3-Element series array phase response at each element

The narrowband performance of series feed network elements makes the choice of feed networks for arrays of parasitically-fed subarrays evident: corporate.

The remaining decisions to be made to complete the array architecture were the number and placement of subarrays. The number of subarrays required is determined by the target directivity. For LMCS applications, directivities from 20 to 40 dB are required, depending on the system being considered. For reflector-based systems, user terminal antenna directivities of 36 dB are typical. Since one of the objectives of this work was to

determine how planar arrays could compete with reflectors, given the new parasitically-fed techniques presented here, an aggressive target of 36 dB was set to allow a direct comparison to be made.

To achieve 36 dB directivity, approximately 64 24-patch subarrays were required (1536 patches). Having the number of elements be a base 2 number simplifies the corporate feed network to a simple binary tree. To achieve maximum efficiency, a uniform aperture distribution was desired, as this also simplified the feed network design. The subarray spacing was optimized to bring grating lobes below -15 dB. Measured and simulated array patterns will be shown in section 6.9.3.2. The resulting array design is shown in Figure 69.

Figure 69: 1536-patch array consisting of 64 subarrays

6.9.2 Dual-Polarized / Dual-Frequency Arrays

As mentioned in section 6.4, dual polarization with high isolation can be achieved with a three slot technique. Large arrays of subarrays with polarization diversity and frequency diversity are desirable, but the three feed points make the realization of a feed network more complex.

6.9.2.1 *Dual-Polarized / Dual-Frequency Element*

The same considerations for element selection apply here as for section 6.9.1.1. We want a large element with high aperture efficiency. Taking the techniques of section 6.4, and applying them to the element of section 6.9.1.1, the dual-polarized / dual-frequency subarray shown in Figure 70 can be designed.

Figure 70: Dual-polarized / dual-frequency subarray

The resulting performance will be presented and discussed in section 6.9.3.1.

6.9.2.2 Dual-Polarized / Dual-Frequency Array Architectures

The primary difference and difficulty associated with the realization of an array of three-feed-point elements (two for one polarization and frequency, and one for the other polarization and frequency) is the routing of the feed network. As discussed in section 6.9.1.2, corporate feed structures must be used. This makes the collocation of feed networks impossible on a single layer, as a bridging technique of some kind has to be used. This is illustrated in Figure 71, with the required bridges shown as dashed lines.

Figure 71: Collocation of Feed Networks

There are several ways in which the bridging could be accomplished, such as flying leads, adding another dielectric layer to the feed network on which the bridges will

be realized, or adding an entire stripline layer. These methods all have serious shortcomings and are shown in Figure 72.

Figure 72: Bridging techniques

Flying leads are not easily mass produced, mechanically weak, and will not provide much isolation between the two lines being bridged. Although more mechanically sound, the addition of a single dielectric layer will provide no more isolation than the flying lead. Adding the additional ground plane required for a stripline layer means having more than one ground plane, which leads to ground plane continuity problems.

These solutions all require increases in cost and complexity, which are highly undesirable for large volume production. As well, the high isolation between the two polarizations at the array element level is eliminated by the poor isolation in the feed network. Ideally, one would like to be able to create the bridges without additional layers

or processing complexity, while maintaining a high degree of isolation between the feed networks.

A solution does exist. By slot coupling from a line on the feed network layer to a bridge line on the first patch layer, and then back down to a line on the feed network layer, a bridge with high isolation can be created. This method does not require the addition of any new substrate layers, simply more efficient use of the existing ones. This technique is illustrated in Figure 73.

Figure 73: Bridging using slot coupling

The only limitation with this technique is that you can not simply insert a bridge anywhere on the feed network layer, since the bridge trace could interfere or even intersect with a patch. In fact, these bridges should only be run between elements, as shown in Figure 71. Furthermore, to maximize isolation, the bridge lines should be run parallel to the polarization of the patches that they are feeding. This ensures orthogonality with the patches of the opposite sense, thereby minimizing coupling and maximizing isolation. This is illustrated on Figure 71.

With these techniques, it should be possible to make large arrays of dual-polarization / dual-frequency subarrays, while maintaining high isolation.

6.9.3 Results

6.9.3.1 24-Patch Subarray Performance

All the results for both the single-polarization (Figure 66) and dual-polarization / dual-frequency (Figure 70) elements come from the same physical antenna. The results for the 28 GHz vertical polarization are identical with or without the additional 20 GHz horizontal polarization. This claim can be made because of the isolation between the two ports. Isolation between the two ports exceeded 40 dB across the band, as shown in Figure 74.

Figure 74: Isolation between the 28 GHz vertically-polarized port and 20 GHz horizontally-polarized port

6.9.3.1.1 20 GHz Horizontally Polarized Port

The return loss at this port was measured and the results are shown in Figure 75.

Figure 75: Return loss at the 20 GHz horizontally polarized port

One sees that the subarray is indeed reasonably well matched between 20.3 and 21.4 GHz. This corresponds to an impedance bandwidth of 5.3%. The impedance is quite sensitive to the phase at the two slots. The 180° phase relationship can only be maintained over a narrow bandwidth with the simple delay line technique used here (see Figure 70). A more broadband phase delay scheme would have led to a broader bandwidth.

The radiation pattern of the subarray at this port was also measured and can be seen in Figure 76, Figure 77 and Figure 78.

Figure 76: 3-D representation of the radiation pattern of subarray when connected to the 20 GHz port

Figure 77: Contour plot of the radiation pattern of sub array when connected to the 20 GHz port

Figure 78: Principal plane pattern cuts for the 20 GHz horizontal polarization

The peak gain was measured to be 16.14 dB at 20.5 GHz and thanks to the near-field measurement system used to make these measurements, data for all planes exist and when integrated yields a directivity of 17.33 dB. The small mismatch can account for 0.1 dB and the line loss for another 0.9 dB, implying a corrected gain of 17.14 dB.

6.9.3.1.1.1 Performance Summary for the 20 GHz Horizontally Polarized Mode

Gain at 20.5 GHz	10 dB Return Loss Bandwidth	3 dB Gain Bandwidth	Cross- Polarization	Efficiency at 24 GHz (Directivity Method)
17.1 dB	5.3%	~5%	-20 dB	95%

Table 7: Measured performance of the 24-patch dual-polarized / dual-frequency subarray at the 20 GHz port

6.9.3.1.2 28 GHz Vertically Polarized Port

Results in this section describe not only the performance of the array element shown in Figure 66, but also the performance of the 28 GHz vertically polarized port of the dual-polarization / dual-frequency antenna, shown in Figure 70. As discussed in section 6.9.3.1, the results are interchangeable because of the high isolation between the horizontally and vertically polarized ports. The return loss of the 28 GHz port is shown in Figure 79.

Figure 79: Return Loss of the 28 GHz vertically polarized port

One notes a 10 dB return loss bandwidth from 26.28 to 29.16 GHz, which translates into 2.88 GHz or 10.3%. This bandwidth is slightly overstated due to the feed line losses attenuating the return.

Radiation patterns at 27 GHz for this 24-patch subarray are shown in Figure 80, Figure 81, Figure 82, and Figure 83.

Figure 80: 3-D representation of the radiation pattern of subarray when connected to the 28 GHz port

Figure 81: Contour plot of the radiation pattern of sub array when connected to the 28 GHz port

In both Figure 80 and Figure 81, one sees the complete radiation pattern over both 120 degree ranges. The sidelobes at ± 46 degrees azimuth and ± 38 degrees elevation were 17.4 dB down from the 15.40 dB peak at boresight. The feed line loss was 1.8 dB (see Figure 85) making the corrected gain 17.2 dB. The loss due to mismatch was very small and therefore not included in the corrections.

Figure 82: Principal plane pattern cuts for the 28 GHz vertical polarization

The E-plane -3 dB beamwidth is 24.5 degrees and the H-plane beamwidth is 18 degrees. The directivity was found to be 18.89 dB. Cross-polarization levels in the principal planes were very low.

Figure 83: Contour plot of the cross-polarized radiation pattern of sub array when connected to the 28 GHz port

In Figure 83, one can see that the cross-polarization performance is significantly poorer between the principal planes. The cross-polarization peaks at ± 36 degrees azimuth and ± 22 degrees elevation are 23 dB down from the 15.40 dB peak at boresight.

6.9.3.1.2.1 Performance Summary for the 28 GHz Horizontally Polarized Mode

Gain at 27 GHz	10 dB Return Loss Bandwidth	3 dB Gain Bandwidth	Cross- Polarization	Efficiency at 27 GHz (Directivity Method)
17.2 dB	10.3%	~8%	-23 dB	68%

Table 8: Measured performance of the 24-patch dual-polarization / dual-frequency subarray at the 28 GHz port

The relatively low efficiency of 68% is partially due to the low cost Rogers 4003 substrate material used on the patch layers. Although the effects of lossy material are reduced by parasitically feeding an array, some loss is still to be expected.

The performance of each of the two integrated antennas is excellent individually. But the fact that the two can be collocated with such high isolation between them is truly remarkable.

6.9.3.2 1536-Patch Array Performance

The final antenna to be discussed is the largest. In an effort to compare the performance of parabolic reflectors and planar arrays using parasitically-fed subarrays, a planar antenna of comparable directivity was built. As discussed in section 6.9.2.2, 64 24-patch subarrays were fed with a corporate feed as shown in Figure 69.

The same material stack-up was used here as for the 24 patch subarray. Low cost material was used wherever possible to try to keep the cost of the planar array comparable to that of a reflector. The feed network substrate used Duroid 5880 in an effort to minimize the losses in the feed network. More about the performance of Duroid 5880 at 28 GHz will be said later.

The return loss of the 1536-patch array was measured and is shown in Figure 84.

Figure 84: Return loss of 1536-patch array

The return loss response shows a broadband match, which was best around 28 GHz. Although the graph seems to reflect the return loss of the 24-patch subarray, this would be a partially erroneous conclusion. The return loss was again measured, but this time with a brass plate covering the entire array face, preventing any kind of radiation. There was little difference between the return loss responses of the two cases, especially around 28 GHz.

This indicates that rather than an attenuated version of the return loss of the element, this is actually a measurement of the line loss response combined with the return loss of the element, and the tee power dividers, masked by the return loss of the connector.

A line loss test was performed with a 9.5 cm long, 50 ohm line on the same substrate as the feed network, namely 0.010" Duroid 5880.

Figure 85: Duroid 5880 loss over frequency for a 9.5 cm long line

These results seem to clarify the return loss response shown in Figure 84. A line loss of 0.37 dB/cm at 28 GHz was observed, with an apparent material resonance around 28 GHz. This would have attenuated the return from the brass plate covered elements to the point where the difference between the elements being a matched load or not, could not be observed. It would seem that Duroid 5880 is not the ideal material to use at 28 GHz.

The radiation pattern measurements for the 28 GHz array, as with the dual-polarization / dual-frequency 24-patch array, were made on the planar near field system. Results at 27.5 GHz are shown below in Figure 86, Figure 87, Figure 89, Figure 90 and Figure 91.

Figure 86: 3-D representation of the radiation pattern of the 1536-patch array

Figure 87: Contour plot of the radiation pattern of the 1536-patch array

In both Figure 86 and Figure 87, one sees the complete radiation pattern over both 100 degree ranges. The largest sidelobes are in the principal planes at ± 28 degrees azimuth and ± 18 degrees elevation are 14 dB down from the 27.8 dB peak gain at boresight. These are the grating lobes referred to in section 6.9.1.2 and are predicted by

simulation. Simulated array patterns based on simulated subarray radiation patterns are shown in Figure 88.

Figure 88: Simulated array radiation patterns

Sidelobes at ± 35 azimuth and ± 38 elevation were also present. This is to be expected, as sidelobes are present in the subarray radiation pattern at this angle as well (see Figure 81).

Figure 89: Principal plane pattern cuts for the 1536-patch array

Figure 90: Principal plane pattern cuts for the 1536-patch array near boresight

From the principal plane pattern cuts, one can determine that the H-plane and E-plane -3 dB beamwidths are 2.15 and 3.37 degrees respectively. The directivity of the array was calculated to 36.3 dB.

Figure 91: Contour plot of the cross-polarized radiation pattern of the 1536-patch array

Cross-polarized radiation was at a maximum at the four peaks at ± 17 degrees elevation and ± 28 degrees azimuth. The level at these peaks was 2.8 dB, 25 dB down from the maximum gain.

The 28 GHz measurements were made with a planar near-field measurement system. When making measurements of large arrays, planar near-field measurement systems offer the additional advantage of being able to measure the aperture distribution of a planar array. This is very useful as a debugging tool. Any manufacturing problem, such as misalignment, delamination or deformation is rapidly detected and the problem array elements can be singled out.

The aperture distributions for the 1536-patch array at two frequencies are shown in Figure 92, and Figure 93.

Figure 92: Measured aperture distribution of the 1536-patch array at 27.5 GHz

In Figure 92, one sees a uniform aperture distribution in both amplitude and phase. This was expected for a corporately-fed array.

Figure 93: Measured aperture distribution of the 1536-patch array at 28.5 GHz

Approaching the band edges of the sub array, one sees the weaker coupling to the outer patches of the subarrays as well as the phase ripple of the aperture associated with band edge performance. The incongruencies in the aperture distribution lead to degradation in the radiation pattern and are manifested primarily by increased grating lobes.

6.9.3.2.1.1 Performance Summary for the 1536-Patch Array

Gain at 27.5 GHz	Sidelobe Levels	Pattern Bandwidth (-10 dB sidelobes)	Cross- Polarization	Efficiency at 27.5 GHz (Directivity Method)
27.8 dB	-14 dB	~4.5%	-25 dB	7.1%

Table 9: Measured performance of the 1536-Patch Array

The construction of this array has really underscored the challenge facing the millimeter-wave flat plate array designer. Eliminating the feed network is the key to increasing the efficiency of large arrays. Having eliminated the portion of the feed network required to feed the final 24 patches, important gains in efficiency were realized.

Considering the high losses of the material at and around 28 GHz, the achieved efficiency is remarkable. If the line loss had increased linearly between 25 and 35 GHz, instead of increasing sharply at 28 GHz, an additional 4.7 dB of gain would have been measured. With the resulting gain of 32.5 dB, an efficiency of 41.7 % would be achieved. Such efficiency begins to rival that of the parabolic reflector.

Sidelobes were slightly under the design goal of -13 dB for a uniform aperture distribution. Cross-polarization levels were low even at their worst-case angles and extremely low in the principal planes. Bandwidth was adequate at 4.5%.

When considering the performance of this antenna as an active array, one begins to see the full potential for this kind of array technology. Placing a 1/4 Watt amplifier at each of the 64 array elements would translate into an EIRP (effective isotropic radiated power) of 46 dBW.

7. Summary and Future Work

The primary objective of this thesis was to make important advancements in low-profile antenna arrays to improve their efficiency. This was accomplished by eliminating as much of the feed network as possible. For directivities exceeding ~ 20 dB, however, hybrid arrays involving some use of feed networks is required.

Broadband, multiple resonance techniques were demonstrated and their limitations were explored. Capable of achieving gains of 12 dB and bandwidths exceeding 37%, these corporately-fed antennas demonstrated unique characteristics that could make them well suited for broadband or multi-band systems. The lack of phase control of the radiating elements seems to be the limiting factor in realizing the next increase in aperture size while maintaining high aperture efficiency. Further development of these antennas using mixed dielectric material on a given layer for phase control could prove to be the breakthrough required to larger antennas with even larger bandwidths.

The series-fed subarrays were shown to be capable of realizing large apertures while still achieving high aperture efficiencies. In addition, the reduction in the number of required layers lead to a simpler and less expensive structure. The bandwidth of these arrays was sufficient for use in emerging fixed terminal applications. The limits for aperture sizes were established and little room for dramatic increases in directivity exists.

The series configuration provided the opportunity to apply novel techniques, the most notable of which is the ability to provide orthogonal polarizations in different frequency bands from the same aperture with very high isolation.

The bulk of the future work in this area lies in the integration of these subarrays in large arrays. A number of aspects require further investigation. The first task would be to evaluate materials at millimeter-wave frequencies to find substrates with better performance than shown in Figure 85. Implementation of a dual-polarization / frequency array using the techniques discussed in section 6.9.2.2 should be investigated.

For loss reduction, an option would be to replace the microstrip feed network with a waveguide network. The key here would be to develop a waveguide to microstrip patch transition. Another option, using active array techniques, requires the development of low cost and simple packaging techniques.

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IMAGE EVALUATION TEST TARGET (QA-3)

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Rochester, NY 14609 USA
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Fax: 716/288-5989

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