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High-Efficiency Passive and Active Phased Arrays and Array Feeds for Satellite Communications

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High-Efficiency Passive and Active Phased Arrays and Array Feeds
for Satellite Communications

Zhenchao Yang

A dissertation submitted to the faculty of
Brigham Young University
in partial fulfillment of the requirements for the degree of
Doctor of Philosophy

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ABSTRACT

High-Efficiency Passive and Active Phased Arrays and Array Feeds for Satellite Communications

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Doctor of Philosophy

Satellite communication (Satcom) services are used worldwide for voice, data, and video links due to various appealing features. Parabolic reflector antennas are typically used to serve a cost effective scheme for commercial applications. However, mount degradation, roof sag, and orbital decay motivate the need for beam steering. Limited scan range beam steering opens a third option for electronic beam steering with lower cost than full aperture phased arrays and higher tracking speed and accuracy than mechanical-only steering.

Multiple high efficiency passive patch array feeds were designed, fabricated, and measured, including a $2 \times 2$ MSA array, a stacked shorted annular patch antenna, and an SIW-fed hexagonal array feed based on PTFE material, achieving performance comparable to a horn feed. For multiband dual polarization applications, passive MSA feed solutions are also provided. Multiple MSA array feeds with high isolation were designed for dual band dual polarization applications. More functionality can be realized with multi-layer PCB techniques for complex communication scenarios.

Limited scan range electronic beam-steering with a parabolic reflector fed by an active array feed which only needs gain control was demonstrated experimentally, leading to a low cost and effective solution for active beam scanning. A cost-effective flat-panel phased array with limited scan range electronic beam-steering was proposed by tiling high efficiency $4 \times 4$ passive subarrays and performing beam scanning at the tile level. The sidelobe issue was also investigated to comply with the pattern mask requirement set by FCC.

To enable better use of circularly polarized (CP) MSAs for electronically beam-formed antenna systems, the impact of mutual coupling on the performance of high-sensitivity dual-polarized receivers for satellite communications applications was analyzed. A new analysis method for intrinsically dual-CP MSAs based on an equivalent circuit model and Jones matrices was proposed and validated to overcome the port isolation challenge. The model provides accurate estimates of impedances and S-parameters, as well as field parameters such as axial ratio. The feasible region for XPI and impedance mismatch factor is found for dual CP antennas. The circuit model enables multiple useful applications. Effective decoupling and matching schemes were proposed and demonstrated, leading to a high isolation, good match, and wide AR bandwidth dual CP MSA for satellite communications.

Keywords: Passive and Active Array Feed, Microstrip Array, Multiband, Dual Polarization, Circular Polarization, Limited Scan Range Beam Steering, Satellite Communications
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CHAPTER 1. INTRODUCTION

Satellite communication (Satcom) services are used worldwide for voice, data, and video links due to various appealing features. One satellite can create a wide footprint, which theoretically enables global coverage except for pole areas by using three satellites. Another advantage is that satellites as central hubs directly establish links with end-users, solving the last mile problem fundamentally. Satellite communications are appealing for rural, remote, and ocean areas and high altitude air, where other telecommunication infrastructures are expensive or impossible to build.

There are three types of satellite orbits, low earth orbit (LEO), medium earth orbit (MEO), and geostationary orbit (GEO). A satellite on LEO or MEO is always moving relative to the earth, which requires the corresponding receiver with either wide beam pattern to cover most part of the hemisphere or the capability of beam steering widely to track the satellite. One serious drawback for these two orbits is dozens of satellites have to operate on orbits coordinately and synchronically in order to provide a consistent service for global coverage. This means a large amount of initial investment is needed to build and launch satellites before making profit from providing services. Iridium plan is one perfect example for illustrating how risky such ambitious scheme would be.

On the other hand, DirecTV, Dish Network, and Hughes Network System demonstrate a more profitable, feasible path to implement satellite communications - to use GEO satellites. Two main services are provided by them, direct broadcast service (DBS) and fixed satellite service (FSS). Thanks to fixed satellite position in the sky, a terrestrial transceiver only needs a fixed high gain antenna to communicate with a GEO satellite. Parabolic reflector antennas are typically used to serve a cost effective scheme for commercial applications.

However, mount degradation, roof sag, orbital decay, and in-motion terminals like luxury cars, ships and airplanes motivate the need for beam steering. A full aperture-type electronically steerable phased array panel can fulfill the task perfectly, but is too costly for consumer use [4].
Mechanical steering based on inertial and GPS systems is still the only feasible method to date in commercial applications in spite of bulky size and heavy weight.

1.1 Limited Scan Range Electronic Beam Steering Systems and Passive Array Feeds

Limited scan range beam steering fills the technique and application gaps between full aperture phased arrays and mechanical steered dishes. From a technical point of view, precise beam steering in a limited range plus coarse mechanical steering opens a third option for electronic beam steering with lower cost and higher tracking speed and accuracy than mechanical-only steering, particularly for multi-feed systems. There are also three applications that conventional fixed-mount dishes cannot serve. First, current dish installation demands accurate pointing. Lowering the accuracy requirement would increase the productivity of dish installation. Second, mount degradations caused by sagging roof and weather disasters require dish re-pointing, which costs millions of dollars in the Satcom industry. Automatic self-adjustment on beam direction avoids huge labor cost for trivial cases. Last but not least, a fuel-depleted satellite becomes inclined on its orbit, forming a constant ‘8’ shape movement in the sky. Spectrum use on this kind satellite is cheap due to the unstable connection, but the terrestrial stations have to track the movement by beam steering. Limited scan range beam steering is a cost-effective solution for these three applications.

This dissertation considers two research paths to achieve limited scan range electronic beam steering: (1) active array feeds on a parabolic reflector; (2) actively driving multiple passive sub-array tiles of an array panel. The progresses on these two areas are shown in Chapters 4 and 5, respectively.

The first fundamental question which must be answered before building an array is to choose the right antenna type for the array elements. Microstrip antennas (MSA) possess many appealing features like low profile, low manufacturing cost and easy integrateability with circuit components [5], so they could be a promising replacement for the traditional horn feeds and candidate for the element of a subarray tile. Meanwhile, the sensitivity of the receiving antenna system is strongly influenced by radiation efficiency, because ohmic losses feed can increase noise and attenuate the signal. The poor radiation efficiencies of the microstrip antenna [6, 7] must be addressed in order to be competitive as a feed or tiled array for satellite communication applications.
To be a useful feed antenna for a reflector, its feed pattern shape also needs to be designed carefully to satisfy the aperture and spillover efficiency requirement. Chapter 2 deals with these topics.

A simple single band single polarization feed is not able to satisfy diverse needs. Dual band operation is commonly required for two different scenarios. One is transmitting/receiving (Tx/Rx) two-way communication in a designated radio band with small frequency separation generally. The isolation between bands is required to be very high, at least 40 dB in the Tx band, so that the high transmitting power level would not affect the receiver performance. The other is to receive signals at two different radio bands from the same satellite. Often dual polarization is exploited to increase communication channels or the data rate, either in dual linear or dual circular polarizations (LP/CP), depending on the band and application. Dual CP receivers need to be designed carefully, because high isolation and good pattern shape/cross polarization level can only be maintained in much narrower bandwidth, compared with linear polarization [8]. Chapter 3 shows planar array feed solutions for multi-functional scenarios.

Early systems used linear polarization (LP) due to the simplicity and low cost of feed antennas. For more recently introduced satellite communication (Satcom) services, circular polarization (CP) dominates because it avoids the need for polarization alignment of terrestrial feeds. Coupling between orthogonal linear polarized ports is typically very low (-30 dB or less). To realize dual CP with high isolation, a quadrature hybrid is needed to transform LP to CP, but the hybrid is lossy and undesirable for satellite communications applications where efficiency is paramount. There are surprisingly few hybrid-free, intrinsically dual CP microstrip antenna (MSA) designs in the literature. One reason for this is that it is challenging to obtain high isolation between ports while maintaining low return loss and good polarization quality. Chapter 6 answers the impact of coupling on the sensitivity of dual polarized receivers and illustrates a new analysis method for compact and efficient dual CP MSA design.

1.2 Contributions

My research contributes to the antenna and propagation community can be summarized as delivering low cost high efficiency passive and active array feeds and planar arrays and enabling high performance dual CP MSAs for satellite communications. The contribution can be divided into three groups as follows.
High efficiency passive array feeds

- Careful comparison of numerical models, measurements on dish, and range measurement of radiation loss and efficiency to obtain best benchmarked results in this passive array feed application.

- Demonstrated the highest measured radiation efficiency 93% ever reported for a 2x2 microstrip array feed with an FR-4 compatible fabrication process.

- Designed a high efficiency stacked shorted annular patch antenna that is much smaller than a commercial horn feed but competitive in measured SNR performance.

- Designed a high efficiency SIW-fed hexagonal planar array which has the best simulated G/T performance reported to date among non-horn type feeds and can match the performance of a commercial horn feed.

- Designed a 2x2 MSA array feed with the highest reported isolation between Tx/Rx ports for VSAT terminals that require low Tx signal bleeding into the Rx signal path.

- Provided a high efficiency terrestrial MSA array feed solution for communications with two satellites operating in different bands collocated at the same orbital slot, and verified high efficiency of a highly complex antenna design with multi-layer PCB fabrication techniques.

Low cost electronic beam-steering solutions with a limited scan range

- Demonstrated limited scan range electronic beam-steering with a parabolic reflector fed by an active array which only needs gain control, leading to a low cost and effective solution for active beam scanning.

- Proposed a cost-effective flat-panel phased array with limited scan range electronic beam-steering by tiling high efficiency 4x4 passive subarrays and performing beam scanning at the tile level.
High efficiency intrinsically dual CP MSA

- Provided a lower bound on sensitivity degradation due to mutual coupling in a dual polarized receiver, which shows if the LNAs are designed optimally, coupling can be -10 dB with minimal impact on sensitivity.

- Established a guideline for intrinsically dual CP MSA design based on an equivalent circuit model and Jones matrix analysis, and determined the feasible region for polarization quality and impedance mismatch factor.

- Proposed and demonstrated in simulation an effective decoupling scheme to overcome the port isolation issue for intrinsically dual CP MSAs, leading to high isolation, good impedance matching, and wide AR bandwidth as required for dual CP satellite communications.
CHAPTER 2. HIGH EFFICIENCY PASSIVE ARRAY FEEDS

2.1 Figures of Merit for a Terrestrial Antenna in Satellite Communication Systems

One key figure of merit for a wireless communication system is signal-to-noise ratio (SNR) from the receiver output, which is defined as

$$\text{SNR} = \frac{\lambda^2 S^{\text{inc}} G}{4\pi k_B B T_{\text{sys}}},$$  \hspace{1cm} (2.1)

where

$$T_{\text{sys}} = \eta_{\text{rad}} T_a + (1 - \eta_{\text{rad}}) T_p + T_{\text{rec}},$$  \hspace{1cm} (2.2)

$k_B \simeq 1.38 \times 10^{-23} \text{J/K}$ is Boltzmann’s constant, $B$ is the system bandwidth, $\lambda$ is the wavelength at the operating frequency, $S^{\text{inc}}$ is the incident power density, $G$ is the receiving antenna gain, $T_a$ is the received antenna noise temperature coming from external noise source such as the sky and ground, $T_p$ is the physical temperature of the antenna, $T_{\text{rec}}$ is the equivalent noise temperature of the receiver, and $\eta_{\text{rad}}$ is the antenna radiation efficiency.

Since the incident power density is purely determined by the transmitting antenna and propagation, it is convenient to use gain over system temperature ($G/T_{\text{sys}}$) for a receiving antenna as

$$\frac{G}{T_{\text{sys}}} = \frac{G}{\eta_{\text{rad}} T_a + (1 - \eta_{\text{rad}}) T_p + T_{\text{rec}}}. $$  \hspace{1cm} (2.3)

When only considering a parabolic reflector feed, the inverse sensitivity $T_{\text{sys}}/\eta_{\text{ant}}$ is a convenient receiver figure of merit that is independent of antenna aperture size with the antenna noise temperature expanded as

$$\frac{T_{\text{sys}}}{\eta_{\text{ant}}} = \frac{\eta_{\text{rad}} \eta_{\text{sp}} T_{\text{sky}} + \eta_{\text{rad}} (1 - \eta_{\text{sp}}) T_{\text{gr}} + (1 - \eta_{\text{rad}}) T_p + T_{\text{rec}}}{\eta_{\text{rad}} \eta_{\text{ap}}}, $$  \hspace{1cm} (2.4)
where \( T_{\text{sky}} \) is the sky temperature, \( T_{\text{gr}} \) is the ground temperature, \( \eta_{\text{ap}} \) is the antenna aperture efficiency, and \( \eta_{\text{sp}} \) is the spillover efficiency.

When considering a dual circularly polarized communication system, the partial inverse sensitivity including polarization efficiency can be defined as

\[
\frac{T_{\text{sys}}}{\eta_{\text{ant}} \eta_{\text{pol}}} = \frac{\eta_{\text{rad}} T_{\text{ant}} + (1 - \eta_{\text{rad}}) T_{\text{ext}} + T_{\text{rec}}}{\eta_{\text{rad}} \eta_{\text{ap}} \eta_{\text{pol}}},
\]

where \( \eta_{\text{pol}} \) can be expressed in terms of axial ratio (AR) as

\[
\eta_{\text{pol}} = \frac{(\text{AR} + 1)^2}{2(\text{AR}^2 + 1)}.
\]

Although there are many parameters in antenna design, (2.4) illustrates what final antenna parameters need to be considered in a wireless communication system. The S-parameters of the receiving antenna only influence \( T_{\text{rec}} \) in the equation. To simplify the analysis, the S-parameters are assumed to be perfect, i.e. no reflection and coupling, and detail discussion on the effect of S-parameters on the receiving performance can be found in Section 6.1. In this case, only radiation efficiency, aperture efficiency, and spillover efficiency, these three antenna parameters determine the final system performance for a parabolic reflector based ground terminal. Among these efficiencies, radiation efficiency plays the most important role because it can affect both signal and noise. Based on the typical ground terminal parameters, the impact of these three efficiencies on SNR can be plotted as shown in Fig. 2.1. As illustrated in Table 2.1, 1 dB radiation efficiency change can cause 2.4 dB SNR variation based on a typical MSA array feed.

For this reason, high radiation efficiency is the first priority for a feed antenna design, which becomes our research start point consequently. The rest of this chapter shows our promising solutions for passive high efficiency non-horn type compact feed antenna design.

<table>
<thead>
<tr>
<th>1 dB efficiency change</th>
<th>( \eta_{\text{ap}} )</th>
<th>( \eta_{\text{sp}} )</th>
<th>( \eta_{\text{rad}} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>SNR variation in dB</td>
<td>1</td>
<td>1.25</td>
<td>2.4</td>
</tr>
</tbody>
</table>
2.2 High Radiation Efficiency Microstrip Antenna

2.2.1 Introduction

Microstrip antenna arrays are a promising replacement for traditional bulky horn feeds for satellite communications terminals [9], and offer low profile, low manufacturing cost, integrability with circuit components, and beam steering capability, but the poor radiation efficiencies of MSAs is a serious obstacle to widespread adoption [6]. The sensitivity of a reflector and feed system is strongly influenced by radiation efficiency, because ohmic losses in the feed increase noise and attenuate the signal. As conventional metallic horn feeds have inherently high radiation efficiency, the efficiency of MSA arrays must increase in order to be competitive as feeds for satellite communications applications.

To address this issue, we have investigated the impact of key substrate parameters on the radiation efficiency of MSA arrays. Based on the analysis, guidelines for substrate selection were developed and used to design two MSA array feeds, which were fabricated and tested for efficiency comparison.
2.2.2 Substrate Parameters

The three critical properties of a substrate material for an MSA are relative dielectric constant $\varepsilon_r$, loss tangent $\tan \delta$ and thickness. Since $\tan \delta$ describes the dielectric loss directly, a low $\tan \delta$ material is nearly always preferable. The effects of substrate $\varepsilon_r$ and thickness will be illustrated by simulation results. A simple probe-fed MSA element based on 60 mil ISOLA IS680-280 resonant at 12 GHz was used as a reference design. The parameters $\varepsilon_r$ and thickness were varied, and for each parameter set, the resonant frequency was tuned to 12 GHz to ensure impartial comparisons.

Substrate relative dielectric constant

Theoretically, higher $\varepsilon_r$ should lead to lower radiation efficiency and narrower bandwidth [5], because the patch size becomes smaller and the electromagnetic field is more concentrated. Figure 2.2 shows this trend, but it is a weak effect, because the surface wave mode becomes more prevalent for high $\varepsilon_r$. Radiation patterns shown in Fig. 2.3 reflect this, as the surface wave deforms radiation patterns for higher permittivity. A material with higher dielectric constant tends to have a higher $\tan \delta$ as well. When miniaturization is not a key driver, low $\varepsilon_r$ materials are therefore preferred as recommended in [10].

Substrate thickness

As shown in Fig. 2.4, thinner substrate leads to lower radiation efficiency and narrower bandwidth. This is caused by narrower fringe field on the radiating patch sides, which can be viewed as a pair of thinner magnetic dipoles backed with ground plane. Although a thick substrate can improve radiation efficiency and bandwidth, unwanted radiation modes can be excited, degrading the antenna cross-polarization performance and broadside directivity, as illustrated in Fig. 2.5. Moreover, price is roughly proportional to substrate thickness, so thicker substrates are more costly for commercial applications. Based on these considerations, a thickness of 60 mils was selected for the feed designs presented below.
Figure 2.2: The influence of substrate permittivity $\varepsilon_r$ on radiation efficiency and bandwidth. Higher $\varepsilon_r$ leads to lower radiation efficiency and narrower bandwidth even on 60 mil thick material at 12 GHz. The trends would be more obvious on a thinner material with less surface waves launched.

Figure 2.3: The influence of substrate permittivity $\varepsilon_r$ on directivity pattern at the H-plane. As $\varepsilon_r$ increases, the pattern is deformed due to strong surface waves.
Figure 2.4: The influence of substrate thickness on radiation efficiency and bandwidth. Larger substrate thickness can improve radiation efficiency and bandwidth.

Figure 2.5: The influence of substrate thickness on directivity pattern at the H-plane. Higher cross polarization is caused by larger substrate thickness.
2.2.3 Array Designs

Two different $2 \times 2$ MSA array feeds were designed on 60 mil ISOLA IS680-280 for comparison. One is a typical MSA array and the other is a cavity-backed MSA array. The cavity-backed MSA array has top and bottom grounds stitched through tightly spaced vias and is well known for its surface wave-free operation [11], but it suffers from higher conductor loss due to more tightly restricted fields. The ground edges are a quarter wavelength at the center frequency from the patch edges, which provides enough ground area to obtain low back lobes while preventing strong diffraction and surface waves from deforming the main lobe. Both arrays consist of four half-wavelength separated patch elements, distribution feed network with quarter wavelength impedance transformers and vertical launched coaxial connector, as shown in Fig. 2.6.

2.2.4 Simulation and Measurement

A radiation efficiency measurement was conducted in a reverberation chamber [12] and the average value is used as the final result. Figure 2.7 shows that the measured efficiency agrees well with the simulation for both arrays. Compared with measured peak radiation efficiency 81% of the cavity-backed array, the MSA array can achieve 93% efficiency. To the best of our knowledge, this is the highest reported radiation efficiency for a $2 \times 2$ MSA array based on material compatible with an FR4 fabrication process [6, 7, 13].

Figure 2.6: Fabricated MSA and cavity-backed MSA arrays.
Figure 2.7: Simulated and measured radiation efficiency over frequency. The results agree well with the highest reported radiation efficiency 93% for a typical $2 \times 2$ MSA array.

2.2.5 Summary

We have developed substrate design guidelines for high efficiency microstrip antennas. Using 60 mil ISOLA IS680-280, common and cavity-backed MSA array feeds were designed, simulated and measured. The radiation efficiency of 93% was achieved for the MSA array, which is possibly the highest reported radiation efficiency for a $2 \times 2$ MSA array at Ku band to date. Still higher efficiency can be obtained by using a lower loss but more expensive and more difficult to manufacture PTFE-based composite material. The high efficiency design guidelines for MSA build a solid foundation for the following passive and active planar MSA array designs with more complexity and functionality.

2.3 High Efficiency Ku Band Stacked Shorted Annular Patch Antenna Feed

2.3.1 Introduction

Microstrip patch antennas are attractive for satellite communication (SatCom) applications due to low cost, low profile, and ease of fabrication. With careful substrate selection, high ef-
Efficiency microstrip patch arrays can be designed to serve as feeds for parabolic reflectors with performance nearly as good as traditional horn feeds and much smaller size and weight [14]. To achieve radiation efficiency comparable to a horn feed, however, low loss PTFE-based substrate must be used. This material is relatively expensive and requires special processes during PCB fabrication. The efficiency of microstrip patch array feeds is also reduced by the distribution network, and surface waves launched by patch elements on the substrate surface and unwanted radiation from transmission lines can decrease spillover efficiency. For these reasons, an improved patch antenna design is desirable for use as a compact, lightweight, low cost reflector feed.

Efficiency can be increased by eliminating transmission lines and exciting the antenna’s radiating structures using spatial coupling. One approach is to arrange coplanar parasitic patches around the driven patch and driving them through tight edge coupling [15]. This improves efficiency, but the antenna gain is not high enough to illuminate a dish efficiently. Another approach is to utilize natural power spreading via surface wave modes. In [16], planar circularly symmetric electromagnetic band gap (PCS-EBG) structures were used to direct the surface wave to radiate constructively across a large aperture with the substrate surface serving as the transmission line. The excited surface waves were not axially symmetric, leading to different E-plane and H-plane patterns and poor reflector illumination. A PCS-EBG based feed antenna would also be relatively large in size, leading to undesirable reflector blockage. These aperture coupled antennas can provide low loss, but for use as reflector feeds, higher gain and improved illumination patterns are needed.

Further improvements to gain and aperture efficiency while maintaining the low loss of a spatially coupled aperture are difficult to achieve and require a creative design. While most high gain antennas are arrays or other laterally extended structures that require lossy interconnects, gain can also be enhanced by increasing the axial extent of the antenna. This type of gain enhancement has been realized using dielectric superstrates, Fabry-Perot cavities, stacked EBG layers, and various types of lenses, but these techniques generally suffer from poor radiation efficiency and high fabrication cost, making them unsuitable as SatCom reflector feeds. This section provides a feed design based on a suspended shorted annular patch with a stacked parasitic patch that resolves all of these issues and realizes high radiation efficiency and high gain in a compact structure.
2.3.2 Antenna Design

The proposed antenna design combines both lateral and vertical extensions into a circular microstrip antenna to achieve high gain, high efficiency, and low side lobes. The shorted annular patch (SAP) antenna was developed to eliminate surface waves for microstrip patch antennas [17]. Our design exploits air substrate to mitigate surface wave and eliminate dielectric loss, and the shorting via array in [17] is replaced with a thin metal post. The shorting post is located at the center of the patches where the current of the dominant radiating mode $TM_{11}$ is negligible and acts merely as mechanical support. As a result, the patch diameter is approximately half the free space wavelength at 12 GHz.

The SAP antenna can achieve 10 dB gain, but it is not high enough to illuminate a typical SatCom reflector. A parasitic coupled patch is well known and commonly used for widening bandwidth, but a parasitic patch also influences gain, depending on the distance from the main patch [18, 19]. The shorting post of the SAP can be extended to the parasitic patch to support both patches, as long as the joint between the post and ground is strong enough. Since the shorting pin is at the center of both patches, fabricating and assembling such an antenna is relatively simple. The resulting SSAP feed design is shown in Figures 2.8.

Matching the SSAP at 12 GHz was accomplished by adjusting the radius of the main circular patch and the feed pin location like a normal patch design. As a dish feed, the pattern shape needs to be optimized for high aperture and spillover efficiency as well, which is mainly determined by the distance between patches and the radius of the parasitic patch. Using the model described in Section 2.3.3, the design was optimized for the maximum signal-to-noise ratio (SNR) when connected to a low noise amplifier and used as a receiving feed for a parabolic reflector. The parasitic patch maximizes the antenna gain at around half wavelength distance, which is related to the directors in a Yagi-Uda antenna but at a different spacing. To demonstrate the radiating mechanism, E-fields around the optimized SSAP on different planes are plotted in Fig. 2.9. The cross section view in Fig. 2.9(a) shows the top patch re-radiates based on its fringe field with opposite direction of the main patch E-field. Figs. 2.9(b) and 2.9(c) further show that the two patches work in the same mode but with $180^\circ$ phase difference. Considering the half-wavelength distance between the patches, the SSAP can be treated as an end-fire array of patch antennas with two elements.
A direct probe feed from a subminiature version A (SMA) connector or RF circuitry to the SAP was used in the prototype, but could readily be replaced with a transmission line feed or direct connection to a low noise amplifier in a final design. The design uses only a single layer PCB and retains its high efficiency even with low cost, low quality substrates.

At Ku band, besides linearly polarized free-to-air (FTA) signals, circularly polarized direct broadcasting service (DBS) signal are also provided. To receive DBS signals, a hybrid and the second orthogonal linearly polarized port are needed to create dual circular polarizations [20]. The SSAP can be easily fabricated on the ground of the hybrid.

### 2.3.3 Simulation and Measurement Results

To optimize the SSAP structure for the best illumination pattern as a reflector feed antenna, primary patterns were simulated first with an FDTD code, then secondary patterns reflected from the dish were calculated by the physical optics (PO) approximation. Spillover noise, sky noise, and electronics noise were included in the PO model to estimate the SNR [21]. This hybrid model was then embedded in an optimizer and the design was tuned to maximize the overall system SNR as shown in Fig. 2.10. The optimized SSAP configuration for a parabolic dish with a 0.6 focal length to diameter ratio (f/D) is illustrated in Fig. 2.11.

The simulated and measured S-parameters agree well as shown in Fig. 2.12. The simulated -10 dB reflection bandwidth is 480 MHz, which approximately covers the whole Free-To-Air (FTA) band from 11.7 to 12.2 GHz. The deviation between measurement and simulation of $S_{11}$ is caused by fabrication and assembly errors. Hand assembly leads to inaccuracies in the geometry.
Figure 2.9: E-field distribution at 12 GHz: (a) cross section view on a yz plane, (b) xy plane on the SAP, (c) xy plane on the top patch. The distributions show the two patches working in the same mode with $180^\circ$ phase difference.
Figure 2.10: Optimization flow chart for on-reflector simulation.

Figure 2.11: Configuration of the SSAP antenna feed with dimensions on side view.
such as a tilt of the fabricated spool relative to the ground plane and misalignment of the SMA connector with the center of the hole on the ground. In addition, the cases with no shorting post and no top patch are simulated as well, and the S-parameter results are included in Fig. 2.12. This shows that the shorting post only causes a small resonant frequency shift due to the current disturbance and the top patch does not affect the impedance match significantly. As a result, all methods capable of enhancing bandwidth for patch antennas can be applied to the SSAP. Multiple stacked patches or coupling feed are promising techniques [22, 23].

Simulated and measured radiation patterns in the E-plane and H-plane are shown in Figs. 2.13 and 2.14, respectively. The patterns were measured using a near-field scanner system. The simulated radiation patterns show a -8 dB taper at the dish rim and only -23 dB highest sidelobe level, which implies good spillover performance. The SSAP gain is stable as the ground size varies as shown in Fig. 2.15, so smaller ground can be used if the space requirement is strict. The SSAP gain at 56 mm ground size is 3 dB higher than that of SAP, verifying the radiating mechanism explained in Section 2.3.3.

The SNR for the fabricated antenna was measured with a commercial signal strength meter on a standard SatCom offset parabolic reflector which had an opening angle equivalent to a symmetric reflector with a f/D of 0.74. A horn feed and a microstrip antenna (MSA) array feed
consisting of $2 \times 2$ square patches [14] were also simulated and measured in the same setups. The three feed antennas are compared side by side in Fig. 2.16. The modeled horn was somewhat simplified compared to the actual horn, but was qualitatively similar in function. Table 2.2 lists the efficiencies and simulated and measured SNR at 12 GHz. The measured SNR agrees well with simulation in relative value between each feed type. The absolute difference in the measured and simulated SNR values is caused by unmodeled effects including feed support scattering, mispointing, atmospheric attenuation, and uncertainty in the satellite transponder EIRP. The measured SNR
fluctuates over a 4 dB range depending on outdoor conditions. As a result, the absolute difference is unimportant.

The measured SNR for the SSAP feed was 3 dB higher than the MSA feed and only 0.7 dB lower than the horn. Due to the absence of lossy distribution networks and dielectric substrates, the radiation efficiency of the SSAP is as high as that of the horn feed. The measured SNR value for the SSAP feed was reduced by an additional test connector not used in the horn feed setup. The additional connector loss was 0.2 dB, corresponding to roughly 0.5 dB reduction in SNR. Without this connector, the difference between the actual horn and SSAP performance would be even smaller.

Both SSAP and MSA feeds were optimized to a symmetric reflector with a 0.6 f/D. Tests were done with a 0.74 f/D dish so that spillover efficiency degrades. To demonstrate that the performance of the SSAP is better for dishes with smaller f/D and wider opening angle, the efficiencies and SNR of the three feeds were simulated as a function of f/D, as shown in Fig. 2.17 and 2.18. The SSAP performs best on an f/D = 0.56 dish with both efficiencies and SNR comparable to the best horn-fed dish. The MSA feed is poorer over all f/D values due to its higher sidelobe level and a flatter illumination pattern. By eliminating surface waves and introducing a director, the SSAP overcomes those difficulties and produces a highly efficient illuminating pattern.
Figure 2.16: Fabricated $2 \times 2$ MSA and SSAP feeds compared with the measured commercial horn feed.

Table 2.2: Simulated efficiency and measured SNR results of different antenna feeds at 12 GHz on an $f/D = 0.74$ dish

<table>
<thead>
<tr>
<th>Antenna feed</th>
<th>$\eta_{ap}$</th>
<th>$\eta_{sp}$</th>
<th>$\eta_{rad}$</th>
<th>$\text{SNR}_{\text{sim}}$</th>
<th>$\text{SNR}_{\text{meas}}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>MSA</td>
<td>62%</td>
<td>70%</td>
<td>93%</td>
<td>11.8 dB</td>
<td>9.5 dB</td>
</tr>
<tr>
<td>SSAP</td>
<td>75%</td>
<td>84%</td>
<td>98%</td>
<td>14.3 dB</td>
<td>12.6 dB</td>
</tr>
<tr>
<td>Horn</td>
<td>76%</td>
<td>92%</td>
<td>98%</td>
<td>15.2 dB</td>
<td>13.3 dB</td>
</tr>
</tbody>
</table>

2.3.4 Summary

A stacked shorted annular patch antenna was designed and optimized as the feed for a terrestrial reflector in Ku band satellite communications. The feed is lighter and simpler than a traditional horn feed. Without distribution networks and dielectric substrates, the antenna loss is lower than that of microstrip antenna feeds, and spurious radiation from transmission lines and surface waves is eliminated completely, leading to high radiation efficiency. The half wavelength stacked patch increases the gain and shapes the pattern to illuminate a standard Ku band SatCom dish efficiently. The overall SNR performance is only 0.7 dB lower than a commercial horn feed,
Figure 2.17: Simulated aperture, spillover, and radiation efficiencies of three types of feed as a function of f/D.

Figure 2.18: Simulated SNR of three types of feed as a function of f/D. Measured values on an f/D = 0.74 dish.
as verified by simulation and measurement on an $f/D = 0.74$ parabolic reflector. The SSAP is optimal for a dish with $f/D = 0.56$ and matches the best horn performance for that geometry. To our knowledge, this SSAP is the first reported non-horn feed antenna that achieves performance comparable to a horn feed.

Future work includes dual circularly polarized and wideband variations of the SSAP feed.

### 2.4 Radial SIW Fed Hexagonal Array Feed

#### 2.4.1 Introduction

Planar array feeds have appealing features including low cost, low profile, and easy fabrication. Several microstrip array feeds were reported in literatures. However, there are some limitations for current 2x2 array feed solutions. First, a 2x2 array feed may not be able to provide high spillover efficiency for a common $f/D$ dish. Its sidelobe level increases in spite of narrower main beamwidth with larger element distance. Second, a microstrip feed network on the same layer with patches has strong stray radiation if a thick material is chosen for high radiation efficiency, which decreases the spillover efficiency further. A stripline network can eliminate this problem, but the increased fabrication cost in multiple laminations and complex via configurations prohibits its commercial application. In addition, both these quasi-TEM transmission lines contribute a big portion of loss to the whole array performance. Only when radiation and spillover efficiencies are high enough, a planar array feed can compete with a commercial horn feed comprehensively. This white paper introduces a new radial substrate integrated waveguide (SIW) fed hexagonal microstrip array, which shows much improved radiation and spillover efficiency and bandwidth than a 2x2 microstrip array as a reflector feed.

#### 2.4.2 Design Concept and Configuration

The new array feed adopts a hexagonal configuration with six elements. Compared with a conventional 2x2 square array, it allows shorter distance between elements while creates higher gain, which means lower sidelobe levels and narrower main beam width in the primary gain pattern. They are two important factors to improve the spillover efficiency directly.
To feed this hexagonal array, a radial SIW power divider is designed. It provides the shortest feed distance for each element with equal phases and requires only one substrate instead of two in the stripline case. Moreover, SIWs are less lossy than striplines if the substrate is chosen correctly, due to absence of conductor loss. Because the SIW network is fully shielded except for the input connector and output pins to patches, the gain pattern would not be deteriorated by any stray radiation. The above features improve radiation efficiency and preserve the high spillover efficiency improved by the hexagonal configuration. The fully shielded SIW network would not increase the fabrication cost much, as only one substrate are needed between antenna and RF circuits, and transitions between layers can be accomplished by global vias without expensive back-drilling and sequential lamination.

The configuration of the hexagonal array feed is depicted as the following figure. A coaxial connector feeds the radial SIW from the center. The SIW distributes the power radially to six SIW-to-probe transitions, which feed six microstrip elements with equal amplitude and phase. A circular cavity is formed by adding a upper ground and vias to suppress the surface wave propagation.

### 2.4.3 Simulated Result

The S-parameter performance of the hexagonal array is shown in the following figure. The hexagonal array achieves 13% (1.6 GHz) bandwidth, which results from good impedance compensation for the feed pin under the microstrip element by the SIW-probe transition.

The radiation patterns of the array at the H-plane are compared with a commercial horn in the following figure. It shows the array feed can achieve symmetrical patterns with low sidelobe and low cross polarization as a horn feed, providing efficient illumination and avoiding spillover noise effectively.

Integrated with a PO based reflector model, the system performance for the hexagonal array feed was calculated. As shown in Table 2.3, this PTFE material based SIW-fed hexagonal array feed matches a horn feed in all efficiencies, leading to the same simulated SNR value. It has the best simulated G/T performance reported to date among non-horn type feeds.
Figure 2.19: Configuration of the hexagonal array feed: (a) antenna array layer, (b) SIW layer, (c) stack-up.
Figure 2.20: Simulated $S_{11}$ of the hexagonal array feed.

Figure 2.21: Simulated gain patterns of the hexagonal array feed.

Table 2.3: Simulated efficiency and SNR results of different antenna feeds at 12 GHz on an f/D = 0.74 dish

<table>
<thead>
<tr>
<th>Antenna feed</th>
<th>$\eta_{ap}$</th>
<th>$\eta_{sp}$</th>
<th>$\eta_{rad}$</th>
<th>SNR$_{sim}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>MSA</td>
<td>62%</td>
<td>70%</td>
<td>93%</td>
<td>11.8 dB</td>
</tr>
<tr>
<td>Hexagonal MSA</td>
<td>77%</td>
<td>92%</td>
<td>97%</td>
<td>15.2 dB</td>
</tr>
<tr>
<td>Horn</td>
<td>76%</td>
<td>92%</td>
<td>98%</td>
<td>15.2 dB</td>
</tr>
</tbody>
</table>
CHAPTER 3. MULTIBAND DUAL POLARIZED PASSIVE ARRAY FEEDS

A simple single band single polarization feed is not adequate for modern communication applications. Operating in multiband and dual polarization is commonly required for several scenarios.

One is two-way communication (Tx/Rx) in a designated radio band with Tx on one polarization and Rx on the other. The isolation between bands is required to be very high, at least 40 dB in the Tx band and as high as 100 dB for high power applications, so that the high transmitting power level would not affect the receiver performance.

Dual polarization is widely used in Satcom for operators to double the number of channels. Early launched satellite broadcasting services use linear polarization (LP), but circular polarization (CP) dominates for more recently introduced services. It is important planar array feeds can operate easily in dual polarization.

A new satellite with transponders in a different band can be collocated with an existing satellite to save the cost of another geostationary orbital slot. It requires terrestrial dish feeds operating simultaneously at both bands with the same phase center. In addition, the two satellites can operate in different polarizations, one in LP and the other in CP.

The following three sections demonstrate promising solutions for these scenarios, respectively.

3.1 Tx/Rx Planar Array Feed for Very Small Aperture Terminal (VSAT)

The traditional horn type feed for VSAT is very bulky, because an orthomode transducer (OMT) and 90° waveguide transition are needed. Comparatively, planar array feeds (PAF) offer significant advantages for commercial applications, including low profile, low cost and easy manufacturing. However, existing dual polarized planar antennas typically cannot achieve the required 40 dB isolation [24], which makes the receiver vulnerable to leakage of the high power signal from
the transmitter. To overcome this challenge, a high efficiency planar passive array feed antenna with over 50 dB isolation across the transmitting (Tx) band is presented.

3.1.1 VSAT Array Feed Design

The proposed array feed consists of four slot patch elements with two orthogonal polarizations for dual band operation respectively and two distribution networks for Tx/Rx band as shown in Fig. 3.1. Each antenna element is designed to receive a Ku band satellite downlink signal (11.7 to 12.2 GHz) and transmit a corresponding uplink signal (14 to 14.5 GHz), focused by a parabolic dish antenna.

The distance between array elements is a half wavelength at 12 GHz. A 1:4 equal phase distribution network for the Rx band and a 180° phase shifted distribution network for the Tx band are used. All parts are designed on one layer and fed by vertical mount SMA connector, as shown in Fig. 3.1. After LNA and PA are integrated on the back side, the feed points of the array can be directly connected with their outputs through vias or pins. The final product only needs a three layer printed circuit board (PCB), which reduces the fabrication cost tremendously.

Element design

The antenna element configuration is shown in Fig. 3.1. The top ground plane with vias along all sides prevents surface waves on the upper surface substrate, which would otherwise degrade the performance of the patch antenna in the efficiency and increase the backlobe level. Independent Tx/Rx dual band operation can be produced with reasonable isolation by exciting TM10 and TM01 fundamental orthogonal modes in different length from orthogonal patch sides. The feed points must be located at the middle points of their patch sides to prevent unwanted modes from being excited, which is important to guarantee good isolation.

Different feed structures for Tx/Rx bands are designed respectively for easier impedance match and higher isolation. For the higher Tx band, the impedance match is easier than that of the Rx band, because of the lower input impedance resulted from the longer patch side responsible for radiation. Quarter wavelength transformers can be applied effectively to match 50 Ohm transmission lines with the horizontal mode. For the Rx band, a capacitive coupling structure is used to
match the vertical mode with a 100 Ohm transmission line that makes easier match. The capacitive coupling structure is in a slim bar shape, which acts a match circuit for the antenna.

The array element designs are tuned so that the antenna provides a good impedance match, high Rx/Tx port isolation, low loss, and high radiation efficiency. In particular, the slot patch design optimizes the current distribution on the copper surface and the fields in the PCB dielectric to achieve much higher radiation efficiency than existing phased array antennas and enables usage of this antenna for Satcom applications.

**Feed distribution network**

The two feed networks for Tx/Rx on the same layer with improved isolation between the two outputs were designed as shown in Fig. 3.1. A simple equal phase feed network is used for Rx band to minimize loss and achieve high efficiency. A longer feed network surrounding the
outskirts of the array is used for the Tx band. Because the phase difference between the left side patches and right side patches fed from outside edges is 180 degrees intrinsically, the feed point for Tx network is therefore offset by 0.138 inch to produce 180 degree phase shift at the required operating frequency. Meanwhile, at the Rx port the energy coupled from the left two patches can be canceled out by the right two patches, because of the 180 degree phase difference. The feed network therefore achieves both proper Tx feeding phase and high isolation between Tx and Rx.

The array can be fed by a 50 Ohm microwave connector vertically for testing or a via to integrate with electronic circuits. The characteristic impedance values of the transmission lines along the Tx/Rx distribution networks are listed in Fig. 3.1. Grounded coplanar waveguide (GCPW) was chosen as the transmission line for its small size, low stray radiation, and good shielding properties.

3.1.2 Simulation and Measurement Results

The proposed array feed was fabricated on 32 mil Roger RO4003C material, which reduces the manufacturing cost and complexity drastically compared with traditional multi-stage horn feeds. Simulation and measurement are carried out to validate the design. The simulated and measured S-parameter results agree well for both bands, as shown in Fig. 3.2 and Fig. 3.3. The difference partly comes from the inaccuracy in the fabrication process and the SMA connectors feed structure. The isolation across the whole Tx band is around -55 dB, which can protect the receiver amplifier from strong transmit signal bleed-through. Isolation in the Rx band is also lower than 30 dB.

The measured and simulated radiation patterns of the antenna feed at 12.2 (Rx) and 14 GHz (Tx) are shown in Fig. 3.4 and Fig. 3.5. The peak gain of the measured pattern is 11.98 dBi, showing the good antenna efficiency. The -10 dB beamwidth is 90 degrees, equal to the spread angle of our reflector dish, which guarantees the good dish illumination and reasonable spillover efficiency.

3.1.3 Summary

A planar Tx/Rx dual band slot patch passive array feed antenna with high isolation and good satellite communication capability was designed and demonstrated. The array feed func-
Figure 3.2: Simulated and measured input reflection coefficient over frequency.

Figure 3.3: Simulated and measured isolation over frequency. The isolation achieves 55 dB in the Tx band.
Figure 3.4: Simulated and measured gain patterns at 12.2 GHz.

Figure 3.5: Simulated and measured maximum normalized patterns at 14 GHz.
tionally equivalent to a standard horn-type feed antenna plus OMT provides a promising way to minimize two-way link feed size and demonstrates that a planar antenna is a feasible and effective solution for low cost, high performance ground terminals. The same design concept can be applied to other satellite communications frequencies as well, including Ka band and C band.

3.2 Dual Polarized Array Feed

Broadcasting satellites use two orthogonal polarizations to double the number of channels. To receive a full channel list, a dish feed needs to operate in both polarizations. This section demonstrates dual polarization (linear or circular) can be realized conveniently for planar array feeds.

3.2.1 Dual Linearly Polarized Array Feed Integrated with RF Circuits

For an MSA array feed, dual linear polarization can be realized using square patch elements with the distribution networks on the antenna layer similar with the design shown in Section 3.1. Since the whole array only occupies one layer, low noise amplifiers and downconverter circuits can be integrated with the array feed conveniently through vias connecting the circuits with the array outputs. Fig. 3.6 shows a fabricated dual LP array feed integrated with RF circuits for Ku band free-to-air (FTA) services. The array is designed on a 60 mil RO4533 board, while the circuits are on another 20 mil substrate. A sheet of prepreg material is used to bonding the two substrates together. Signals from the output of the feed are downconverted and can directly be demodulated by an FTA receiver to get TV/radio streams. The planar array feed is much small than a horn feed and may have a lower manufacturing cost in high volume production, providing a promising replacement for current feed products.

3.2.2 Dual Circularly Polarized Array Feed

A dual circularly polarized array feed was designed and fabricated to receive direct broadcasting satellite (DBS) CP signals. Because a quadrature hybrid is needed to convert LP to CP, a multilayer PCB was adopted to reduce the array feed size, improve isolation between the two LP
feed networks, avoid unwanted radiation interference from the feed networks. The schematics of the array feed and the fabricated version are illustrated in Figs. 3.7 and 3.8, respectively.

Although the array feed achieves reasonable dual CP quality, the simulated radiation efficiency is only around 80% and the illuminating patterns are not narrow enough, leading to marginal DBS TV reception. One key to solve the problems is to use an intrinsically dual CP MSA as the array element, which eliminates the need of a hybrid and allows a more flexible array configuration. This important topic will be investigated thoroughly in Chapter 6.

3.3 Multiband Dual Polarization High Efficiency Array Feed for Ku/Reverse Band Satellite Communications

The direct-broadcast satellite (DBS) market demands more downlink bandwidth than is available in the existing Ku (11.7-12.7 GHz) and Ka (18.3-20.2 GHz) bands. To meet this need,
the International Telecommunication Union allocated 17.3-17.8 GHz as a downlink broadcast band in 1992 and the service was recently approved by the U.S. Federal Communications Commission. This application of the 17.3-17.8 GHz band is referred to reverse band (RB), to distinguish the downlink service from the existing use of this band for DBS feeder uplinks. Because this new service could lead to station-to-station interference problems at both the earth and space ends, multiple investigations have been carried out to resolve technical issues surrounding RB implementation [25].

To utilize the new band, a new satellite with RB transponders must be launched. Geo-stationary orbital slots are a rare resource and expensive to occupy. If the new satellite could be collocated with an existing Ku band satellite, the RB service could be introduced without requiring a new orbital slot. The key technical development required to enable this opportunity is a dish antenna feed for terrestrial terminals with phase centers of the Ku band and RB feeds at the same focal point and high isolation between bands.
Figure 3.8: Fabricated dual CP array feed for Ku band DBS: (a) antenna layer, (b) distribution network.

The required multiband feed must provide dual linear polarization (LP) in Ku band and dual circular polarization (CP) in RB. Multiband horn feeds are widely used in some applications [26], but the horn solution has complex machined feeding structures and high fabrication cost, and the phase center can shift with frequency. Planar passive array feeds have been introduced as a potential substitute for horn feeds due to low cost, low profile, easy fabrication, and the convenience for integration with low noise block (LNB) circuits [9, 27]. Multiband dual polarization planar arrays have been developed for synthetic aperture radar (SAR) [28, 29], and remote sensing [30]. For Ku/RB applications, the small separation between bands, the need for multiple polarizations, specified radiation pattern shape to feed a reflector, and high radiation efficiency make this an unusually challenging antenna design problem.

To meet these requirements, a dual band dual polarized planar passive phased array feed was designed and simulated, with four highly isolated ports and dual polarizations at each band.
To feed the antenna, a multilayer PCB technique was used. This raises an important research question regarding the achievable radiation efficiency of an antenna array with complex multilayer feed network. In the following subsections, the general configuration and concepts of the array feed are described first. Based on careful material selection with a focus on jointly maximizing antenna efficiency and manufacturability, the antenna element designs, the stripline distribution networks, and the complete stack-up configuration are presented. The simulation process for optimizing the illumination pattern is detailed as well. Various simulation results on the antenna and its corresponding system performance follow.

3.3.1 Multiband Array Feed Design

Array feed design concepts

Conventional multiband planar arrays use stacked patch [30] or perforated patch elements [28]. For satellite communications, high radiation efficiency and high isolation are required. Stacked and perforated patches may have poor radiation efficiency, and for multiband arrays, if the element spacing is optimized for one band, it is suboptimal for the other. Moreover, the low frequency ratio (1.439) of the Ku and RB satellite downlink bands makes high isolation difficult to achieve for collocated arrays. An interleaved array design is therefore superior to conventional approaches for the Ku/RB feed application.

The feed design goal is to illuminate a dish with f/D 0.6-0.9 efficiently. Based on the results of [9], the $2 \times 2$ planar array configuration was chosen due to its simplicity and high efficiency. To obtain collocated phase centers, an interleaved configuration for Ku and RB elements was adopted as shown in Fig. 3.9(a). In this way, the element distance for each band can be optimized independently to obtain the correct illumination pattern for the specified reflector geometry. The same configuration can be applied into a $4 \times 4$ or even larger array to illuminate a reflector with larger f/D efficiently or make the array active for beam steering as Fig. 3.9(b).

For microstrip-type antennas, a thick substrate is preferred to obtain high radiation efficiency and wide bandwidth [14]. On the other hand, distribution networks require a moderately thin substrate to avoid wide transmission lines, reduce radiation loss, and achieve low coupling between branches. Because of this conflict between substrate thickness preferences and also the
Figure 3.9: Interleaving array configurations for dual band operation with the same phase center: (a) $2 \times 2$ array, (b) $4 \times 4$ array.

complexity of implementing independent distribution networks for each band and polarization, a multilayer PCB technique was adopted.

**Ku and RB antenna element designs**

For a feed antenna, the overall performance is determined by the port input impedance, spillover efficiency, aperture efficiency, and radiation efficiency. Of these, a decrease in radiation efficiency has the largest detrimental impact on sensitivity, since loss in the antenna structure adds thermal noise and attenuates the received signal. Air or foam substrates are best for low loss, but are inconvenient in fabrication. Teflon is the next best choice, but its large z-axis coefficient of thermal expansion (CTE) increases the failure rate of vias during lamination and its weak mechanical strength often causes hole registration misalignment. PTFE ceramic perform better in these respects, at the expense of slightly higher loss tangent. As a result, RO3003 laminate (Rogers, Inc.) was chosen as the substrate due to its low dielectric constant, low loss tangent, smooth copper foil, and small thermal expansion coefficient.

In the interleaved array configuration, the issue of coupling between Ku and RB patches arises because of the close distance. Coupling not only wastes the received power, but also deforms beam patterns. Compared with conventional patch antennas, cavity-backed patch antennas have better isolation, confining the fringing field on radiating sides and eliminating surface wave modes [31]. Additionally, low cross polarization and good pattern shape can be maintained while
taking advantage of a thick substrate for high radiation efficiency and wide bandwidth [14]. The
locations of feeding pins are on the center axes to excite symmetric orthogonal modes for good
isolation, and can be adjusted for impedance matching. A shorting pin was added at the center of
each patch and extended to layers beneath the antenna array to suppress parallel plate modes and
improve impedance matching. In addition, two pairs of slots can be added into the patch to shrink
its size for both frequency band when normal patch size is too large for interleaving configuration.

Simulations were conducted using EMPIRE (IMST GmbH) based on the FDTD method.
Dielectric and conductor loss effects were taken into account. Specifically, surface roughness of
copper foils was modeled, since roughness can lead to significant loss as frequency increases. All
these considerations result in a precise efficiency estimate, which has been verified in [14] by some
rigorous reverberation chamber measurements. To speed up the simulation, modelling the cavities
and the following channelized stripline networks was simplified by using metal walls instead of
numerous vias to connect upper and lower grounds. In the optimization process, two parameters,
the antenna substrate thickness and the cavity size, are particularly important. A smaller cavity or
a thinner substrate confines the field more tightly, leading to lower cross-polarization and coupling,
but both radiation efficiency and bandwidth decrease. Finally, 60 mil thickness was chosen for the
antenna substrate. The optimized Ku and RB elements are shown in Fig. 3.10.

Distribution network designs

Channelized striplines were chosen for all networks due to their capability for superior
shielding property, which maintains good isolation and radiation patterns of the array feed. One
drawback is that substrates have to be hollowed to feed grounded co-planar waveguides (GCPW)
transformed from striplines. When integrating the feed with LNB circuits, direct transitions from
striplines to microstrips on the circuit layer can be made. Corporate feed networks were adopted
for both bands so the phase of each element can be kept the same across the operating band. Quar-
ter wavelength impedance transformers were used at every T-junction. For the RB network, four
hybrids were used after element feeding pins to form dual CP. Although sequential rotational con-
figuration could have improved the axial ratio, it is not practical with a single-layer feed network,
and may not be as effective for array feeds as it is for aperture arrays.
The thickness for each stripline substrate is 20 mil, which takes into consideration both stripline width for easy routing and loss for high efficiency. FEP (Dupont, Inc.) with the lowest loss tangent among common bonding materials was used in the two stripline subassemblies to satisfy the critical loss requirement. 2 mil thickness was chosen because it can cover the copper patterning while avoiding via failure due to thermal expansion. The final layouts of the Ku and RB feed networks are shown in Figs. 3.11(a) and 3.11(b).

**Complete dual Ku/RB array feed**

To feed the antenna elements from the two stripline subassemblies correctly, the feeding via stubs underneath striplines must be kept as short as possible, which can be realized by sequential lamination and back-drilling techniques. To bond the antenna substrate and two stripline subassemblies in sequential lamination, 2929 bondply (Rogers, Inc.) was chosen, because it is a thermal set prepreg which no longer flows after its first lamination cycle and has the cure temperature lower than the FEP melt point. 4 mil 2929 is needed to provide sufficient bonding strength.
Figure 3.11: Distribution network layouts for: (a) Ku band, (b) RB. Dimensions: $w_1 = 0.29$, $w_2 = 0.55$, $w_3 = 0.8$, $w_4 = 0.94$, $w_5 = 0.84$, $l_1 = 2.07$, $l_2 = 4.18$, $l_3 = 3.4$, $l_4 = 2.72$, $l_5 = 3.15$, $l_6 = 3.02$, $d_1 = 2.24$, $d_2 = 2.2$, $d_3 = 2.32$, $d_4 = 1.94$, $d_5 = 4.08$, $d_6 = 2.01$, $r_1 = 1.5$, $r_2 = 3$, $r_3 = 0.97$, $r_4 = 2.92$, $s_1 = 1.8$, $s_2 = 1.6$, $pad_1 = 0.66$, $pad_1s = 0.98$, Pad2 = 0.76, trw = 3.6, trl = 0.4, caw = 13, and cal = 5.4. Values are in mm.
between the ground layers. To reduce the complexity of the passive structures needed for the CP feed network and avoid intersections between feeding vias and striplines, the Ku band network was arranged between the antenna and the RB stripline network. The complete array structure is shown in Fig. 3.12.

Feed pattern optimization

The ideal reflector illumination pattern should be similar to the curve $\sec^4(\theta/2)$ with appropriate taper at the dish rim. Larger taper reduces spillover noise, whereas smaller taper increases aperture efficiency. Although -10 dB taper at the edge of the reflector is a common rule of thumb, better performance can be obtained by optimizing the antenna pattern for a particular dish geom-
Figure 3.13: Optimization flow chart for on-reflector simulation.

For the 2x2 dual band array, the illumination pattern can be controlled by adjusting element spacing.

To optimize element distances for the best illumination pattern, the array feed must be simulated with a reflector. This was accomplished by forming the primary patterns using an FDTD code and modeling reflection from the dish using the physical optics approximation as shown in Fig. 3.13. Electronics noise was included using microwave network theory following [21]. This hybrid model was then embedded in the same built-in optimizer in EMPIRE and the design was tuned to maximize the aperture efficiency, spillover efficiency, sensitivity, and overall system SNR. The optimized array configuration for an $f/D$ of 0.9 is shown in Fig. 3.10. The distances between elements for Ku band and RB are 14.7 mm ($0.6\lambda_0$ at 12.2 GHz) and 9.7 mm ($0.57\lambda_0$ at 17.55 GHz), respectively.

**Fabrication guidelines**

To shield the feed networks and provide high isolation, shielding vias and feeding vias with counterbores are required. These are fabricated through sequential lamination, and prepreg bonding materials must also be selected carefully. FEP (Dupont, Inc.) has the lowest loss tangent of commonly available materials, but being a thermal plastic bondply it would melt and flow in each lamination step, leading to poor hole registration and lamination integrity. 2929 (Rogers, Inc.) was chosen as the sequential laminating prepreg, because it is a thermal set prepreg which no longer flows after its first lamination cycle. The cure temperature is lower than the FEP melt point,
so FEP can be used once before the 2929 laminations. Therefore, FEP was only used in the two stripline subassemblies to satisfy the critical loss requirement, while Rogers 2929 were utilized to attach stripline subassemblies and the antenna core. Thickness is also critical. 2 mil FEP is chosen because it is thick enough to cover copper patterning while thin enough to avoid via failure due to thermal expansion. 4 mil 2929 is needed to cover copper features and provide sufficient bond strength. The final stack-up of the array is shown in Fig. 3.14.

Sequential lamination technology is adopted to realize buried and blind vias, which can feed antenna elements effectively. Dupont FEP film is chosen to bond the stripline substrates to minimize the transmission line loss, due to its perfect electrical performance (\(\tan \delta = 0.001\)).

The whole fabrication process is as follows: (a) etch copper into designed patterns for the antenna core and two upper cores of the stripline networks; (b) rout out cavity areas in three networks cores; (c) laminate two stripline subassemblies with FEP; (d) drill and plate all three sets of shielding vias; (e) laminate Ku network subassemblies and antenna core with Rogers 2929; (f) drill and plate Ku feed vias, then make counterbores; (g) laminate subassemblies with Rogers 2929; (h) drill and plate RB feed vias, then make counterbores; (i) rout out arrays from the whole panel.
Table 3.1: Simulated array feed efficiencies and system performance

<table>
<thead>
<tr>
<th>Port</th>
<th>$\eta_{ap}$</th>
<th>$\eta_{sp}$</th>
<th>$\eta_{rad}$</th>
<th>$T_{sys}/\eta_{ant}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ku V-pol</td>
<td>71%</td>
<td>77%</td>
<td>84%</td>
<td>308 K</td>
</tr>
<tr>
<td>Ku H-pol</td>
<td>73%</td>
<td>79%</td>
<td>83%</td>
<td>290 K</td>
</tr>
<tr>
<td>RB LHCP</td>
<td>68%</td>
<td>72%</td>
<td>81%</td>
<td>354 K</td>
</tr>
<tr>
<td>RB RHCP</td>
<td>68%</td>
<td>72%</td>
<td>79%</td>
<td>362 K</td>
</tr>
</tbody>
</table>

3.3.2 Simulation Results

Following optimizations of the array layout, the complete model of the dual band array feed was simulated. The S-parameters of all four ports are shown in Fig. 3.15. For Ku band, the -10 dB impedance bandwidth achieves 640 MHz from 12.15 GHz to 12.79 GHz, and the coupling between all ports is below -20 dB. Since hybrids were used in RB, the coupling between two CP ports is more meaningful than the reflection. The -15 dB bandwidth is 350 MHz, from 17.47 GHz to 17.82 GHz. Other coupling levels are below -26 dB.

Directivity patterns are illustrated in Fig. 3.16. Compared with the RB array, the Ku band array has a narrower beamwidth, lower cross polarization level, and higher sidelobe level. This leads to higher aperture and spillover efficiencies as listed in Table 3.1. The inverse sensitivity $T_{sys}/\eta_{ant}$ is a convenient receiver figure of merit that is independent of the dish size. The parameters of LNAs in the receiver were set to industry standard values with 0.8 dB $F_{\min}$. The radiation efficiency is approximately 80% for all four ports. To our knowledge, this is the highest simulated radiation efficiencies reported to date for such a complex multilayer antenna array design.

3.3.3 Summary

This work presents the first reported planar array feed design for collocated Ku/RB dual polarization satellite communications. With the proposed array feed on ground terminals, a new RB satellite can be launched to share the same orbital slot with an existing Ku band satellite. Simulation results have demonstrated that this planar array feed is promising for dual band receiving with good isolation between ports and high efficiencies for all four ports. The key development of this work is that careful attention to details in feed network design and material selection makes the
Critical difference between a planar antenna that does not approach viability in manufacturability or system-level receiver sensitivity and one that meets industry standard performance requirements for satellite communications.

Figure 3.15: Simulated S-parameters of the dual Ku/RB array: (a) Ku vertical polarized port, (b) Ku horizontal polarized port, (c) RB LHCP port, (d) RB RHCP port.
Figure 3.16: Simulated directivity patterns of the dual Ku/RB array: (a) Ku vertical polarized port, (b) Ku horizontal polarized port, (c) RB LHCP port, (d) RB RHCP port.
CHAPTER 4. ELECTRONICALLY STEERED ARRAY FEED FOR LIMITED SCAN RANGE BEAM STEERING

This chapter introduces the progress of one path to achieve limited scan range beam steering, which is to utilize active array feeds on a conventional stationary parabolic reflector. By setting beamforming weights for elements, different phase distributions on the dish aperture can be formed, resulting in steered beams. The main research issue is how to achieve limited scan range beam steering efficiently in low cost.

4.1 Introduction

Active array feeds provide a cost-effective solution to enable a conventional parabolic reflector antenna the beam-steering capability by replacing its low noise block feed (LNBF). A typical reflector antenna has the feed antenna at the focal point so that the phase distribution on the dish aperture plane is uniform leading to broadside radiation. If the feed antenna is offset from the focal point but still on the focal plane, the phase distribution varies approximately linearly across the dish aperture, which can be considered as a continuous phase shifted aperture. By locating the feed at different position on the focal plane, the beam can be steered off from broadside according to the phase shift rate [32]. This technology has been applied on spot beam satellites to provide customized programs for different areas [33] and multiple pixel receivers in radio astronomy [34]. The disadvantage of these array feeds is that each beam is fixed with one electrically large waveguide type feed and areas between two spot beams are not covered.

Arrays of electrical small elements following with beamforming networks provide a full coverage on foot print with the continuously changeable phase center on the focal plane, which has been implemented in Satcom [35] and radio astronomy [21]. If the feed location is too far from the focal point, the phase distribution cannot be approximated as linear any more, leading to necessary corrections with expensive phase shifters [36]. In radio astronomy, beamforming
is realized in digital processing based on digitized array signals after frequency downconversion, which offers all possible receiving beams. Changing the beam direction only requires different weights for each channel. The drawback is the cost of digital signal processing (DSP) servers, downconverting circuits, and analog-to-digital converter (ADC) required for each channel.

Considering the strict low cost requirement for commercial products, we used a totally different architecture, applying beam weights through variable gain amplifiers (VGAs) and combining each channel before downconversion. The weights are adjusted through the feedback loop through the power detection output based on calibration. Limiting the scan range makes an array feed without phase shifter affordable. In this way, the array feed can cost-effectively steer the secondary beam of a conventional dish by simply varying the gain of each active element. In this section, we demonstrate an $4 \times 2$ active array feed for one dimensional beam steering and show the progress of our $4 \times 4$ active array feed for future two dimensional sweeping.

### 4.2 $4 \times 2$ Electronically Steered Array Feed System

An electronically steered array feed (ESAF) system includes two subsystem, a $4 \times 2$ microstrip antenna (MSA) array with variable gain amplifiers (VGA) and a feedback system for the VGAs, as shown in Fig. 4.1. As demonstrated in [9, 20], a $2 \times 2$ microstrip array feed provides sufficient illumination for a typical offset parabolic reflector. To simplify the system requirement, our first objective was set to double the receiving range in one dimension by beamsteering. As a result, the array consists of four pairs of cavity-backed microstrip antennas which were chosen as its superior isolation performance. Each output of the antenna pair is connected to a low noise amplifier (LNA) to minimize the large loss occurred at the VGA stage so that an acceptable signal-to-noise ratio (SNR) can be maintained. A T-junction type corporate network combines signals from the four channels. The array feed was designed to receive free-to-air (FTA) signals from 11.7 to 12.2 GHz. The fabricated and assembled array feed is shown as Fig. 4.2.

A commercial low noise block (LNB) was utilized to downconvert the combined Ku band signal to intermediate frequency (IF), followed by a T-junction connector to allow TV receiver and the feedback system to process IF signals simultaneously. Since the LNB needs an external DC power supply from the receiver, a DC block was used to protect the feedback system. A rectifier type power detector was used to transform the IF signal into DC current. Because the
Figure 4.1: Block diagram of an active planar feed array for beam steering.

Figure 4.2: System board of the active planar feed array for Ku band.
detector can receive signals from 10 to 8000 MHz, an IF bandpass filter was added before it to exclude unwanted noise. To make the detector work in the linear response region, an IF amplifier was placed before the filter. A microcontroller reads the analog input, run acquiring or tracking algorithms, and output addresses and digital signals to the digital-to-analog converter (DAC). The analog signals are the control voltages for the four VGAs. This feedback system was constructed as shown in Fig. 4.3.

With the two subsystems, an LNB, and a TV receiver, on-reflector testing can be carried out as shown in Fig. 4.4. The dish was first pointed to a satellite providing FTA signals, Galaxy 19 in our experiment, by maximizing the SNR with a typical LNBF. The ESAF was then put on the dish with a hand-made mounting structure based on the positioning platform of a tripod which offers adjustments in roll, pitch, yaw, height, and focus. To ensure the center of the ESAP is located at the focus point, fine tuning in the five freedoms of the platform for the maximum SNR was conducted in the condition of activating the center two channels. Based on the above two alignments, optimal beam weights can be found for each dish position controlled by a precise motor. Fig. 4.5 shows
a measured SNR comparison of a horn feed and the ESAF as the dish was steered to different relative azimuth angle. Although the ESAF gives lower SNR values, its working range with SNR larger than 4 dB is two times the range of the horn. The variation of the ESAF SNR is caused by a simplified testing procedure which re-optimizes beam weights only when SNR drops below 4 dB. In addition, there were only five variable gain values used in the measurement to save search time. Lower SNR of the ESAF is caused by the packaged LNA with 1.5 dB higher noise figure (NF), lower antenna radiation efficiency, transmission line loss before LNAs, and large NF of VGA at some states.

Since our acquiring and tracking are based on the power detector output, the absolute values are meaningless because the output includes noise contribution which can be dominant. A calibration is necessary to extract the noise response of the different optimal beam weights, which was accomplished by pointing the dish to the cold sky and recording the power detector output. When the satellite was back to the steering range of the ESAF, the same set of beam weights was swept again with the power detector output recorded. The acquisition is finished by setting the
beam weights to the one with the largest difference between two output sets. The tracking mode was switched on after acquiring. To maintain the link, 3 point search method was used for tracking. In this way, the microcontroller only switches the beam weights to the two neighbors of current status. If a better position is found, the VGAs would be set to the corresponding weights. Our test showed the ESAF can track the satellite successfully as the dish is steered within the $4^\circ$ range maintaining SNR for Satellite video link.

### 4.3 $4\times4$ Electronically Steered Array Feed System

Based on the successful $4 \times 2$ ESAF, a 4x4 ESAF board was designed with improved performance based on high efficiency antenna elements and less transmission lines before LNAs. The top and bottom sides of the ESAF are shown in Figs. 4.6(a) and 4.6(b). The stackup consists of a 60 mil RO4003C laminate for cavity-backed microstrip antennas, a 20 mil RO4003C laminate for
RF front end circuits and a 16:1 combining network, and an FR4 epoxy laminate applied between the two RO4003C laminates. A thickness of 60 mil was chosen for the microstrip antennas to allow high radiation efficiency while keeping surface waves and cross polarization under reasonable levels [14]. Each microstrip element is backed by a via-fenced cavity to increase isolation between elements. The RF front end circuit is same with that of the $4 \times 2$ ESAF, except for the inner four channels which have a $360^\circ$ transmission line between LNAs and VGAs to fit all components on the same layer and simplify the 16:1 combining network. The combining network was realized by three stages of T-junctions and quarter wavelength transformer at the output of each junction. With VGAs for all 16 elements, this ESAF system can steer the beam in two dimensions with a $4^\circ \times 4^\circ$ range.

### 4.4 Summary

One dimensional limited scan range beam-steering was demonstrated experimentally by the $4 \times 2$ ESAF with VGAs. Acquiring and tracking functions were realized with the power detector based feedback system. The ESAF does not require phase shifters and uses a conventional parabolic reflector as the main radiator, leading to a low cost and efficient solution for limited scan
range beam-steering. To improve ESAF performance, high efficiency MSA element can be designed based on the guideline proposed in Section 2.2 and the transitions between antenna elements and LNAs should avoid using a long transmission line. These considerations were added into the $4 \times 4$ ESAF. Two dimensional beam-steering capability and better performance are expected and will be demonstrated experimentally in the future.
CHAPTER 5. LIMITED SCAN RANGE STEERED BEAM VSAT SYSTEM BY TILING 4×4 MICROSTRIP SUBARRAY TILE

This chapter introduces the progress of another path to achieve limited scan range beam steering, which is to excite passive subarray tiles instead of individual elements of a large array panel. This reduces the cost of RF front-end circuits significantly. Key research questions include: (1) How to design a high efficiency subarray tile with complex distribution networks? (2) Can the transmitting pattern of a full panel array comply with the pattern mask requirement set by FCC? The following two sections answer these two questions, respectively.

5.1 Planar 4×4 Array Tile Design for a Limited Scan Range Steered Beam VSAT System

5.1.1 Introduction

Very small aperture terminal (VSAT) service provides two-way communications for a wide area with geostationary satellites. A VSAT typically requires a parabolic reflector antenna fed by a horn antenna to establish a connection. Transmitting (Tx) and receiving (Rx) signals are divided and isolated by an orthogonal mode transducer (OMT). A block upconverter (BUC) and a low noise block downconverter (LNB) are responsible for Tx and Rx, respectively. All the above parts constitute an outdoor unit (ODU), which is so heavy and bulky that mount degradation occurs occasionally. Recent research on planar array feeds (PAF) [37,38] demonstrates a promising solution for limited scan range beam steering, but a PAF-fed reflector is still inconvenient and burdensome for applications where rapid terminal deployment and easy portability is required such as single soldier communications. To deploy an antenna conveniently and establish a communication link quickly, a compact lightweight planar antenna panel with beam-steering capability would be an ideal solution.

A full phased array panel can offer a broad beam-steering range and has been used for many radar applications. However, the high cost of transmitting/receiving (Tx/Rx) modules with
variable phase and gain for all antenna element prohibits many commercial applications [39]. On the other hand, if a fixed-beam antenna is used, pointing the main lobe precisely at a VSAT satellite requires a time-consuming installation. Since VSAT satellites are located on a geosynchronous orbit, VSATs do not need to track a satellite once it is acquired and locked in. A planar phased array with a limited scan range of beam steering offers a good trade-off between cost and functionality for this scenario.

The limited scan function can be realized cost-effectively by integrating multiple elements into one tile with a passive distribution network, providing a T/R module for each tile, and actively forming a beam using tiles as array elements. The cost of RF active components can be reduced proportionally depending on the amount of antenna elements in a tile. A $4 \times 4$ array tile only costs $1/16$ of a full active array on RF active components while still supporting limited scan range beam steering. To achieve high G/T and EIRP required for VSAT communications, design concept and practical consideration of a $4 \times 4$ Tx/Rx microstrip antenna (MSA) array tile was described.

5.1.2 Design Concept and Practical Considerations

The VSAT service discussed in this section operates at different frequencies with orthogonal linear polarizations for Rx and Tx. A $4 \times 4$ Tx/Rx MSA array tile essentially consists of 16 dual band MSAs and two 1:16 distribution networks for the two bands. From the fabrication cost perspective, it would be ideal to arrange the whole array on one layer, which can only be realized by using series-corporate feeding networks for both bands [40, 41]. The major issue for this solution is narrow operational bandwidth in impedance match and beam squint, which is a problem particularly for Rx where high G/T across 750 MHz bandwidth must be maintained. A corporate network should be used for Rx to meet the bandwidth requirement, while the loss in Tx effective isotropically radiated power (EIRP) due to beam squint can be compensated by adjusting the power amplifier output. Therefore, a multilayer distribution network is more practical for the situation.

Since the bottom side of an array tile is reserved for RF circuits, it is common to choose striplines to build distribution networks between antenna and RF substrates [20, 42] due to good shielding and straightforward design rules as microstrip lines (MSL). However, the fabrication of a multilayer PCB in such a stack-up requires three bonding processes and complex back-drilling for the transitions between layers. Instead, substrate integrated waveguides (SIW) provide perfect
shielding and high power capacity without high fabrication cost due to easy stack-up and has been used widely to excite MSA array through slots [43, 44] or coupling probes [45]. Because to feed a dual polarized MSA directly from an SIW is not realistic, the distribution network can be divided into two stages with the first stage on the antenna layer and the second stage in the SIW layer [46].

Electrical performance and manufacturability are the two main factors to be considered in substrate material selection. Due to the importance role radiation efficiency plays in signal-to-noise ratio (SNR) [47], a low loss tangent substrate should be used. Air or foam substrates are ideal for low loss, but are inconvenient in fabrication. Teflon is the next best choice, but its large z-axis coefficient of thermal expansion (CTE) increases the failure rate of vias during lamination and its weak mechanical strength often causes hole registration misalignment. PTFE ceramic perform better in these respects, at the expense of slightly higher loss tangent. As a result, RO3003 laminate (Rogers, Inc.) was chosen due to its low loss tangent, smooth copper foil, and small thermal expansion coefficient. Bondply 2929 (Rogers, Inc.) was chosen as the sequential laminating prepreg, because it is a thermal set prepreg which no longer flows once its cure process is done.

Besides material, right substrate thickness is important for achieving high performance. For an MSA working at Ku band, 60 mil thickness can result in high radiation efficiency and wide bandwidth with acceptable cross polarization [14]. SIWs also require use a thick substrate for lower loss. However, substrate material cost is approximately proportional to thickness, and a thick multilayer stack-up can increase the failure rate of via plating. Therefore, 60 mil thickness is chosen for SIWs as a trade-off between those factors. 20 mil thickness is used for the RF circuit substrate to facilitate the use of MSLs.

Element design

The antenna element can be a cavity backed rectangular microstrip patch antenna with modes $TM_{10}$ and $TM_{01}$ resonant at different frequencies. The cavity is used to mitigate the surface waves launched by a thick-substrate MSA [16, 31] so that patterns are not deformed and coupling between elements is reduced. Although 60 mil thickness RO3003 was used, a bare rectangular MSA cannot achieve the Rx bandwidth requirement. To improve the antenna bandwidth, two parasitic side-coupling strips were added besides the Tx radiating edges to widen the Rx radiating
edges [15]. The MSA is fed by a high impedance MSL from the Tx port and a coupling-bar from the Rx port. In this way, no slot is needed on the center patch leading to intact orthogonal modes underneath.

**CPWG feed network**

To support broadband satellite communications, the combining networks for the array tile should be highly efficient and provide equal amplitudes and phases across the operating bandwidth. Although a series feed network introduces lower loss than a cooperate feed due to shorter distance, its output phase distribution varies with frequency, resulting in narrow bandwidth and beam squint. As discussed in Section 5.1.2, a parallel combining network should be used for Rx to ensure the wideband high G/T performance. Since MSLs can have severe spurious radiation on a 60 mil substrate particularly at discontinuities and high coupling between lines, grounded coplanar waveguide (CPWG) was chosen to build the first stage of the distribution networks.

To fit the Tx distribution network into the rest of the tile area, a series-parallel hybrid network was proposed to feed Tx ports. To keep the network simple, the distance between elements in x direction shrinks to 360° electrical length for the CPWG so that only a quarter-wavelength 25Ω CPWG is needed for impedance matching. In addition, another 2-to-1 CPWG combining network can be arranged in the saved area. The 8-to-1 combining point needs to be shifted 90° electrical length to align antenna feeding phases of two adjacent rows. The opposite feed point shift for the upper and lower two rows introduces another 180° phase difference. By combining these two outputs in the SIW layer with 180° out of phase, the Tx signal reaching each antenna element has the same phase, while the coupled Tx power at the Rx port on two levels is cancelled, leading to high isolation between Tx and Rx [9].

**SIW network and transitions**

The second stage of distribution networks based on SIW lies between the antenna layer and the RF layer. Transitions between different layers with small reflection and coupling are important. Although there are various types of transitions between an MSL and an SIW on different layer [48], a direct shorting via feeding backed with a shorting wall is used due to its simplicity and
easy fabrication. However, a thin prepreg between two substrates can couple some power transformed from a via transition mode to a parallel plate mode. Generally, this issue can be solved by putting global shielding vias around transition vias. However, the smallest area surrounded by shielding vias in this stack-up is the same with the SIW networks, which results in cavities in the prepreg layer with resonant frequencies close to the operating frequency of the SIWs. Due to the material property, the resonant cavity can have a very low quality factor, so power coupled from via transition and dissipated inside the prepreg can be large. For this reason, shielding vias were not put on the board, and the small amount of coupled power just propagates through parallel plates and either radiates at edges or dissipates inside the material.

A wider SIW width makes the field distribution inside an SIW less concentrated, leading to lower loss, while the width has to be limited by the tile size. At each junction of the 2-to-1 SIW combiners, ridges can be added to minimize the negative effect of discontinuities on S-parameters. In this way, the Tx performance from S-parameters to radiation patterns is sacrificed in order to realize the best Rx performance. More balanced Tx/Rx performance in terms of bandwidth and beam squint, if required, can be achieved by substituting series-corporate networks for pure corporate networks for Rx and re-optimizing the Tx series-corporate networks [49].

### 5.1.3 Summary

Phased-tile arrays can be a valuable solution for SatCom applications requiring beam steering only in a limited scan range. A high efficiency $4 \times 4$ MSA array tile can provide a solid foundation for future phased-tile array design. To achieve high performance required by SatCom within restricted tile size limits, the layer stack-up configuration is carefully discussed with considerations on manufacturability, cost, and efficiencies of MSA, distribution networks, and transitions. A method to enhance bandwidth and radiation efficiency of a rectangular Tx/Rx dual band MSA was proposed. Distribution networks consisting of two CPWG and SIW stages for Tx and Rx to minimize fabrication cost and maintain good performance are described.
5.2 Tile Array Simulation

5.2.1 4×4 Tile Array

Currently a full aperture array for commercial satellite communication is not available because of the prohibitive cost and design complexity. These disadvantages are caused by the fact that each antenna element needs its own transmitting/receiving (T/R) module with phase and gain controller inside. This architecture is necessary for scenarios like radar, which need to scan the beam across a broad range. However, a limited scan range is actually able to satisfy requirements for many commercial applications such as mobile direct broadcasting satellite and very-small aperture terminal (VSAT), for which the target is stationary and the antenna can be roughly pointed to the satellite at installation.

The limited scan function can be realized cost-effectively by integrating N×N elements into one tile with a passive distribution network and only providing a T/R module for one tile. The cost of the active components will drop significantly relative to the original full aperture array, which is an appealing feature especially for commercial products. One key for this limited scan array is to design a high efficiency active N×N sub-array. The final array will unite M×M tiles with a combining network and the total size is determined by the application requirement. A 4×4 tile array with each tile consisting of 4×4 elements is shown in Fig. 5.1.

Since elements inside a tile need to be fed by passive networks, the boarder area of the tile may be used to facilitate the network design, especially for dual polarization systems (two way or same frequency). In this case, the distance between tiles \(d_t\) would be larger than 4 times the element distance \(d_e\) referred in Fig. 5.1. A simple 1-D array factor equation was used to calculate the array factors of 4 elements, 4 tiles, and 16 elements for two different cases as shown in Fig. 5.2 and 5.3. The results show that unequal element distance at tile junctions causes a non-monotonic sidelobe level, which makes the pattern fail the sidelobe level requirement. The reason for this phenomenon is the misalignment between the grating lobes in the tile array factor and the zeros in the element array factor. As a simple rule of thumb, when implementing a rectangular array, consistent element distances should be used to suppress unwanted sidelobes.

A second issue is edge tapering. To serve as a VSAT terminal, the transmitting pattern of the whole antenna must comply with the pattern mask requirement as shown in Fig. 5.4. The
array factor pattern at 13 GHz of a $16 \times 16$ array with element distance equal to 16 mm, which is necessary for an acceptable link between terrestrial terminals and a VSAT satellite, was added into the same figure. The comparison illustrates that it is hard to meet the mask requirement if uniformly exciting all elements. To solve that, edge element excitation tapers on the array have to be implemented.

Besides the above issues, there are three additional difficulties in antenna design for VSAT application. High antenna efficiency is required to achieve the required G/T. High isolation between transmitting and receiving (Tx/Rx) is important for good receiving performance. The last challenge is the need for low loss distribution networks for both bands while maintaining low cost. These challenges have been addressed in Section 5.1.

5.2.2 Summary

Limited scan range beamsteering can be implemented cost-effectively by using a tile array with each tile consisting of a passive network fed subarray. To maintain a monotonic sidelobe level, the same element distance through the whole array panel must be maintained. The sidelobes of an uniformly excited array exceeds the regulatory pattern mask, requiring edge element tapers. With these design aspects, a viable tiled array for Satcom applications can be realized.
Figure 5.2: Array factors of 16 elements with different element distance inside a tile and between tiles at 13 GHz.

Figure 5.3: Array factors of 16 elements with same element distance inside a tile and between tiles at 13 GHz.
Figure 5.4: Uniformly distributed gain patterns, mask requirement, and array factor of a $16 \times 16$ array with element distance equal to 16 mm at 13 GHz.

Figure 5.5: Gain patterns after applying 1:2 taper on the outmost 3 elements.
Figure 5.6: Gain patterns after applying 1:2 taper on the outmost 4 elements.
Dual orthogonal polarization is widely used to improve the data throughput of satellite-based voice, data, and video services. Early systems used linear polarization (LP) due to the simplicity and low cost of feed antennas. For more recently introduced satellite communication (Satcom) services, circular polarization (CP) dominates because it avoids the need for polarization alignment of terrestrial feeds. To achieve high quality dual CP operation, SatCom terminals normally use a horn feed antenna with a dielectric septum in the waveguide transition [50].

Dual CP horns have good polarization performance and achieve the high radiation efficiency required for SatCom applications, but are large and heavy. For lightweight mobile terminals, it is desirable to replace bulky horns feed with a low profile planar phased array feed (PAF). Planar PAFs offer low cost fabrication, simple mass production using PCB technology, and can be readily integrated with amplifiers and other transceiver electronics circuits [9, 14, 20]. PAFs also allow cost-effective limited scan range electronic beam steering [38].

For microstrip antennas (MSAs) that could be used as elements in electronically steered phased arrays and array feeds, coupling between orthogonal linear polarized ports is typically very low (-30 dB or less). However, coupling between ports of an intrinsically dual CP MSA (not relying on a quadrature hybrid) is much higher [51]. Section 6.1 analyzes the effect of mutual coupling on the sensitivity of dual polarized receivers. The coupling effect is further investigated in Section 6.2 based on a proposed equivalent circuit model and Jones matrix formulation. Applications of the analysis including an efficient decoupling method are demonstrated.
6.1 Effect of Mutual Coupling on the Sensitivity of Dual Polarized Receivers

6.1.1 Introduction

Coupling between ports of a dual-polarized antenna impacts various types of communications links differently. For a diversity antenna or multiple input multiple output communications channel, coupling reduces the amount of unique information contained in the two output signals. For application as a feed antenna for dish-type satellite communications terminals, it is possible for the cross-polarization discrimination of the secondary beam formed by the dish to be adequately high, while the coupling between feed antenna output ports for the orthogonal polarizations can still be poor. In this case, the impact of the coupling is not related to poor diversity, but instead it leads to an increase of the equivalent receiver noise temperature of the system. In order to enable better use of CP microstrip antennas for compact electronically beamformed antenna systems, we present an analysis of the impact of mutual coupling on the performance of high-sensitivity dual-polarized receivers for satellite communications applications.

6.1.2 Correlation Coefficient and S-parameters

The envelope correlation is commonly used to analyze coupling in multi-antenna communication systems. This quantity is the correlation between output signals for a pair of antennas in a multipath environment with uniformly distributed incoming waves and is related to the antenna S-parameters by [52]

\[ \rho_e = \frac{|S_{11}^* S_{12} + S_{21}^* S_{22}|^2}{(1 - |S_{11}|^2 - |S_{21}|^2)(1 - |S_{22}|^2 - |S_{12}|^2)}. \]  

(6.1)

This expression shows that no mutual coupling is a sufficient condition for zero correlation, but zero correlation is only a necessary condition for no coupling.

To understand the practical import of this result, a dual CP slotted circular patch antenna was designed and simulated. The antenna produces radiation patterns with highly orthogonal circular polarizations across the main beam range at 12.5 GHz, as shown in Fig. 6.1. It leads to a low value for the correlation coefficient at 12.5 GHz, as shown in Fig. 6.2. On the other hand, as is commonly observed for planar dual CP antennas, the port coupling is still large. At the design
center frequency, $S_{12}$ is -7.3 dB, which does not meet the common industry requirement of -20 dB for CP satellite communication feed antennas. This example illustrates the difficulty of achieving high port isolation for planar dual CP antennas and highlights the need to understand the impact of isolation on the performance of the receiver system.
6.1.3 Effect of Mutual Coupling on Sensitivity

When high coupling between two orthogonal ports exists, the reverse noise waves emitted from the input ports of low noise amplifiers (LNAs) couple into other channels, increasing system noise [53]. The noise degradation can in principle be eliminated by impedance matching the LNAs to the active impedances at the antenna ports, but in practice the active impedance can have a negative real part and perfect matching is not practical.

An alternate approach to minimizing receiver noise under coupling is to eliminate the LNA reverse noise wave. To simplify the analysis of this possibility, the following assumptions are made: the system impedance $Z_0$ is real, the amplifier optimal source impedance is $Z_{opt} = Z_0$, the reflection coefficients of the antenna ports are $S_{11} = S_{22} = 0$, and all components in the two receiver channels are identical.

We will characterize the LNA through the optimal source reflection coefficient $\Gamma_{opt}$, the minimum noise figure $F_{\text{min}}$, and noise resistance $R_n$ [54]. The condition for no reverse noise wave can be shown to be [55]

$$R_n = \frac{T_{\text{min}}Z_0}{4T_0},$$

where $T_0 = 290K$. This value happens to be the lower bound for $R_n$ [56].

Active reflection coefficients can be found from the following equation [56]

$$\Gamma_{\text{act},m} = \frac{1}{w_{f,m}^*} \sum_{n=1}^{N} w_{f,n}^* S_{A,nm},$$

where $w_f$ is the beamformer weights referred to forward wave amplitudes at the antenna input ports with matched loads. The beamformer weights at the receiver output ports, $w$, relate to $w_f$ through $w_f = G^H w$ where $G = g\sqrt{Z_0}(I + S_R)(I - S_A S_R)^{-1}$ and $S_R$ is the scattering matrix corresponding to the receiver impedance matrix $Z_R$.

Assuming $Z_R = R_0 I$ and $w^H = [1 0]$, $w_f^H$ would be equal to $[g\sqrt{Z_0} 0]$. So $\Gamma_{\text{act},1} = S_{11} = 0$ and $\Gamma_{\text{act},2} = w_{f,1} S_{12}/w_{f,2} = \infty$. Based on these results, the receiver temperature can be calculated using [56]

$$T_{\text{rec}} = \frac{\sum_{m=1}^{M} |w_{f,m}|^2 (1 - |\Gamma_{\text{act},m}|^2) T_m}{\sum_{m=1}^{M} |w_{f,m}|^2 (1 - |\Gamma_{\text{act},m}|^2)},$$

(6.4)
where
\[ T_m = T_{\text{min},m} + \frac{4R_{n,m}T_0|\Gamma_{\text{act},m} - \Gamma_{\text{opt},m}|^2}{Z_0|1 + \Gamma_{\text{opt},m}|^2(1 - |\Gamma_{\text{act},m}|^2)}. \] (6.5)

Substituting \( R_n, \Gamma_{\text{opt}} \) into \( T_m \) yields
\[ T_m = \frac{T_{\text{min}}}{1 - |\Gamma_{\text{act},m}|^2}. \] (6.6)

This leads to the receiver noise temperature
\[ T_{\text{rec}} = \frac{T_{\text{min}}}{1 - |S_{12}|^2}. \] (6.7)

The denominator is identical to the ratio of radiated power to input power in a dual-polarized transmitting scenario without reflection and loss, which is a consequence of reciprocity. Practical receivers have noise resistance \( R_n \) larger than the minimal value which can be expressed as \( R_n = nR_{n,\text{min}} \), then the receiver temperature can be obtained from
\[ T_{\text{rec}} = \frac{T_{\text{min}}[1 + (n - 1)|S_{12}|^2]}{1 - |S_{12}|^2}. \] (6.8)

Assuming the analyzed receiver working at Ku band has a perfect antenna with 100% aperture and radiation efficiencies and its \( F_{\text{min}} \) is equal to 1 dB, the inverse sensitivity defined in Section 2.1 as a function of \( S_{12} \) at different values of the LNA noise resistance parameter \( R_n \) is shown in Fig. 6.3.

If reflection is not negligible, the receiver temperature in terms of noise resistance, reflection, and coupling coupling can be obtained from
\[ T_{\text{rec}} = \frac{T_{\text{min}}[1 + (n - 1)|S_{11}|^2 + (n - 1)|S_{12}|^2]}{1 - |S_{11}|^2 - |S_{12}|^2}. \] (6.9)

6.1.4 Summary

Low envelope correlation coefficient in a dual polarized receiver does not necessarily mean low mutual coupling. Port coupling, which can be difficult to minimize for planar dual CP antennas, increases system noise and reduces receiver sensitivity. The impact of coupling on receiver
performance is minimized when the LNA noise parameters are such that no reverse noise wave is emitted, and this case provides a lower bound on the sensitivity degradation due to coupling. If the LNAs are designed optimally, the coupling can be -10 dB or poorer with minimal impact on sensitivity.

An intrinsically dual CP antenna with good CP quality and high coupling is suboptimal and may be insufficient for Satcom. In the following section, the coupling between the two ports of an intrinsically dual CP MSA is analyzed based a new method, and a solution for the high coupling problem is also proposed.

6.2 Analysis of Intrinsically Dual Circularly Polarized Microstrip Antennas Using an Equivalent Circuit Model and Jones Matrix Formulation

6.2.1 Introduction

Low-profile PAFs for circularly polarized SatCom services require a compact dual CP planar antenna element design with high port isolation, high radiation efficiency, and good polarization quality. Port isolation may seem to be less important than efficiency and polarization perfor-
mance, but it is actually a critical requirement for SatCom terminals. Coupling between ports on the receive side means increased noise and lower SNR even with perfect design of the electronics [2,56]. For SatCom terminals that receive on one polarization and transmit on another, typically with the two ports tuned to adjacent frequency bands, coupling of the transmitted signal into the receive signal path also degrades SNR.

No existing dual-CP planar antenna simultaneously realizes the design goals of high isolation, high efficiency, and high quality polarization. Linearly polarized waves can be converted to CP with an EBG surface [57] or a planar circular polarizer [58]. The surfaces are narrowband, increase manufacturing cost, and reduce radiation efficiency. Traveling wave and leaky wave antennas such as the radial line slot antenna (RLSA) [59] and the cross antenna [60] also generate CP. For these types of dual CP antennas, the termination at the non-driven port must absorb some of the power from the active port to reduce reflected waves, which degrades the radiation efficiency and noise performance. These antennas are also relatively large and are difficult to integrate into phased arrays.

Another approach is to excite the ports of a dual linear polarized antenna with a 90° phase difference using a quadrature hybrid. Dual linear polarized patch antennas with a well-designed hybrid offer excellent dual-CP polarization performance. The hybrid occupies a large amount of PCB real estate and increases the size and cost of the planar feed. More significantly, it adds loss, which for SatCom applications means increasing the size of the reflector antenna required to achieve an adequate link margin.

To eliminate the drawbacks of additional structures to convert from linear to circular polarization, the obvious solution is a two-port microstrip antenna (MSA) that intrinsically radiates dual CP fields. Intrinsically dual-CP MSAs have been designed and used as elements in electronically steered phased arrays and array feeds since the early 90s [51]. Early designs suffered from poor isolation between the antenna ports, and surprisingly little work have been done on intrinsically dual-CP elements since then. A satisfactory dual-CP MSA is one of the few major remaining unsolved problems in antenna theory.

To overcome the port isolation challenge with intrinsically dual-CP MSA designs, we develop a new equivalent circuit model for dual-CP patch antennas and derive a Jones matrix description of the polarization properties of the two-port antenna. The model answers the following
open questions: (1) One might think that orthogonality of radiated modes would correspond to zero coupling between the antenna ports. How it is possible for an MSA to achieve good cross polarization discrimination/isolation (XPD/XPI) while at the same time exhibiting poor isolation? (2) What is the feasible region for the tradeoff between XPI and impedance mismatch factor? (3) How does the impedance bandwidth of an MSA relate to the axial ratio bandwidth? (4) How can the Wheeler cap method for measuring radiation efficiency be applied to dual-port antennas with multiple adjacent resonances? (5) Is there a simple and effective decoupling network that can improve port isolation while retaining high XPI? Section 6.2.3 deals with question 1. Questions 2 and 3 are analyzed in Section 6.2.4. The Wheeler cap method and decoupling scheme are discussed in Section 6.2.5 as applications of the proposed circuit model and Jones matrix representation. This work will help to enable a new generation of planar PAFs and phased arrays for circularly polarized SatCom services.

6.2.2 Jones Matrix

Communication systems utilizing orthogonal polarizations as two independent channels can be analyzed using the Jones matrix formulation [61]. The system Jones matrix for a dual-port antenna system links its circuit parameters with electromagnetic (EM) field properties. A block diagram for a typical dual-polarization communication system is shown in Fig. 6.4 with Jones matrices for system components indicated.
The transmitting (Tx) and receiving (Rx) antennas can be modeled as series equivalent circuits. The antennas have two ports with each corresponding to one orthogonal polarization channel. By modeling the power amplifier (PA) as a Thévenin equivalent circuit, the excitation currents for the Tx antenna are

\[ i_A = (Z_{A,Tx} + Z_S)^{-1} v_g = J_g v_g \]  

(6.10)

in terms of the antenna and source network impedance matrices \( Z_{A,Tx} \) and \( Z_S \) and the source open circuit voltage vector \( v_g \). The radiated electric (E) field is related to the far-field patterns of the Tx antenna by

\[
\begin{bmatrix}
E_{u,\text{rad}}(\vec{r}) \\
E_{v,\text{rad}}(\vec{r})
\end{bmatrix} = \frac{1}{I_0} \begin{bmatrix}
E_{u1,Tx}(\vec{r}) & E_{u2,Tx}(\vec{r}) \\
E_{v1,Tx}(\vec{r}) & E_{v2,Tx}(\vec{r})
\end{bmatrix} \begin{bmatrix}
i_{A,1} \\
i_{A,2}
\end{bmatrix} = J_{\text{Tx,oc}} i_A,
\]  

(6.11)

where \( u \) and \( v \) indicate orthogonal components of the field corresponding to the principal polarizations for port 1 and 2 of the antennas respectively. \( E_{pn,Tx}(\vec{r}) \) is the radiated field in the \( \hat{p} \) polarization obtained by exciting port \( n \) of the Tx antenna with input current \( I_0 \) and all other non-driven ports open circuit loaded (OCL).

By the electromagnetic reciprocity theorem, the open circuit voltages induced at the Rx antenna terminals by the incident field is

\[
\begin{bmatrix}
v_{oc,1} \\
v_{oc,2}
\end{bmatrix} = \frac{c_1}{I_0} \begin{bmatrix}
E_{u1,Rx}(\hat{k}_{inc} r) & E_{v1,Rx}(\hat{k}_{inc} r) \\
E_{u2,Rx}(\hat{k}_{inc} r) & E_{v2,Rx}(\hat{k}_{inc} r)
\end{bmatrix} \begin{bmatrix}
E_{u,inc} \\
E_{v,inc}
\end{bmatrix} = J_{\text{Rx,oc}} E_{\text{inc}},
\]  

(6.12)

where \( c_1 = 4\pi jre^{jkr}/(\omega \mu) \). \( E_{pn,Rx}(\hat{k}_{inc} r) \) is the field in the \( \hat{p} \) polarization radiated by the Rx antenna in the incident wave direction when excited as a transmitter with the input current \( I_0 \) in port \( n \) and the OCL condition for non-driven ports. The loaded voltage at the antenna terminals is

\[ v_L = Z_L (Z_{A,Rx} + Z_L)^{-1} v_{oc} = J_L v_{oc}. \]  

(6.13)
The complete transceiver system can be described by concatenating the Jones matrices of the four stages to obtain

\[ v_L = J_L J_{RX,oc} J_{TX,oc} J_g v_g = J_{RX} J_{TX} v_g. \] (6.14)

If the medium and materials in the system are reciprocal and the same types of antennas are used for both the transmitter and receiver, the relationship

\[ J_{RX} = c_1 z_L J_{TX}^T \] (6.15)

follows from the reciprocity of the system.

Based on the system Jones matrices, the polarimetric figures of merit, cross polarization discrimination (XPD) and cross polarization isolation (XPI), are [61]

\[ \text{XPD}_u = \frac{|J_{RX,11}|^2}{|J_{RX,21}|^2}, \quad \text{XPD}_v = \frac{|J_{RX,22}|^2}{|J_{RX,12}|^2} \] (6.16)

\[ \text{XPI}_1 = \frac{|J_{TX,11}|^2}{|J_{TX,21}|^2}, \quad \text{XPI}_2 = \frac{|J_{TX,22}|^2}{|J_{TX,12}|^2}. \] (6.17)

The axial ratio (AR) can be calculated from XPI by [62]

\[ AR = \frac{\sqrt{\text{XPI}} + 1}{\sqrt{\text{XPI}} - 1}. \] (6.18)

This analysis provides a complete Jones matrix model for the radiation and polarization properties of the system. The next step is to derive the network parameters from equivalent circuit models of the transmitting and receiving antennas.

### 6.2.3 Equivalent Circuit Model

**Dual-polarized antenna equivalent circuit models**

Antenna performance parameters can be linked to the physical operating mechanisms of the structure using an equivalent circuit model. Whereas a dipole is commonly modeled as a series resonant circuit, a microstrip antenna is better represented by a parallel resonant circuit [63, 64].
By perturbing the patch shape, a pair of orthogonal modes can be produced simultaneously by a single feed point, with the phase difference between the modes tuned to realize CP radiation. In [65], the circuit model of a CP MSA is built by cascading two parallel resonance circuits which represent the two degenerate modes. For a two-port, dual-resonant patch of this type, dependent generators [64] and $\pi$ or T-shape equivalent circuits [66] allow the antenna impedance behavior to be accurately represented.

Figure 6.5: Circuit model for an intrinsically dual circularly polarized MSA: (a) Rx, (b) Tx.
To enable design optimization and provide insight into the physical mechanisms that govern antenna performance, a dual CP MSA equivalent circuit must represent both the impedance and the polarimetric properties of the antenna. To achieve this, following [65] two parallel resonant circuits are used to represent the orthogonal radiating modes. A key challenge is that while cascading the two resonant circuits works fine for single port CP MSAs, this simplistic model cannot support multi-port operation. Early work in [51] used a circuit model to analyze a dual CP MSA, but the predicted axial ratio (AR) values did not match measured results because the two resonant circuits were cascaded directly and the second port was simply modeled as a load coupled with the resonant circuits through two transformers. By using transformers to model signal coupling into radiating modes [67, 68], each mode operates independently and the ports can be assigned adjustable coupling level to each mode. In this way, the model can accommodate different antenna configurations such as additional ports and a range of feed probe location. The complete proposed circuit models for dual CP MSAs are shown in Fig. 6.5 for the Tx and Rx cases.

In the model, each port connects in series with two transformers independently coupling into two parallel RLC circuits representing the orthogonal horizontally and vertically radiating modes. Since the second port position is always mirrored from the first one to create the CP in opposite sense, the transformer on the horizontal branch coupling with port 1 has a current direction opposing that of the other port. By the reciprocity result of Section 6.2.2, it is sufficient to calculate only the receiving Jones matrix. As illustrated in Fig. 6.5(a) for the Rx scenario, with the assumption that the orthogonal modes radiate perfect linearly polarized waves without ohmic loss, the short circuit currents induced by an incident wave add by Norton’s theorem are

\[
\begin{bmatrix}
i_{Sh} \\
i_{Sy}
\end{bmatrix} = \frac{c_1}{V_0} \begin{bmatrix}
E_h(-\hat{k}^{inc} r) & 0 \\
0 & E_v(-\hat{k}^{inc} r)
\end{bmatrix} \begin{bmatrix}
E^{inc}_h \\
E^{inc}_v
\end{bmatrix} = J_{sc}E^{inc}_{LP},
\]

(6.19)
where $E_p(-\hat{k}_{\text{inc}}r)$ is the field radiated by the $\hat{p}$ polarization mode in the incident wave direction when excited with the input voltage $V_0$. The power density radiated by the $\hat{p}$ mode is

$$S_{\text{rad}}^p(r) = \frac{P_{\text{rad}}^p G_p(\hat{r})}{4\pi r^2} = \frac{|V_0|^2 G_p(\hat{r})}{8R_p \pi r^2}$$

(6.20)

where $G_p(\hat{r})$ is the partial gain of the $\hat{p}$ mode in the $\hat{r}$ direction. Assuming that the input voltage $V_0$ is real, combining (6.20) and (6.21) leads to

$$|E_p(r)| = \frac{V_0}{\sqrt{R_p}} \sqrt{\frac{G_p(\hat{r}) \eta}{4\pi r^2}}.$$  

(6.22)

Since the two radiated fields have the same phase center, both $E_h(\vec{r})$ and $E_v(\vec{r})$ have a phase term $e^{-jk_r}$ with the initial phase assumed to be zero. Substituting (6.22) into (6.19) gives

$$i_S = c \begin{bmatrix} G_h(-\hat{k}_{\text{inc}}) \sqrt{R_h} & 0 \\ 0 & G_v(-\hat{k}_{\text{inc}}) \sqrt{R_v} \end{bmatrix} \begin{bmatrix} E_{\text{inc}}^L \hfill \text{LHCP} \\ E_{\text{inc}}^R \hfill \text{RHCP} \end{bmatrix}.$$  

(6.23)

where $c = j\lambda / \sqrt{\pi \eta}$.

To transform an incident CP wave into its LP representation, we use

$$\begin{bmatrix} E_{\text{inc}}^L \\ E_{\text{inc}}^R \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & 1 \\ -j & j \end{bmatrix} \begin{bmatrix} E_{\text{inc}}^L \hfill \text{LHCP} \\ E_{\text{inc}}^R \hfill \text{RHCP} \end{bmatrix}.$$  

(6.24)

The output load voltages can be found by summing the induced voltages from each current source. The contribution from the $i_{Sh}$ source can be obtained from

$$v_h = Z_h (i_{Sh} + N_h v_{L1}/Z_L - N_h v_{L2}/Z_L),$$

(6.25)

$$v_v = -N_v Z_v (v_{L1} + v_{L2}),$$

(6.26)

$$v_{L1} = -N_h v_h + N_v v_v, \quad v_{L2} = N_h v_h + N_v v_v.$$  

(6.27)
Solving for the output voltages leads to

\[ v_{L1} = \frac{-N_h Z_h Z_L}{2N_h^2 Z_h + Z_L} i_{Sh} \]
\[ v_{L2} = \frac{N_h Z_h Z_L}{2N_h^2 Z_h + Z_L} i_{Sh}. \]  
(6.28)

The voltage responses caused by \( i_{SV} \) can be found similarly with different expressions of \( v_h \) and \( v_v \) as

\[ v_h = \frac{N_h Z_h}{Z_L} (v_{L1} - v_{L2}) \]  
(6.29)
\[ v_v = Z_v \left( i_{SV} - \frac{N_v v_{L1}}{Z_L} - \frac{N_v v_{L2}}{Z_L} \right). \]
(6.30)

Arranging both responses into a matrix form leads to

\[
\begin{bmatrix}
  v_{L1} \\
  v_{L2}
\end{bmatrix} =
\begin{bmatrix}
  \frac{-N_h Z_h Z_L}{2N_h^2 Z_h + Z_L} & \frac{N_h Z_v Z_L}{2N_h^2 Z_v + Z_L} \\
  \frac{N_h Z_h Z_L}{2N_h^2 Z_h + Z_L} & \frac{N_h Z_v Z_L}{2N_h^2 Z_v + Z_L}
\end{bmatrix}
\begin{bmatrix}
  i_{sh} \\
  i_{SV}
\end{bmatrix} = \mathbf{J}_L \mathbf{i_S}. \]   
(6.31)

Combining (6.23), (6.24), and (6.31) leads to

\[
v_L = c \mathbf{J}_L \begin{bmatrix}
  G_h (\frac{-k_{\text{inc}}}{\sqrt{2R_h}}) & G_v (\frac{-k_{\text{inc}}}{\sqrt{2R_v}}) \\
  -jG_v (\frac{-k_{\text{inc}}}{\sqrt{2R_v}}) & jG_v (\frac{-k_{\text{inc}}}{\sqrt{2R_v}})
\end{bmatrix}
\begin{bmatrix}
  E_{\text{inc}}^{LHCP} \\
  E_{\text{inc}}^{RHC}\n\end{bmatrix} = \mathbf{J}_R \mathbf{E}_{\text{inc}}^{CP}. \]
(6.32)

The Z-parameters of the antenna are

\[
\mathbf{Z} = \begin{bmatrix}
  N_h^2 Z_h + N_v^2 Z_v & -N_h^2 Z_h + N_v^2 Z_v \\
  -N_h^2 Z_h + N_v^2 Z_v & N_h^2 Z_h + N_v^2 Z_v
\end{bmatrix}. \]  
(6.33)

The corresponding S-parameters can be found from the impedance matrix according to [69]

\[
\mathbf{S} = \begin{bmatrix}
  \frac{(Z_{11} - Z_0)(Z_{22} + Z_0) - Z_{12}Z_{21}}{\Delta Z} & \frac{2Z_{12}Z_0}{\Delta Z} \\
  \frac{2Z_{12}Z_0}{\Delta Z} & \frac{(Z_{11} + Z_0)(Z_{22} - Z_0) - Z_{12}Z_{21}}{\Delta Z}
\end{bmatrix}, \]
(6.34)

where \( \Delta Z = (Z_{11} + Z_0)(Z_{22} + Z_0) - Z_{12}Z_{21} \).
Figure 6.6: Simulated S-parameters based on full wave analysis and the circuit model for an intrinsically dual CP MSA. The circuit model based results agree well with the full wave simulation results.

**Circuit model validation**

To validate the dual-polarization MSA circuit model, a rectangular patch antenna based on 5 mil Rogers RT/duroid 5870 was optimized for dual CP operation at 13.15 GHz using the finite element method (HFSS, Ansys, Inc.) with the configuration detailed in Fig. 6.6. Because the two orthogonal modes need to be excited equally for CP, all turn ratios of transformers are set to $1/\sqrt{2}$ for calculation convenience in the circuit model. A series inductor at each port was added to compensate the parasitic effect of the feeding pins. The circuit model component values were then selected to match the numerically simulated Z-parameters. The resulting agreements over frequency between the actual antenna and the circuit model as shown in Figs. 6.6 and 6.7 demonstrate the effectiveness of the circuit model with respect to the MSA input impedances.

The next step is to validate the radiated fields predicted by the circuit model. Assuming that the orthogonal modes radiate in an identical gain pattern, the system Jones matrix can be obtained from the circuit model and used to predict the XPD/XPI and AR. The AR values calculated from the circuit model are compared to a full wave simulation of the actual antenna in Fig. 6.8. Agreement is good over the full bandwidth range shown. These comparisons of the circuit model
to full-wave simulations show that the circuit model closely matches the actual antenna in terms of its impedance and radiation properties.

### 6.2.4 Fundamental Performance Bounds for Intrinsically Dual CP MSAs

It is known that there exists a tradeoff between dual CP performance in terms of AR and impedance match including effects of coupling and reflection [51], but no quantitative analysis of this effect appears to be available in the literature. By adjusting the resonance frequency separation between the two orthogonal modes in the proposed circuit model, the tradeoff can be illustrated clearly as Fig. 6.9. In this section, the feasible region for maximum power transfer at different cross polarization performance levels is derived. A relationship between the optimized 3 dB AR bandwidth and -10 dB mode impedance bandwidth of an intrinsically dual CP MSA is derived for two different loading conditions for the non-driven port.
Figure 6.8: Simulated axial ratio based on full wave analysis and the circuit model. Field estimation based on Jones matrix and circuit model matches well with full wave simulation. Bandwidth can be enhanced by using a wider bandwidth antenna.

Figure 6.9: S-parameters and XPD as a function of frequency separation for an ideal $Q = 120$ MSA with the center frequency at 12.5 GHz. XPD is maximized at the optimal frequency separation, but matching and isolation are poor.
Feasible region for XPI and impedance mismatch factor

Given the relationship between radiation pattern orthogonality and the impedance matrix of a multport antenna [56], one would expect that perfect polarization isolation implies high isolation between ports of a dual-CP antenna. Counter to this intuition from basic theorems of array antennas, it has been observed that it is actually challenging to obtain both high isolation and good AR simultaneously for intrinsically dual-CP MSAs [51]. The goal here is to quantify using the circuit model from the previous section the apparent contradiction that the polarization isolation can be good for a dual-CP MSA while at the same time the port isolation is poor.

To simplify the analysis, we assume the source impedance $Z_L$ is purely real and equal to $R_L$, $R_h = R_v = R_{rad}$, $N_h = N_v = 1/\sqrt{2}$, and the orthogonal modes radiate in an identical gain pattern.

To incorporate both reflections from individual ports and coupling between ports into one figure of merit, we quantify the two-port impedance matching quality using the impedance mismatch factor (IMF) [70]

$$\text{IMF} = 1 - |S_{11}|^2 - |S_{21}|^2.$$  \hspace{1cm} (6.35)

When the impedance mismatch factor is minimized, maximum power transfer to the two-port antenna is achieved.

AR is directly related to the ratio of the two orthogonal LP fields through [71]

$$\text{AR} = \left( \frac{A^2 + 1 + [A^4 + 1 + 2A^2 \cos(2\phi)]^{\frac{1}{2}}}{A^2 + 1 - [A^4 + 1 + 2A^2 \cos(2\phi)]^{\frac{1}{2}}} \right)^{\frac{1}{2}} \approx \sqrt{(A_e^2 + 0.15^2 \phi_e^2)} \text{ (dB)}, \hspace{1cm} (6.36)$$

where $A$ and $A_e$ represent the amplitude ratio on linear and dB scales, respectively, $\phi$ is the phase of the ratio, and $\phi_e$ is the phase error between $\phi$ and the values $\pm90^\circ$ required for perfect CP. For the Tx circuit model as shown in Fig. 6.5(b), the field ratio is the voltage ratio $v_h/v_v$. This reduces the problem of determining the feasible region for XPI and the impedance mismatch factor to the maximum power transfer condition for a given AR.

Using the reciprocity property of the Jones matrix (6.15) and the load Jones matrix (6.31) from the Rx circuit model analysis in Section 6.2.3, the voltage ratio between the two orthogonal
modes $v_h/v_v$ by exciting port 2 is

$$\frac{v_h}{v_v} = \frac{N_h Z_h (2N_v^2 Z_v + R_L)}{N_v Z_v (2N_h^2 Z_h + R_L)} = \frac{Z_h (Z_v + R_L)}{Z_v (Z_h + R_L)}. \quad (6.38)$$

The ratio can be simplified by expressing the input impedance of mode $m$ (h or v) in the form $Z_m = R_{rad}/(1 + jx_m)$ due to the parallel circuit characteristic

$$\frac{v_h}{v_v} = \frac{R_{rad} + (1 + jx_v)R_L}{R_{rad} + (1 + jx_h)R_L}. \quad (6.39)$$

To find the maximum power transfer condition for an arbitrary AR value, we let the voltage ratio be an arbitrary complex number $a + jb$. The voltage ratio equation can be expanded to obtain $x_m$ in terms of $R_{rad}$, $a$, and $b$,

$$x_m = c_m(R_{rad} + R_L)/R_L, \quad (6.40)$$

where $c_m = (a - 1)/b$ and $(a^2 + b^2 - a)/b$ for horizontal and vertical modes, respectively. The radiated power for each mode can be expressed as

$$P_{rad,m} = \frac{|v_m|^2}{2R_{rad}} = \frac{|v_s|^2}{2R_{rad} \left| \frac{2N_m^2 Z_m + R_L}{N_m Z_m} \right|^2} = \frac{|v_s|^2}{4 |R_{rad} + R_L (1 + jx_m)|^2 / R_{rad}}. \quad (6.41)$$

Substituting (6.40) into (6.41) to simplify the denominator leads to

$$4(1 + c_m^2) \left( \frac{R_{rad} + R_L}{R_{rad}} \right)^2. \quad (6.42)$$

By setting the first derivative of the denominator expression with respect to $R_{rad}$ equal to zero, the maximum power transfer condition for both modes is

$$R_{rad} = R_L, \quad (6.43)$$

which is as expected from the maximum power transfer theorem of network theory.
Under the maximum power transfer condition, the peak radiated power is

\[ P_{\text{rad},m} = \frac{|v_s|^2}{4R_L(4 + x_m^2)}, \quad (6.44) \]

and the voltage ratio simplifies to

\[ \frac{v_h}{v_v} = \frac{2 + jx_v}{2 + jx_h} = Ae^{j\phi}. \quad (6.45) \]

The reactances \( x_h \) and \( x_v \) can be expressed in terms of \( A \) and \( \phi \) as

\[ x_h = 2(cot \phi - \csc \phi / A) \quad (6.46) \]
\[ x_v = 2(A \csc \phi - cot \phi). \quad (6.47) \]

The total radiated power is

\[ P_{\text{rad}} = \frac{|v_s|^2}{16RL} \left[ \frac{1}{1 + (\cot \phi - \csc \phi / A)^2} \right. \\
\left. + \frac{1}{1 + (A \csc \phi - \cot \phi)^2} \right]. \quad (6.48) \]

By finding the zero of the derivative of (6.48) with respect to \( A \) as

\[ \frac{\partial P_{\text{rad}}}{\partial A} = \frac{|v_s|^2}{16RL} \left\{ \frac{2(cot \phi - \csc \phi / A) \csc \phi A^{-2}}{[1 + (\cot \phi - \csc \phi / A)^2]^2} \right. \\
\left. + \frac{2(cot \phi - A \csc \phi) \csc \phi}{[1 + (\cot \phi - A \csc \phi)^2]^2} \right\} = 0, \quad (6.49) \]

the maximum power transfer condition for a given \( \phi \) value is

\[ A = 1. \quad (6.50) \]

Because the constant AR (dB) contour is approximately a circle in the \((A_c, 0.15\phi_c)\) coordinate, as illustrated by (6.37), the maximum power transfer condition for a given \( \phi \) value is not necessarily valid for a given AR. The power transfer with zero amplitude error may not be larger than that of cases with some amplitude error but smaller phase error for the same AR.
To prove that the maximum power transfer condition (6.50) holds for all AR values, the derivative of (6.48) with respect to $\phi$ at $A = 1$ was found to be less than zero as

$$\frac{\partial P_{\text{rad}}}{\partial \phi} = \frac{|v_s|^2}{8R_L} \left\{ - \frac{2(\cot \phi - \csc \phi)^2 \csc \phi}{1 + (\cot \phi - \csc \phi)^2} \right\} < 0. \quad (6.51)$$

This implies that transferred power drops as polarization becomes more circular with increasing $\phi$ when $A = 1$. It can be concluded that for a given AR value the maximum radiated power still occurs at $A = 1$, because (6.49) and (6.51) mean the power transfer of an arbitrary $(A, \phi)$ case is smaller than that of the same $\phi$ value with $A = 1$ case which is further smaller than the case with $A = 1$ producing the same AR with the original arbitrary case, i.e.

$$P_{\text{rad}}(A_x, \phi_x) < P_{\text{rad}}(1, \phi_x) < P_{\text{rad}}(1, \phi_y), \quad (6.52)$$

where $x$ indicates an arbitrary coordinate and $(1, \phi_y)$ has the same AR with $(A_x, \phi_x)$.

This shows that the conditions for maximum power transfer for a given AR value are that the mode radiation resistance equals the load resistance and the magnitude of the mode voltage ratio equals one. The first condition is an extension of the maximum power transfer theorem, and the second condition indicates that the maximum power transfer occurs at a frequency that is approximately in the center of the resonances of the two orthogonal modes tuned for a given AR value.

The condition (6.50) for maximum power transfer further simplifies the expressions for $x_m$ (6.46) (6.47) and AR (6.36) to

$$x_h = -x_v = 2(\cot \phi - \csc \phi) \quad (6.53)$$

$$\text{AR} = \cot \frac{\phi}{2}. \quad (6.54)$$

The S-parameters can be calculated from the impedance values (6.43) and (6.53). Using this model, the feasible region for impedance mismatch factor and XPI for a lossless dual CP MSA is shown in Fig. 6.10. To validate the feasible region, performance figures of merit for several actual intrinsically dual CP antennas are indicated with markers.
Figure 6.10: Feasible region for XPI and impedance mismatch factor $\eta = 1 - |S_{11}|^2 - |S_{21}|^2$ for a dual-CP MSA antenna. Actual dual CP antennas are indicated with markers: ◇ simple probe fed rectangular patch, × optimized probe fed rectangular patch, • fork fed corner-cut square patch [1], △ thin diagonally slotted circular patch, * thick diagonally slotted circular patch [2], + T-coupled corner-cut square patch [3], ▽ dual LP patch with a 90° hybrid (i.e., a non-intrinsically dual-CP MSA). For a good intrinsically dual CP antenna, reflection and coupling are close to -6 dB as attained by the best design (∗).

For the intrinsically dual-CP designs shown in Fig. 6.10, none lie outside the feasible region, as expected. The best family of designs is those closest to the boundary of the feasible region. A non-intrinsically dual-CP antenna consisting of a dual LP MSA fed by a quadrature hybrid is also shown, and has performance well outside the feasible region for intrinsically dual-CP antennas. This clearly reveals the tradeoff between the better XPI and matching performance of the quadrature fed design and the much smaller size and lower loss of the intrinsically dual-CP antennas.

The poor port isolation of the intrinsically dual-CP designs is just as detrimental for SatCom applications that require high sensitivity as the higher loss of the quadrature fed patch. As demonstrated in [2], a good dual CP antenna with poor reflection and coupling cannot achieve good SNR performance when embedded in a receiving system, even with perfectly designed low noise amplifiers (LNAs). It can be seen that the design problem becomes one of reducing the port coupling
of the intrinsically dual-CP antennas. After applying the circuit model to bandwidth optimization and radiation efficiency measurement, the coupling issue will be dealt with in Section 6.2.5.

**AR bandwidth limits**

The AR bandwidth of a dual LP MSA with a $90^\circ$ hybrid is mainly determined by the phase and magnitude difference between the two outputs of the hybrid. The bandwidth of the hybrid is normally larger than impedance bandwidth of a probe-fed single layer MSA. Because an intrinsically dual CP MSA obtains a $90^\circ$ phase shift by tuning two orthogonal modes resonant at two different frequencies, the phase and magnitude differences change rapidly away from the center frequency due to the sharp slope of the modal reactance, also leading to narrow AR bandwidth. In this section, the relationship between the impedance bandwidth of the dominant mode of an MSA and its AR bandwidth when operating as a dual-CP antenna is investigated for the non-driven port either open circuit loaded (OCL) or $50\Omega$ loaded.

To operate in CP for the $50\Omega$ loaded case, the voltage ratio (6.45) must be $\pm j$. Assuming the ratio to be $j$ results in $x_h = -x_v = -2$. Comparing the assumption $Z_m = R_m/(1 + jx_m)$ with the input impedance expression with quality number of a parallel resonant circuit [69] gives

$$x_m = 2Q\Delta f/f_0,$$

where $Q$ is the quality number of the MSA excited with single LP mode. The modal frequency shift required for CP operation is

$$\Delta f_h = -\frac{f_c}{Q-1}, \quad \Delta f_v = \frac{f_c}{Q+1},$$

where $f_c$ is the center frequency where perfect dual CP operates. If the AR decreases to 3 dB when the frequency is shifted $\Delta f_{AR}$ from the center frequency, $x_m$ at that frequency becomes

$$x_m = 2Q\frac{\Delta f_m + \Delta f_{AR}}{f_c - \Delta f_m}.$$
Substitute (6.56) into (6.57) leading to

\[
x_h = -2 + B_{AR}(Q - 1) \quad x_v = 2 + B_{AR}(Q + 1),
\]

(6.58)

where \( B_{AR} = 2\Delta f_{AR}/f_c \). The relationship between the 3 dB AR bandwidth of an intrinsically dual CP MSA and the mode quality factor \( Q \) can be obtained by combining (6.36), (6.45), and (6.57). Using a 10 dB matching threshold, the impedance bandwidth is [72]

\[
B_{-10\text{dB}} = \frac{\Delta f_{-10\text{dB}}}{f_c} = \frac{VSWR - 1}{Q\sqrt{VSWR}} = \frac{2}{3Q}.
\]

(6.59)

This relationship between the 3 dB AR bandwidth of an intrinsically dual CP MSA and the impedance bandwidth of the MSA in LP mode is shown in Fig. 6.11.

The AR bandwidth for one of the CP ports is sensitive to the loading condition on the second port. The above result was derived for a 50\( \Omega \) load. To study the dependence on loading condition, we will change from a match to the extreme case of an OCL. The voltage ratio with maximum power radiated becomes

\[
\frac{v_h}{v_v} = \frac{1 + jx_v}{1 + jx_h},
\]

(6.60)

and the condition for CP is

\[
\Delta f_h = -\frac{f_c}{2Q - 1} \quad \Delta f_v = \frac{f_c}{2Q + 1}.
\]

(6.61)

Fig. 6.11 illustrates that the 3 dB AR bandwidth varies almost linearly with -10 dB impedance bandwidth, with ratios of 1.03 and 0.53 for 50\( \Omega \) loading and OCL at the nondriven port, respectively. The AR bandwidth for OCL agrees with [73, 74], in which it is concluded that the 3 dB AR bandwidth is 35% of the mode resonant frequency difference. By sacrificing the quality of the impedance match, we see that an intrinsically dual CP MSA can achieve twice the AR bandwidth of an MSA with an OCL at the non-driven port.
6.2.5 Applications of the Circuit Model

Multiport Wheeler cap method for dual-CP antennas

The Wheeler cap method [75] provides an accurate and convenient measurement of radiation efficiency for small and planar antennas [6] and connected arrays [76]. This method relies on choosing a reasonable circuit model for the antenna under test. While a series resonant circuit model is valid for wire antennas, a parallel resonant circuit shows better efficiency accuracy for a single port LP MSA [63]. In [77], a physically meaningful equivalent circuit modeling method based on Characteristic Modes (CM) was proposed, which generates impedance and field results that agree well with full wave simulations across a wide bandwidth. The CM formulation assumes that the antenna is lossless, however, so the Wheeler cap method with the CM-based circuit model requires a complex calculation based on a modified CM formulation for the lossy case. This calculation was simplified in [65] by using multiple simple resonant circuits with each linked to different modes. The remaining open problem is that we know of no literature on the Wheeler cap method that deals with multiple port antennas.
The Wheeler cap method has not been applied to multiport antennas primarily due to a lack of a good circuit model for the multiport case. The circuit model proposed in this paper solves this problem for dual-CP MSAs. Since the circuit model is accurate over a wide bandwidth, as shown in previous sections, when it is used with the Wheeler cap method, the extracted radiation efficiency is accurate over the full operating bandwidth of these two-port antennas.

The Wheeler cap method was applied to the dual CP MSA used as validation in Section 6.2.3. The total dissipation can be separated into radiation and loss resistors in parallel. Adding a conductive cap eliminates radiation, so that only loss resistors are left in the circuit model. Due to the added capacitance from capping the MSA, the resonant frequencies are shifted slightly higher [65], so the capacitor values should be adjusted accordingly while the inductors and the loss resistors are kept the same for the capped case. The impedance obtained from the circuit model matches closely with the full wave simulation as shown in Fig. 6.12. The resulting circuit model with loss separated from radiation is shown in Fig. 6.13.
With the circuit model shown in Fig. 6.13, the radiation efficiency is

\[ \eta_{\text{rad}} = \frac{|v_h|^2 G_{hr} + |v_v|^2 G_{vr}}{|v_h|^2 G_h + |v_v|^2 G_v}, \]

(6.62)

where \( v_h \) and \( v_v \) can be obtained from basic circuit analysis and are affected by the loading condition at the non-driven port. The simulated radiation efficiency using the finite element method and the Wheeler cap method for the intrinsically dual CP antenna of Fig. 6.6 with the non-driven port 50\( \Omega \) loaded are shown in Fig. 6.14. The error is only 3% between the two resonant frequencies 12.98 GHz and 13.32 GHz.

It can be expected that the efficiency slope from the Wheeler cap method would be more accurate with the circuit model including more modes. Therefore the Wheeler cap method can be still valid and accurate for a multi-port antenna by using a correct circuit model based on the method proposed in this section.
Figure 6.14: Simulated radiation efficiency with the non-driven port 50Ω loaded as a function of frequency based on full wave analysis and the Wheeler cap method.

**High isolation dual CP antenna**

A figure of merit that includes impacts of both mismatch and cross polarization on system performance is the system temperature over efficiency $T_{sys}/(\eta_{ant}\eta_{pol})$, explained in Section 2.1. With typical parameter settings the same as Section 6.1.3, the point with best inverse sensitivity occurs between the best CP quality and the perfectly matched LP case as shown in Fig. 6.9 and 6.15. The figure also shows that the higher the LNA quality, the better the optimal system performance shifts towards the highest quality CP case.

Due to the adverse effect of coupling on receiving performance [2], a decoupling network is required to make an intrinsically dual CP MSA useful for SatCom applications. In [78], a general decoupling method is proposed, which utilize a pair of transmission lines at the antenna ports to tune mutual admittance pure imaginary and then bridge a reactance between the two branches to cancel the imaginary admittance. However, since that method only considers decoupling, directly applying it to an intrinsically dual CP MSA [3] would result in deformed patterns with poor AR, because the parallel reactance bridge perturbs the input currents to the antenna ports. In other words, port coupling is improved, but the CP modes mix and AR is degraded.
This can be seen for the probe-fed intrinsically dual CP MSA from Section 6.2.3. The required decoupling network is a reactance in series with the antenna as shown in Fig. 6.16. The decoupling network is particularly simple, as the real part of $Z_{21}$ is zero due to the symmetry of the feeding points. If the decoupling network is designed for a non-driven port 50 $\Omega$ loaded dual CP MSA, the AR at boresight deteriorates severely in spite of a nearly perfect impedance mismatch factor, because the decoupled AR is determined by the open circuit loaded (OCL) case at the decoupling frequency as shown in Figs. 6.17 and 6.18.
Figure 6.17: S-parameters for the MSA with the decoupling and matching networks using the method without consideration of AR.

Figure 6.18: Axial ratio comparison of different stages in the decoupled and matched MSA with the decoupling and matching networks using the method without consideration of AR.
A better decoupling scheme for an intrinsically dual CP antenna should be achieving good
CP at the driven port and then applying the decoupling and matching networks. The antenna in
Fig. 6.6 can be re-optimized by tuning the two resonant frequencies of the circuit model in Fig.
6.13 for good dual CP with the non-driven port OCL. To verify the validity of optimizing an MSA
based on the circuit model, a full wave simulation of the MSA with proportionally adjusted patch
size according to the tuned resonant frequencies was conducted. The comparisons of S-parameters,
Z-parameters, and AR illustrated in Fig. 6.19, 6.20, and 6.21, respectively, demonstrate the tuning
method based the circuit model is accurate.

Based on the optimized OCL dual CP MSA and the same decoupling and matching scheme
in Fig. 6.16, not only small coupling and low return loss can be achieved as shown in Fig. 6.22,
but also the AR maintains the same perfect value at the center frequency and has the bandwidth
enlarged through each stage as shown in Fig. 6.23. The system performance obtained for a cor-
rectly decoupled intrinsically dual CP antenna is no longer limited by the AR bandwidth, but by
the isolation bandwidth. Since the decoupling and matching networks in this solution are small
and simple, the loss is much lower than that of the hybrid required for a non-intrinsically dual-CP
MSA.
Figure 6.20: Z-parameters for the optimized MSA without networks based on the circuit mode and full wave simulation.

Figure 6.21: Axial ratios for the optimized MSA without networks based on the circuit mode and full wave simulation.
Figure 6.22: S-parameters for the optimized MSA with the decoupling and matching networks using the proposed method.

Figure 6.23: Axial ratio comparison of different stages in the proposed high performance dual CP MSA using the proposed method.
The same scheme applies to other feeding methods for MSAs which may have zero mutual admittance. In those cases, good CP should be achieved at the driven port with the non-driven port short circuit loaded (SCL), and an effective decoupling network becomes a shunt reactance. The networks can be realized using either lumped components or distributed transmission line. In the case of zero mutual admittance, an effective decoupling network becomes a shunt reactance which can be conveniently realized as a distributed structure. Designs in [79–81] provide possible stripline embodiments of the proposed decoupling method.

6.2.6 Summary

The proposed equivalent circuit model and Jones matrix analysis for intrinsically dual CP MSAs solves several open problems in antenna theory. The model provides accurate estimates of impedances and S-parameters, as well as field parameters such as axial ratio. The feasible region for XPI and impedance mismatch factor is found for dual CP antennas, which explains the counterintuitive tradeoff between matching and polarization quality observed with these designs. A relationship between impedance bandwidth and axial ratio is also derived, and the role of loading condition on the non-driven port is revealed. The circuit model allows the Wheeler cap method for radiation efficiency measurement to be applied for the first time to multi-port antennas. An effective decoupling scheme to overcome the port isolation issue is proposed and demonstrated, leading to a high isolation, good match, and wide AR bandwidth dual CP MSA for use in applications such as satellite communications where high sensitivity, good noise performance, and high polarization quality are required.
CHAPTER 7. CONCLUSION AND FUTURE WORK

7.1 Conclusion

For high efficiency passive array feeds, we have successfully designed multiple high efficiency array feeds for Ku band satellite communications. Based on design guidelines for high efficiency microstrip antennas, a $2 \times 2$ MSA array achieves the highest reported radiation efficiency of 93% based on material compatible with an FR4 fabrication process at Ku band. A stacked shorted annular patch antenna was designed and optimized as the feed for a terrestrial reflector in Ku band satellite communications. Without distribution networks and dielectric substrates, the antenna loss is lower than that of microstrip antenna feeds, and spurious radiation from transmission lines and surface waves is eliminated completely, leading to high radiation efficiency. The overall SNR performance is only 0.7 dB lower than a commercial horn feed while lighter and simpler. A PTFE material based SIW-fed hexagonal array feed was designed, which can match the performance of a commercial horn feed. It has the best simulated G/T performance reported to date among non-horn type feeds.

High performance MSA array feeds are achievable for dual polarization and multiband applications. A planar Tx/Rx dual band slot patch passive array feed antenna with high isolation and good satellite communication capability was designed and demonstrated. The array feed functionally equivalent to a standard horn-type feed antenna plus OMT provides a promising way to minimize two-way link feed size and demonstrates that a planar antenna is a feasible and effective solution for low cost, high performance ground terminals. Dual linearly and circularly polarized MSA array feeds were also demonstrated. The first reported planar array feed design for collocated Ku/RB dual polarization satellite communications was also presented. With the proposed array feed on ground terminals, a new RB satellite can be launched to share the same orbital slot with an existing Ku band satellite. Simulation results have demonstrated that this planar array feed
is promising for dual band receiving with good isolation between ports and high efficiencies for all four ports.

One dimensional limited scan range beam-steering was demonstrated experimentally by the $4 \times 2$ ESAF with VGAs. Acquiring and tracking functions were realized with the power detector based feedback system. The ESAF does not require phase shifters and uses a conventional parabolic reflector as the main radiator, leading to a low cost and efficient solution for limited scan range beam-steering. To improve ESAF performance, high efficiency MSA element can be designed based on the guideline proposed in Section 2.2 and the transitions between antenna elements and LNAs should avoid using a long transmission line. These considerations were added into the $4 \times 4$ ESAF. Two dimensional beam-steering capability and better performance are expected and will be demonstrated experimentally in the future.

Phased-tile arrays can be a valuable solution for SatCom applications requiring beam steering only in a limited scan range. A high efficiency $4 \times 4$ MSA array tile can provide a solid foundation for future phased-tile array design. To achieve high performance required by SatCom within restricted tile size limits, the layer stack-up configuration is carefully discussed with considerations on manufacturability, cost, and efficiencies of MSA, distribution networks, and transitions. A method to enhance bandwidth and radiation efficiency of a rectangular Tx/Rx dual band MSA was proposed. Distribution networks consisting of two CPWG and SIW stages for Tx and Rx to minimize fabrication cost and maintain good performance are described. Limited scan range beamsteering can be implemented cost-effectively by using a tile array with each tile consisting of a passive network fed subarray. To maintain a monotonic sidelobe level, the same element distance through the whole array panel must be maintained. The sidelobes of an uniformly excited array exceeds the regulatory pattern mask, requiring edge element tapers. With these design aspects, a viable tiled array for Satcom applications can be realized.

Low envelope correlation coefficient in a dual polarized receiver does not necessarily mean low mutual coupling. Port coupling, which can be difficult to minimize for planar dual CP antennas, increases system noise and reduces receiver sensitivity. The impact of coupling on receiver performance is minimized when the LNA noise parameters are such that no reverse noise wave is emitted, and this case provides a lower bound on the sensitivity degradation due to coupling. If the LNAs are designed optimally, the coupling can be -10 dB or poorer with minimal impact.
on sensitivity. To overcome the port isolation challenge with intrinsically dual-CP MSA designs, we proposed equivalent circuit model and Jones matrix analysis for intrinsically dual CP MSAs, which can solve several open problems in antenna theory. The model provides accurate estimates of impedances and S-parameters, as well as field parameters such as axial ratio. The feasible region for XPI and impedance mismatch factor is found for dual CP antennas, which explains the counterintuitive tradeoff between matching and polarization quality observed with these designs. A relationship between impedance bandwidth and axial ratio is also derived, and the role of loading condition on the non-driven port is revealed. The circuit model allows the Wheeler cap method for radiation efficiency measurement to be applied for the first time to multi-port antennas. An effective decoupling scheme to overcome the port isolation issue is proposed and demonstrated, leading to a high isolation, good match, and wide AR bandwidth dual CP MSA for use in applications such as satellite communications where high sensitivity, good noise performance, and high polarization quality are required.

### 7.2 Future Work

Planar antenna radiation efficiency and insertion loss of a planar distribution network heavily depend on substrate material. PTFE material has lowest loss, but the fabrication cost is huge especially when a multilayer PCB is required, compared with a conventional FR-4 fabrication process. To obtain super high efficiency multi-functional planar array feeds with acceptable fabrication cost, more effort is needed to collaborate closely with laminate material manufacturers and PCB fabricators. Another possibility is to eliminate dielectric substrate completely like the SSAP antenna and develop cost-effective antenna feeding techniques applicable to large-scale arrays. To further increase the aperture and spillover efficiencies of a parabolic reflector illuminated by a planar array feed, a patch array feed needs more elements with smaller distance so that the Airy pattern can be matched better (from the receiving perspective) and the illuminating pattern is more like a rectangular shape.

A parabolic reflector fed by an ESAF provides a low cost solution for limited scan range electronic beam steering, because no phase shifters are needed. If increasing the array feed size, the beam can be steered further away from boresight, but the phase distortion would become a serious problem. Co-optimizing array feed configuration with reflector shape could widen the
beam steering range. For the tile array direction, development of a passive subarray with small beam-squinting across the operating bands for both Tx and Rx and high isolation between the two ports is the next objective. In addition, distribution networks that can be tweaked easily to produce different power ratio are highly demanded to facilitate the array edge tapering.

Based on the decoupling and matching schemes for intrinsically dual CP MSAs, a high isolation, good match, and wide AR bandwidth dual CP MSA could be designed on a single layer PCB by realizing those networks in distributed microstrip form. This is very important to promote planar active array feeds in the Satcom industry, because all DBS and most future satellites adopt circular polarization. The single layer compact dual CP MSA element will encourage more engineers to design receiving terminals in a planar profile for applications requiring low profile and light weight such as airliners and single soldiers.
REFERENCES


