Low Cost Ultra-Wideband Millimeter-Wave Phased Arrays

Dissertation

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By

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Abstract

Many high performance wireless applications continue to be integrated onto increasingly small platforms, such as satellites, UAVs, and handheld devices. Low-profile and ultra-wideband antenna arrays have emerged as a potential solution, by allowing many disparate functions to be consolidated into a shared, multi-functional aperture. Simultaneously, the demand for high data rate communications has driven these applications to higher frequencies, with many now exploring the use of the millimeter-wave spectrum. However, existing UWB arrays often utilize complex feed structures which cannot scale to these frequencies. The development of wideband millimeter-wave arrays compatible with low-cost commercial fabrication processes is critical to enabling these small and highly connected platforms.

Tightly Coupled Arrays are one family of low-profile and wideband arrays which have demonstrated superior bandwidth and wide scanning capability. However, the feed design of these arrays is limited to operation below 5 GHz, and suffers from reduced efficiency when scanning. In this work, the feed is modified to improve efficiency by eliminating a Wilkinson power divider, and mitigating the resultant cavity resonances with the application of shorting pins. Likewise, strenuous fabrication requirements are relaxed, allowing fabrication at higher frequencies. This effort is approached initially through the intermediate frequencies in the X-, Ku- and Ka-bands, and is demonstrated to allow the new design to scale up 49 GHz. An $8 \times 8$ prototype
operating over 3.5–18.5 GHz is fabricated and measured to validate the design. Infinite array simulations show VSWR < 2 across this band at broadside, with scanning to ±45° in the H-plane (VSWR < 2.6) and as far 70° in E-plane (VSWR < 2).

At millimeter-wave frequencies, planar co-fabrication of the entire array is critical to achieving repeatable fabrication, by eliminating the need for complex assembly at such small scales. Simultaneously, compatibility with low-cost PCB processes enables the potential for large scale applicability. The limitations of PCB fabrication are discussed, and a planarized balun consisting of only three vias and two metal layers is developed. The design is shown to operate across 24–72 GHz, with VSWR < 2.2 at broadside, and VSWR < 3 for ±45° scans in the E- and H-planes. The design is validated by fabrication of 3 × 3 and 5 × 5 prototype arrays through a commercial PCB vendor. Measurements of the 3 × 3 array show close agreement with simulations.

Finally, we develop a novel measurement technique, necessary for accurate characterization of antennas at or near millimeter-wave frequencies. This comprises a post-processing algorithm which serves to compensate the measured phase response of the array for otherwise undetectable, millimeter-scale movements occurring during sequential measurements. This is developed through a mathematical model, and is verified through measurements at Ku band (18 GHz).
Dedicated to Lindsay.
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# Table of Contents

<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>Abstract</td>
<td>ii</td>
</tr>
<tr>
<td>Dedication</td>
<td>iv</td>
</tr>
<tr>
<td>Acknowledgments</td>
<td>v</td>
</tr>
<tr>
<td>Vita</td>
<td>vi</td>
</tr>
<tr>
<td>List of Tables</td>
<td>x</td>
</tr>
<tr>
<td>List of Figures</td>
<td>xi</td>
</tr>
<tr>
<td>1. Introduction</td>
<td>1</td>
</tr>
<tr>
<td>1.1 Applications of Ultra-Wideband Phased Arrays</td>
<td>1</td>
</tr>
<tr>
<td>1.2 Review of Existing UWB Array Technologies</td>
<td>5</td>
</tr>
<tr>
<td>1.2.1 Tapered Slot</td>
<td>8</td>
</tr>
<tr>
<td>1.2.2 Fragmented Aperture</td>
<td>10</td>
</tr>
<tr>
<td>1.2.3 Connected and Coupled Arrays</td>
<td>12</td>
</tr>
<tr>
<td>1.2.4 Material Loading</td>
<td>15</td>
</tr>
<tr>
<td>1.3 Contemporary Challenges in UWB Phased Arrays</td>
<td>18</td>
</tr>
<tr>
<td>1.4 Contribution and Organization of this Dissertation</td>
<td>21</td>
</tr>
<tr>
<td>2. Design of a Tightly Coupled Array Operating up to 49 GHz</td>
<td>23</td>
</tr>
<tr>
<td>2.1 Initial Design Approach</td>
<td>24</td>
</tr>
<tr>
<td>2.2 Efficiency Improvement and Resonance Mitigation</td>
<td>33</td>
</tr>
<tr>
<td>2.2.1 Split Unit Cell</td>
<td>34</td>
</tr>
<tr>
<td>2.2.2 Surface Waves</td>
<td>38</td>
</tr>
<tr>
<td>2.2.3 Cavity Resonance</td>
<td>48</td>
</tr>
<tr>
<td>2.3 Optimized Performance and Scalability</td>
<td>52</td>
</tr>
<tr>
<td>Section</td>
<td></td>
</tr>
<tr>
<td>---------</td>
<td></td>
</tr>
<tr>
<td>2.4 Prototype Fabrication</td>
<td>58</td>
</tr>
<tr>
<td>2.5 Measurement</td>
<td>66</td>
</tr>
<tr>
<td>2.6 Thermal Analysis</td>
<td>71</td>
</tr>
<tr>
<td>2.7 Conclusion</td>
<td>73</td>
</tr>
<tr>
<td>3. Planar Millimeter-Wave Tightly Coupled Arrays</td>
<td>76</td>
</tr>
<tr>
<td>3.1 Design Challenges</td>
<td>77</td>
</tr>
<tr>
<td>3.1.1 Planar Implementation</td>
<td>77</td>
</tr>
<tr>
<td>3.1.2 Material and Process Selection</td>
<td>79</td>
</tr>
<tr>
<td>3.1.3 Limitations of PCB Processing</td>
<td>80</td>
</tr>
<tr>
<td>3.2 Design Approach</td>
<td>82</td>
</tr>
<tr>
<td>3.2.1 Development of Three-Pin Balun</td>
<td>82</td>
</tr>
<tr>
<td>3.2.2 Addition of a Conducting H-Wall</td>
<td>89</td>
</tr>
<tr>
<td>3.2.3 Sample Design for 5G Frequencies</td>
<td>93</td>
</tr>
<tr>
<td>3.3 Prototype Fabrication</td>
<td>97</td>
</tr>
<tr>
<td>3.4 Measurement</td>
<td>103</td>
</tr>
<tr>
<td>3.5 Conclusion</td>
<td>112</td>
</tr>
<tr>
<td>4. Measurement Techniques for Millimeter-Wave Antennas</td>
<td>113</td>
</tr>
<tr>
<td>4.1 Displacement Correction in Sequential Far-Field Measurements</td>
<td>113</td>
</tr>
<tr>
<td>4.2 Mathematical Model</td>
<td>115</td>
</tr>
<tr>
<td>4.3 Implementation &amp; Results</td>
<td>122</td>
</tr>
<tr>
<td>5. Conclusions and Future Work</td>
<td>127</td>
</tr>
<tr>
<td>5.1 Summary of this Work</td>
<td>127</td>
</tr>
<tr>
<td>5.2 Opportunities for Future Work</td>
<td>129</td>
</tr>
<tr>
<td>5.2.1 Scalability Beyond 70 GHz</td>
<td>129</td>
</tr>
<tr>
<td>5.2.2 UWB Components and Systems</td>
<td>131</td>
</tr>
<tr>
<td>5.2.3 Hardware Reduction in High Gain Phased Arrays</td>
<td>132</td>
</tr>
</tbody>
</table>

Bibliography | 134 |
List of Tables

<table>
<thead>
<tr>
<th>Table</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.1</td>
<td>Optimized parameter values of the split unit cell Ku-Band TCDA, shown in Fig. 2.7.</td>
</tr>
<tr>
<td>2.2</td>
<td>Optimized values of the design parameters shown in Fig. 2.23.</td>
</tr>
<tr>
<td>3.1</td>
<td>Minimum feature sizes for PCB fabrication, corresponding to the parameters shown in Fig. 3.2</td>
</tr>
<tr>
<td>3.2</td>
<td>Allocated Millimeter-Wave 5G and ISM Bands</td>
</tr>
<tr>
<td>3.3</td>
<td>Optimized values of the parameters shown in Fig. 3.15.</td>
</tr>
</tbody>
</table>
List of Figures

<table>
<thead>
<tr>
<th>Figure</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.1</td>
<td>Topside antennas for various communications, radar, and EW tasks aboard U.S. naval assets have doubled from 1980 to 1999. Image from [1].</td>
</tr>
<tr>
<td>1.2</td>
<td>Comparison of defense and emerging commercial distributed sensing networks. (a) Conceptualization of an ad hoc maritime sensor and communications network. Taken from [2]. (b) Intelligent transport network for coordinating autonomous vehicles. Taken from [3].</td>
</tr>
<tr>
<td>1.3</td>
<td>A simple circuit model of an antenna acting as impedance transformer to a load, $Z_L$. The load impedance consists of the free space impedance in parallel with a shorted transmission line, which represents the ground-plane.</td>
</tr>
<tr>
<td>1.4</td>
<td>Illustration of the source of fundamental limitations in the impedance bandwidth of an antenna above a groundplane. (a) Impedance response of the circuit depicted in Fig. 1.3. (b) Illustration of two zeros in the impedance response, when image currents on the groundplane cancel the aperture current (left) and when reflected radiation is out of phase with forward radiation (right).</td>
</tr>
<tr>
<td>1.5</td>
<td>Picture and dimensions of an all-metal, dual-polarized tapered slot array. Taken from [4].</td>
</tr>
<tr>
<td>1.6</td>
<td>Schematic of a fragmented aperture element synthesized via genetic algorithm, operating from 1.7–4.5 GHz at broadside. Each pixel is 1.5 mm square. Taken from [5].</td>
</tr>
<tr>
<td>1.7</td>
<td>Picture of the Current Sheet Array, which used capacitive coupling to approximate a constant current across the aperture, resulting in a 5:1 bandwidth. Taken from [6].</td>
</tr>
</tbody>
</table>
1.8 Picture of a Tightly Coupled Dipole Array with integrated baluns, capable of >7:1 bandwidth. Taken from [7]................................. 15

1.9 Achievable bandwidth of an array over a conducting groundplane, for various substrate materials and thicknesses. Notably $\epsilon_r > 1$ results in a reduction in bandwidth, and $\mu_r > 1$ an increase. Taken from [8]......................... 17

1.10 Sample periodic AMC structure, producing an effective LC circuit. At resonance, high reactive impedance causes in-phase reflection. Taken from [9]................................................................. 18

2.1 Overview of the TCDA unit cell developed in [10] (left) and detail of the integrated stripline balun (center). Transmission line model of the balun is shown on the right. ................................................. 25

2.2 Transmission line model of a Tightly Coupled Dipole Array and feeding network. The reactance $L$ is caused by the dipoles inherent inductance, and $C$ is a result of the capacitive coupling between neighboring elements in the array......................................................... 26

2.3 Proportional scaling of the unit cell design from [10] up to the Ku-band, by reducing all dimensions by a factor 0.22. The resulting minimum feature size and substrate thickness cannot be realized using commercial processes.......................................................... 27

2.4 Distributed element model of a transmission line. Primary constants of the line ($R, L, C, G$) are assumed to remain constant for the length of the line................................................................. 28

2.5 Cross sections of four common PCB-implemented transmission lines, with key parameters labelled. Representative electric field lines of the primary mode are indicated, and the direction of propagation is perpendicular to the plane of the page. From left to right, $C$ decreases and $Z_0$ increases, for constant values of $w, h$, and $\epsilon_r$............. 29

2.6 Illustrated Ku-band TCDA unit cell, showing co-planar dipoles, balun, and Wilkinson power divider. Microstrip ($Z_0 \approx 100\,\Omega$), stripline ($Z_0 \approx 20\,\Omega$), and twinline ($Z_0 \approx 200\,\Omega$) traces are used to equalize fabrication feature sizes despite a wide range in impedances................................. 30
2.7 Design parameters of a Ku-band TCDA utilizing the split unit cell approach. The four metal layers are displayed individually, and are combined according to the layer stack seen at the bottom. Optimized values are given in Table 2.1.

2.8 Simulated matching of the infinite-array design shown in Fig. 2.7. We observe operation from 3–18 GHz for VSWR<2 at broadside and scanning to ±45° in both planes for 3–17 GHz with VSWR<3.

2.9 Ratio of power radiated to power available, at broadside and 45° scans in the E- and H-planes. Losses up to 2.5 dB are observed for wide E-plane scans, which cannot be explained by mismatch alone.

2.10 Schematic layout of a linearly polarized radiator situated inside a waveguide cross-section. The boundary conditions at the PEC and PMC walls match those of an infinite array of identical and aligned radiators.

2.11 Conceptual diagram of an array radiating at an angle into a waveguide, such as in Fig. 2.10. The waveguide width at broadside ($d_1$) is reduced when projected into the scan direction ($d_2 = d_1 \cos \theta$). This results in an altered impedance, looking into the array.

2.12 Geometry of a dielectric slab above a groundplane. The dielectric region is assumed to be isotropic, and infinite in $y$ and $z$. Orientations of the guided TM and TE modes are shown.

2.13 Contour plots of equations (2.5)–(2.8) in the $k_c d, h d$ plane, for slabs of dielectric constant (a) $\epsilon_r = 2$, (b) $\epsilon_r = 8$. Intersections of the curves with the circles represent solutions to the equation pairs. The solid curves correspond to TM modes, whereas dashed curves correspond to TE modes, and the dotted curves represent dielectric slabs of varying thickness.

2.14 Geometry of an infinite, linearly polarized array above a groundplane. Dipole elements are shown to illustrate polarization, however subsequent derivations assume infinitesimal radiators.

2.15 Numerical solutions to the surface wave propagation factor ($\beta_{sw}$) as a function of substrate thickness, for (a) $\epsilon_r = 2$, and (b) $\epsilon_r = 8$. The first three TE and TM modes are labeled.
2.16 Solutions of (2.16) plotted in one quadrant of the $u,v$ plane, for an array spacing of $a = b = 0.5\lambda_0$, on $\epsilon_r = 2$ dielectric of thickness $d = 0.5\lambda_g$. Portions of the TE and TM circles which fall inside the visible region ($u^2 + v^2 < 1$) represent scan conditions at which those surface wave modes will become excited.

2.17 Detail of the visible region from Fig. 2.16, showing azimuth ($\phi$) and elevation ($\theta$) grids. TM$_0$ and TE$_1$ surface wave modes are plotted (solid and dashed curves, respectively), and will be excited at $44.4^\circ$ in the E-plane ($\phi = 0^\circ$) and $65.5^\circ$ in the H-plane ($\phi = 90^\circ$). Results are symmetric about $u$ and $v$ (i.e. for negative values of $\phi, \theta$).

2.18 Illustration of unbalanced currents on the differential line of the Marcdillard balun, resulting in net vertical currents. Unbalanced input feed and open circuit are shown for reference (dashed).

2.19 Strong vertical fields at resonance induce common-mode currents in the normally differential feed line. Notably, such resonance will more commonly occur diagonally between rows, rather than in-line as shown here.

2.20 Diagram of a planar array of dipoles, viewed from above. Dipoles are suspended in a vertical dielectric layer. Key values in determining the onset of a cavity resonance are indicated.

2.21 Infinite array input reflection, showing initial spurious cavity resonance (dashed peak at 15 GHz). Introduction of a shorting post shifts the resonance (dashed peak at 16.5 GHz), and finally increasing the dimension ‘A’ (see Fig. 2.22) pushes the resonance out of the operating band (solid curve).

2.22 Physical representation of the parameter $D_x$ in Fig. 2.20. Addition of a shorting post significantly reduces this value (labeled $D'_x$), as does adjustment of the dimension ‘A’.

2.23 Illustration of the proposed array periodic unit cell. Design parameters are noted as ‘A’–‘X,’ and three transmission line elements are identified. The design as shown has a 90 $\Omega$ input impedance, and requires tapering to operate in a 50 $\Omega$ system.
2.24 Simulated infinite array VSWR of the design in Fig. 2.23, using the values in Table 2.2. Broadside (red), and ±45° scanning in the E-plane (black) and H-plane (blue) are shown. Solid curves refer to a 90Ω impedance match, whereas dashed curves are matched to 50Ω using a taper of length \( \lambda_{hi} \). ................................................. 56

2.25 Infinite array VSWR for scans in the E-plane at ±50° (orange), ±60° (green), and ±70° (violet). Notably, the array maintains 88% of the broadside bandwidth down to ±70°, with VSWR < 2.2. Scanning beyond ±45° is achievable only in the E-plane. ................................................. 56

2.26 Infinite array impedance match of the proposed Ka-band design. Broadside (red), and ±45° scanning in E (black) and H (blue) planes are shown. This design can be fabricated with a minimum feature size of 4 mil................................................................. 58

2.27 Infinite array impedance match of the proposed millimeter-wave band design. Broadside (red), and ±45° scanning in E (black) and H (blue) planes are shown. This design can be fabricated with 3 mil tolerances. 59

2.28 Assembly diagram of the various components of the 8 × 8 array prototype. Detail of the 1 × 8 subarray assembly is shown below. .......................... 60

2.29 Image of two 1 × 8 subarrays before assembly, showing front and back metal layers (gold). The meandered input trace can be seen along the bottom of the lower board, as well as cutouts for the coax connectors. 61

2.30 Photograph of the prototype during assembly. Printed subarrays can be seen (black) with shaped dielectric spacers (white), as well as remaining slots in the groundplane. The superstrate is added last, above the spacers. ................................................................. 63

2.31 Photograph of the assembled prototype array, shown with superstrate removed. The 3D-printed spacers can be seen in white, and the fabricated antenna boards in black. ................................................................. 64

2.32 Photograph of the assembled array and groundplane. The HDPE superstrate is composed of nine separate strips which lie between the antenna boards, as indicated in Fig. 2.28. Future iterations of the array could integrate the superstrate and spacers into a single component to reduce assembly burden and possibility of error. .............................. 65
2.33 Phototgraph of the far-field measurement setup at the NASA Glenn Research Center. The array prototype is inserted in the center of a $5.5\lambda_{low}$ square ground plane.

2.34 Measured scanning patterns of the prototype array in the E-plane (left, red) and H-plane (right, blue), for 5.2 GHz, 8.4 GHz, 13 GHz, and 17.5 GHz (top to bottom). Simulated patterns (dashed black) are shown in each plot for reference.

2.35 Measured broadside co-polarized (solid) and cross-polarized (dashed) gain. Reference directivity (red) is calculated using the total aperture area.

2.36 Illustration of a continuous slot cut into the groundplane for inserting the antenna PCB (top). This structure can radiate when excited by the array feed. The suggested implementation (bottom) breaks up the long slot to prevent in-band radiation. The electric field orientation of co- and cross-polarized radiation is indicated for reference.

2.37 Measured peak in cross-polarized gain (dashed) vs. simulations that include the long slot in groundplane (crosses) show good agreement. Concurrently, we show that by breaking up the slot (circles) we observe 10 dB improvement at 7.5 GHz, and 18 dB improvement at 3 GHz.

2.38 Steady state thermal distributions (white-hot convention) of an infinite array, assuming 1 W RF input power at 17.5 GHz. High atmosphere conditions (even convection across structure with $h_0 = 5$) is shown on the left. Also, thermal distribution in vacuum conditions (no convection) with a $8.3^\circ$K/mW heat sink below the groundplane is shown on the right.

2.39 Observed thermal behavior during simulated high-power transmit operation, showing peak and volumetric average temperature change ($\Delta T$) above ambient for 1 W RF input power at individual frequencies.

2.40 Peak localized temperature change ($\Delta T$) above ambient as the RF input power is varied. The array was operated at 17.5 GHz, corresponding to maximal losses, as shown in Fig. 2.39.
2.41 Transient response of the unit cell showing instantaneous peak temperature, assessed at 17.5 GHz for 1 W RF input power. 

3.1 Illustration of planar (left) and non-planar (right) arrays. In both cases, a groundplane resides below the elements in the $x, y$ plane, and broadside is in $z$-direction. Notably in the planar case, the entire array can be formed from a single PCB, whereas in the non-planar array several PCBs must be combined to form the array. 

3.2 State of the practice Printed Circuit Board fabrication limits, shown from edge. Key limiting parameters are labeled, and provided in the Table 3.1. 

3.3 Circuit model representation of the antenna ($L, C, Z_{\text{sub}}$ and $Z_{\text{sup}}$) and feeding network ($Z_{\text{feed}}, Z_{\text{open}},$ and $Z_{\text{short}}$). Duplicated for clarity from Section 2.1. 

3.4 Model of an infinite array of tightly coupled dipoles above a groundplane fed from an ideal gap source, shown in black, (left) and $S_{11}$ response from 20–90 GHz (right). This model assumes a conducting wall along the H-plane, and will serve as the starting point for the design of the feed. 

3.5 Two conducting vias are added to the model from Fig. 3.4, shorting the dipoles to ground (left). This forms the high impedance shunt shorted stub seen in Fig. 3.3, inverting the $S_{11}$ response of antenna (right). 

3.6 Series open-circuit implemented in the style of [7, 10, 11], using a via opposite the shorting pins (left). VSWR is plotted to better show frequency-dependent behavior (right). Notably, the open circuit is effective in matching the low frequencies, but becomes detuned at high frequencies (grating lobe onset frequency is 90 GHz in this simulation). 

3.7 The open circuit is implemented as a patch on top of the opposite dipole. The $S_{11}$ plot shows all portions of the band are well matched. The strong coupling of the patch, compared to the pin in Fig. 3.6 mitigates detuning at higher frequencies. 

3.8 Conceptual model to route the antenna feed towards the groundplane. This requires that the short microstrip trace above the dipole can match the impedance of the antenna ($\approx Z_0 = 100 \Omega$).
3.9 Analysis of the impact of the microstrip trace in the feeding network. Due to the fabrication limitations, the trace width cannot be less than 3 mil. However, due to the proximity of the dipole, the trace $Z_0$ is low. This causes detuning of the element feed, even in the short length of the microstrip transition (right, $L = 0–10$ mil).

3.10 Side view of the three-pin balun shown in Fig. 3.8 (left), as well as equivalent transmission line model (right). This circuit can be understood by observing the similarity to the balun circuit shown in Fig. 3.3.

3.11 Illustration of the electric field distribution excited by the vertical currents of the input feed (dashed arrows) during cavity resonance.

3.12 Simulated VSWR demonstrating a destructive cavity resonance in the unmodified (central peak), alternate substrate (left peak), and shorting pin (right peak) designs. The resonance can be mitigated with a continuous conducting “H-wall”.

3.13 Top view of a $2 \times 2$ array, indicating the orientation of the resonant length $D$ in (3.1) of the spurious cavity mode. Based on the lengths $D_1$ (unmodified) and $D_2$ (with shorting pin), we expect that the addition of a shorting pin will only slightly increase the resonant frequency. Conversely, inclusion of an H-wall will significantly increase the resonant frequency.

3.14 Proposed unit cell, including integrated balun and H-wall for resonance mitigation. Key components are colorized and labeled for clarity.

3.15 Key design parameters of the nominal unit cell given in Fig. 3.14. Optimized values of these parameters are given in Table 3.3.

3.16 Infinite array VSWR of the array shown in Fig. 3.15, at broadside (black) and $45^\circ$ scanning in the E- (red) and H- (blue) planes. The 5G and ISM bands from Table 3.2 are highlighted for reference, and are observed to fall in the passband of the array.

3.17 Sample layout of a $3 \times 3$ array measurement coupon with the center element fed, shown from top, cutaway view (left) and bottom view (right). Approximate dimensions are included for reference.
3.18 Prototype 5 × 5 array (approximately 12 × 12 × 1.7 mm³) without measurement fixtures, alongside a U.S. penny (0.75 in diameter). This coupon is not used for measurement, but is representative of the array as it would be applied in practice. Black surface seen is the superstrate material (Duroid 5880), and thin tan layers are the prepreg (Rogers 2929).

3.19 (a) Initial fabrication delivered by vendor lacked mounting holes for the coax port. (b) Holes were drilled as rework, but suffered low accuracy due to lack of fiducial markers on de-panelized boards. As this was anticipated, hole diameter was increased to allow for misalignment.

3.20 Simulated $S_{21}$ of the coax to microstrip transition, with the coax port shifted laterally by 0–10 mil (0–0.254 mm). We note at 15 mil offset the signal is completely reflected (not shown in figure).

3.21 Excess substrate found at the edge of the coax footprint is a result of de-panelization. The thickness of the substrate boundary varies between coupons, and must be removed to avoid losses.

3.22 Simulated $S_{21}$ of the coax to microstrip transition, for various values of the excess dielectric shown in Fig. 3.21. Negative values imply too much has been removed, cutting into the coax footprint. We note at 30 mil, $S_{21} < -10$ dB (not shown in figure).

3.23 (a) Mounted 1.85 mm coax port showing alignment of launch pin and microstrip trace. The port is bolted in place, not soldered, and thus can be moved or removed as needed. (b) Picture of a 3 × 3 coupon with coax port mounted; ruler included for scale.

3.24 Four 3 × 3 array coupons selected for measurement, shown with coax ports mounted and resistors soldered. Mounted calibration standard for de-embedding is shown to the right.

3.25 Locations of the active elements as well as failed connections after soldering of 50 Ω resistors.
3.26 (a) Schematic of the far-field test setup used for gain characterization of the fabricated prototypes. (b) Picture of the millimeter-wave range at the NASA Glenn Research Center, with key elements labelled. Open-ended waveguides can be seen mounted at both the source and receive ports. .......................................................... 106

3.27 (a) Picture of the U-band open-ended waveguide source mounted at port 1. (b) Picture of one coupon mounted at port 2. The coax cable feeding the antenna is run inside the mounting post; the entire post rotates during measurement to change the angle of incidence. ........ 107

3.28 Back-to-back array feed probes, used to measure insertion loss of the measurement fixture. .......................................................... 108

3.29 Measured embedded element gain at broadside (blue) compared to simulated values (red) for two measurement coupons, with the array oriented horizontally. Edges of the measurement bands are indicated (dashed lines). .......................................................... 109

3.30 Measured embedded element gain at broadside (blue) compared to simulated values (red) for two measurement coupons, oriented vertically. Edges of the measurement bands are indicated (dashed lines). ....... 110

3.31 (a) Semi-finite simulation model which assumes a periodic boundary along one side of the array to reduce computational burden. This was the model used for the results shown in Figs. 3.29 and 3.30. (b) Complete model including extended substrate and feedline of the measurement coupon (shown from below). .......................................................... 111

3.32 Measurement data of the center element compared against 1 × 3 and 3 × 3 simulations (models shown in Fig. 3.31). As expected, the increased accuracy of the 3 × 3 simulation introduces nulls that are more noticeable at the low frequencies and which are also observed in the measured results. .......................................................... 111

3.33 Measured (solid) and simulated (circles) element patterns at 52 GHz and 56 GHz show close agreement. We observe that dips in the broadside gain (see Fig. 3.32) are a result of changes in the element pattern over frequency. .......................................................... 112

xx
4.1 Measurement setup of a two-element array (the Antenna Under Test) with an incident plane wave from a transmitting horn. The array is physically displaced from the center of rotation by offsets $x$ and $y$. We note that $x$ and $y$ align with the coordinate axes $\hat{x}$ and $\hat{y}$ when the array is oriented to broadside.

4.2 Relative positions of eight elements of a linear array, traveling from $-90^\circ$ to $90^\circ$. We note at broadside, all eight elements are in-phase (zero relative separation) whereas at $\pm 90^\circ$ they are out of phase (0.5\textlambda separation).

4.3 Relative positions of eight elements at some offset from the center of rotation (refer to Fig. 4.1): (a) with $x$-offset only, (b) with $y$-offset only, (c) with $\theta$-offset only, (d) with arbitrary offsets of each element.

4.4 Computed array factors of isotropic radiators with ideal (solid) and distorted (dashed) characterizations, as shown in Fig. 4.2 and Fig. 4.3d, respectively. Notably, even small phase errors ($\sigma = 0.125\lambda$) contribute to significantly increased sidelobe levels and beam skew.

4.5 Eight element paths during measurement, calculated from measured phase data (dashed) and least-squares fitted sine curves (solid).

4.6 Element trajectories after removing measurement errors. Measurement data is shown in black, and fitted curves used for processing in blue.

4.7 Comparison of normalized array patterns using measured gains with ideal phase response (red), raw measured data (black), and error-corrected data (blue). Notably, the error-corrected data closely follows the expected, ideal response.

4.8 Derived measurement offsets in $y$ (solid lines) and $x$ (dashed lines). As expected for this measurement setup, $x$ offsets remain relatively constant, while $y$ offsets change with each measurement.

5.1 Miniaturized unit cell design utilizing an offset feed via, scalable to 90 GHz (left). Asymmetry in the offset feed results in high cross-polarized gain at high frequencies, which can be reduced by duplicating the feed via to form a symmetric unit cell (right, shown without H-wall for clarity).
5.2 Co- and cross-polarized gain (Ludwig’s third definition) of the offset feed design shown in Fig. 5.1a. Notably, co- and cross-polarized gain are nearly equal at high frequencies, implying diagonal polarization.

5.3 Application of frequency diplexing to channelize the UWB antenna output, in order to: (a) overcome band-limited components, (b) reduce oversampling of low frequencies by forming frequency dependent subarrays.
Chapter 1: Introduction

The need for wideband antenna systems is driven by a confluence of factors. The ever-growing need for bandwidth, continually shrinking platforms, and increasingly multi-functional systems are traits observed throughout the defense, scientific, and consumer-electronic communities. Ultra-wideband (UWB) phased arrays provide a highly effective solution to these challenges. However, the widespread acceptance and use of these systems will require robust, compact, and low-cost implementation. The goal of this dissertation is to develop a novel array design, leveraging mass-market compatible fabrication techniques, to provide the needed bandwidth and efficiency for emerging applications in the millimeter-wave spectrum.

1.1 Applications of Ultra-Wideband Phased Arrays

With the advent of smart phones has come a revolution in mobile data consumption. Both the bandwidth of each connection as well as the user base continue to grow, resulting in an exponential growth of cellular and mobile broadband traffic [12]. The challenge of delivering this bandwidth is compounded by the iterative allocation and regulation of the electromagnetic spectrum, which has resulted in scattered frequency bands. Paired with the highly integrated and multi-purpose nature of these devices,
smart phones now operate in as many as seven different frequency bands [13]. Moreover, this effect is only growing. Most recently, various components of the coming Fifth Generation mobile architecture (5G) have been allocated across six new frequency bands [14] in the millimeter-wave spectrum.

Of course this challenge is not unique to consumer devices. Increasingly ubiquitous platforms such as small satellites, unmanned aerial vehicles (UAVs), and autonomous passenger vehicles all face the fundamental challenge of providing multiple wireless services from a small or conformal space. Simultaneously, many of these applications are themselves very wideband. Low-profile, ultra-wideband (UWB) apertures are a fundamentally enabling development for advanced remote sensing techniques, including foliage penetration (FOPEN) [15], through-wall imaging [16], medical imaging [17], and high resolution radar [18]. Likewise, both radio astronomy [19] and Electronic Warfare (EW) [20] depend on the ability to access wide swaths of spectrum. Spread spectrum techniques have emerged as a tool for secure, low probability of intercept [21], and disruption tolerant [22] communications. As well, the need for high-reliability systems such as satellites to be disruption tolerant has led to strong interest in frequency-agile, software defined radio (SDR) technologies [23,24].

In this regard, one critical application of UWB technologies is the formation of a multi-functional aperture. Leveraging time, frequency, and spatial multiplexing, many separate wireless functions which would conventionally utilize separate antennas can be consolidated into a single shared aperture. This concept has a heritage in military platforms, where managing numerous radar, communications, and fire-control systems has been a long-standing challenge (see Fig. 1.1). Early efforts such as the Multifunction Electronically Scanned Adaptive Radar (MESAR) [25,26] had
a modest bandwidth (20%), relying instead on rapid sequencing. Later systems such as the Advanced Multifunction RF Concept (AMRFC) [1] would more significantly take advantage of frequency multiplexing, supporting a 1–18 GHz operating bandwidth, split between two UWB arrays. Similar concepts spanning 2–18 GHz have been studied for aircraft [20]. The complex operational environment encountered by such systems include ad hoc sensing, networking, data relay, and communications with local units as well as satellites or ground stations. But this is no longer very different from the highly multi-functional vehicles, satellites, and even handheld devices being developed today (see Fig. 1.2).
Figure 1.2: Comparison of defense and emerging commercial distributed sensing networks. (a) Conceptualization of an ad hoc maritime sensor and communications network. Taken from [2]. (b) Intelligent transport network for coordinating autonomous vehicles. Taken from [3].
Figure 1.3: A simple circuit model of an antenna acting as impedance transformer to a load, $Z_L$. The load impedance consists of the free space impedance in parallel with a shorted transmission line, which represents the groundplane.

1.2 Review of Existing UWB Array Technologies

A useful tool for understanding the limitations of UWB arrays is by considering the role of the antenna as that of an impedance transformer, from the 50 Ω system to a 377 Ω load, in free space. However this simple picture is complicated by the nearby presence of a reflective groundplane, which is required to avoid bidirectional radiation. This manifests as a parallel short circuit to the free space load, as is shown in Fig. 1.3. As is shown in Fig. 1.4, the parallel short circuit introduces zeros in $Z_L$, which cause total reflection at the input. This fundamentally bounds the achievable impedance bandwidth to the space between two zeros.

Immediately, we see that increasing the separation from the groundplane serves to increase bandwidth by reducing the rate of change of the groundplane reactance. This inverse relationship to the groundplane height presents a miniaturization challenge,
similar to the way the Wheeler and Chu relations [27, 28] place a limit on the size of small antennas. Indeed, matching the reactance stemming from the groundplane is the primary challenge in designing low-profile UWB arrays.

Naturally, one approach by which to access a larger bandwidth is to simply include multiple narrow-band resonators. Multi-resonant antennas such as the stacked patch antenna combine several independently resonant structures with neighboring resonant frequencies to achieve a wider effective bandwidth. This has been demonstrated in [29] to achieve up to 2:1 bandwidth. A logical extension to this approach is to continue adding resonators at a regular rate of scaling, as in the Log-Periodic Antenna (LPA) [30]. However these techniques are inherently dispersive, and the antenna size is determined by the lowest frequency—limiting their use in arrays. Similarly, there exist certain families of UWB antennas which have excellent performance, but are not suitable for electronically-scanning arrays. This includes antennas which do not radiate normal to the groundplane, such as wideband monopoles [31], as well as electrically large antennas which would exceed the grating lobe spacing, such as spirals [31].

Despite these limitations, various groundplane-backed UWB arrays have been developed with operational bandwidths ranging from an octave up to a decade or more. In this section, we will review the primary UWB array design approaches, providing a brief introduction to their modes of operation as well as relative strengths and weaknesses.
Figure 1.4: Illustration of the source of fundamental limitations in the impedance bandwidth of an antenna above a groundplane. (a) Impedance response of the circuit depicted in Fig. 1.3. (b) Illustration of two zeros in the impedance response, when image currents on the groundplane cancel the aperture current (left) and when reflected radiation is out of phase with forward radiation (right).
1.2.1 Tapered Slot

Perhaps the most common and well-understood UWB array element is the exponentially tapered slot, also variously referred to as a “flared notch” or “Vivaldi” antenna. These have been in use for over four decades [32], and have demonstrated over 10:1 bandwidth and scanning up to 60° [4] (see Fig. 1.5). Further, the very simple layout of tapered slot antennas makes them easy to modify for specific needs, such as dual-polarizations [33], modularity [34], or power handling [35].

The element consists of a gradually tapered slotline, which serves to significantly reduce the high impedance of the propagating wave. The slot line is typically excited by a simple balun. In this way, a relatively low radiation resistance ($R_{rad}$) is seen at the feed, and prevails over the groundplane inductance ($L_{GP}$) such that $R_{rad}||L_{GP}$ is mostly real. Therefore, Vivaldi bandwidth is mostly determined by the bandwidth of the impedance taper from the feed to free space. Just like conventional microstrip impedance transformers [36], this can be accomplished for very wide bandwidths using gradual tapering profiles.

However, the primary drawback of tapered slot antennas is their length. Due to the need for wideband impedance tapering, these antennas are typically designed to be a quarter wavelength (0.25 $\lambda$) at the lowest frequency of operation—translating to 2–3 $\lambda$ at the high frequency. This makes them extremely bulky and heavy for low frequency applications. Additionally, the vertical currents running along the slot’s length can cause high cross-polarized radiation when scanning. In some cases, cross-polarized radiation even exceeds the co-polarized radiation, at as little as 30° scanning [37]. Elevated cross-polarization can be combated to a certain extent with
Figure 1.5: Picture and dimensions of an all-metal, dual-polarized tapered slot array. Taken from [4].
active compensation algorithms in dual-polarized arrays, but are ultimately limited in their angular coverage and bandwidth [37].

Numerous variants of the tapered slot antenna exist [34,38], which seek to address size and cross-polarization issues. These generally utilize a more aggressive tapering profile or truncated fins to reduce height [33]. In particular, the Balanced Antipodal Vivaldi Array (BAVA) [38] feeds the flared slotline directly from a stripline input, eliminating the need for a balun, and thereby achieves a $\lambda_{hi}/2$ profile. The low profile improves cross-polarization performance, however limits bandwidth to a 4:1 range, and scanning to $\pm 45^\circ$ [33]. Additionally, the asymmetric layout of BAVA elements can lead to asymmetry in the E-plane element pattern [38]. More recently, an unbalanced-fed and capacitively coupled corrugated notch has been demonstrated to achieve up to 7:1 bandwidth in a $\lambda_{hi}/2$ profile [39,40].

1.2.2 Fragmented Aperture

With the advent of accurate and reasonably fast electromagnetic modeling codes in the mid 1990’s came a revolution in the ability to analyze complex radiating structures. As a result, the use of evolutionary optimizers became a popular tool for tackling complex electromagnetic problems [41]. Naturally, one such problem was the development of planar, broadband radiators with better gain profiles than canonical elements such as a spiral or bowtie [42]. This would be achieved by treating the aperture as a blank canvas, dividing the available space into small discrete regions which are treated as either conductive or insulating. An iterative Genetic Algorithm (GA) assigns each region, simulates the electrical performance, and compares the result
against a set of goals. As seen in Fig. 1.6, the resulting antennas are composed of discrete, visually fragmented scattering structures.

Early on, it was determined that a critical feature in achieving broadband performance in such arrays was allowing for electrical contact between neighboring elements [42]. In this way, many elements (electrically small at such low frequencies) could work together to support the necessary radiating currents. However, this also implied a minimum size of the total array in order to support those frequencies. Initial work focused on bidirectional radiators, which were able to produce 10:1 bandwidth [42]. However, when placed above a groundplane this is significantly reduced, to around 3:1 [5]. Much interest has also been paid to the optimization of multi-layered, frequency selective absorbing layers, to overcome this limitation. These have
been applied to achieve up to 33:1 bandwidth in a planar array above a groundplane, though with a $2.75\lambda_{hi}$ profile, and only $> 50\%$ efficiency [43,44].

The fragmented aperture approach yields very novel and non-intuitive designs, and has also been applied successfully to improve the design of reflectors [45] and Frequency-Selective Surfaces (FSS) [46], among others. However, GA optimization naturally has a high computational burden—the fine resolution designs observed in [42] require access to parallel supercomputing clusters to converge, though lower resolution designs suitable for optimization on a desktop PC have also proven useful [5]. Likewise, this computational expense limits the optimization to planar designs, and prevents the co-optimization of feed structures, which may lend to increased bandwidth in low-profile designs. That being said, with the continuing advances in computational power, we expect it is merely a matter of time before one may encounter three-dimensional GA-synthesized designs. Finally we wish to note that while good performance has been demonstrated, blind reliance on optimizers makes it challenging to learn from, and meaningfully expand on, these designs.

1.2.3 Connected and Coupled Arrays

Intuitively, the upper frequency bound of a phased array is determined by the onset of grating lobes, which bounds the element spacing to $\approx \lambda/2$. Likewise if these elements are isolated from one another (i.e. no mutual coupling), then the current distribution falls to zero at the element edges, preventing the element from supporting significantly lower frequencies. These two conditions result in a limited fractional bandwidth. We have seen that tapered slots overcome this by extending the element orthogonal to the groundplane. Similarly, the effective length of the
element can also be expanded horizontally if neighboring elements are connected to one another, either electrically or through a strong mutual coupling. In this way, long continuous currents, and thus long wavelengths, can be supported along the aperture across numerous electrically small elements.

The consideration of implementing distributed sources to produce an aperture current are seen as early as 1970 [47]. More recently, the use of interconnected elements to expand the low frequency response of an array was developed by Munk [6]. This approach was inspired the treatment of a phased array of dipoles as a continuous current sheet, by Wheeler [48]. Munk’s implementation consisted of an array of dipoles, which were capacitively coupled using inter-digitated tips. Called the “Current Sheet Array” (CSA), this array demonstrated nearly 5:1 bandwidth over a groundplane. The CSA is shown in Fig. 1.7

Arrays utilizing capacitive coupling between element tips are often referred to as Tightly Coupled Dipole Arrays (TCDA). Conversely, “Connected Arrays” utilize a direct electrical connection between radiating elements [49]. Direct connection serves the same purpose of supporting long wavelengths across multiple elements, and is capable of very large bandwidths in free space. However, bandwidth is starkly limited in the presence of a groundplane (2:1 bandwidth observed in [49]). Indeed, the series capacitance observed in TCDAs is useful in partially mitigating the large shunt inductance stemming from the groundplane, at low frequencies.

One drawback to Munk’s CSA was the need for bulky external baluns to provide a balanced excitation to the dipole elements. Due to the electrically small volume of the unit cell across most of the band, integrating a balun above the groundplane is challenging, and can significantly reduce the array bandwidth [50]. However, recent
work has produced a folded Marchand balun which can be integrated directly within the unit cell volume [7]. Moreover, the balun can be exploited as an additional impedance matching network, expanding the bandwidth of the array. This has been demonstrated in [7] to produce a 7:1 bandwidth in an array only $\lambda_{hi}/2$ thick (see Fig. 1.8), as well as in [10] for a 6:1 bandwidth with scanning to ±60° from broadside in all planes. Alternatively [51] feeds a coupled dipole element directly from the unbalanced source, to achieve a 5:1 bandwidth.

Additionally, several variants and corollaries of connected and coupled arrays have been demonstrated. This includes the magnetic dual of the connected array, the “Long Slot Array” [52], which uses long continuous slots to produce a constant magnetic current along the aperture. Paired with ferrite loading, this has demonstrated 10:1
Figure 1.8: Picture of a Tightly Coupled Dipole Array with integrated baluns, capable of >7:1 bandwidth. Taken from [7].

bandwidth [52]. Similarly, a TCDA with resistive FSS loading is shown in [11], which effectively doubles the array bandwidth (to >13:1) by eliminating the λ/2
groundplane resonance. Likewise, the concept of tight coupling between elements has
been used to extend the bandwidth of already wideband elements such as spirals.
In this way, the Interwoven Spiral Array (ISPA) [53] achieves a 10:1 bandwidth for
VSWR < 2. However, the ISPA suffers from high cross-polarization at low frequencies,
as well as resonances when scanning.

1.2.4 Material Loading

Material loading can impact the impedance behavior of the array. In particular,
bulk dielectric is often found between the aperture and groundplane, as the antenna
substrate. By reducing the wavelength of the propagating signal, this is an effective
tool for reducing the antenna height. However, increasing the dielectric constant of
the propagation media results in a reduction of the wave impedance. This in turn
increases the reactance of the groundplane (see Fig. 1.3), decreasing the achievable
bandwidth of the array [54].

Conversely, dielectrics can be placed above the radiating aperture, where they are
referred to as a superstrate. Here they act as an intermediate matching stage, reduc-
ing the free space impedance seen at the aperture. In particular, if the superstrate is
\( \frac{\lambda}{4} \) in thickness it behaves analogous to a quarter-wave impedance transformer. By
reducing the radiation impedance, the total load on the antenna \( R_{rad} | Z_{GP} \) is reduced.
This increases the impedance bandwidth of the array. Further, wave slowdown in the
superstrate reduces the effective angle of incidence of free-space waves, resulting in
a more constant impedance behavior across the scanning volume. For this reason,
superstrates are often used to improve scanning performance and are sometimes re-
ferred to as Wide-Angle Impedance Matching (WAIM) layers. However, excessively
thick or high dielectric constant superstrates can support surface waves, resulting in
severe mismatch at the resonant frequency.

Unlike dielectrics, magnetic materials (\( \mu_r > 1 \)) used in the substrate can increase
the bandwidth of low-profile arrays. By increasing the wave impedance, the shunt
reactance of the groundplane is suppressed, resulting in an primarily real load across
frequency. Unfortunately, magnetic materials are generally heavy and lossy, limiting
their use in many applications. The maximum achievable bandwidth for an array
above a groundplane when utilizing various dielectric or magnetic substrate materials
is shown in Fig. 1.9.
Figure 1.9: Achievable bandwidth of an array over a conducting groundplane, for various substrate materials and thicknesses. Notably $\varepsilon_r > 1$ results in a reduction in bandwidth, and $\mu_r > 1$ an increase. Taken from [8].

Similarly, if the PEC groundplane were replaced with a perfect magnetic conductor (PMC) boundary, in-phase reflection ($\Gamma = 1$) would allow the antenna to sit directly above the groundplane. While these materials do not exist in nature, much ongoing research pertains to the synthesis of artificial magnetic conductors (AMC) which can emulate this response over a finite band. Generally this involves the design of subwavelength periodic or corrugated surfaces (see Fig. 1.10), which can resonate to produce a high surface impedance [9]. However, these materials are only effective over a narrow band.

A larger groundplane distance serves to increase bandwidth by reducing the reactive impedance seen at low frequencies. However this is ultimately limited by the destructive resonance which occurs when the groundplane separation reaches a half

17
Figure 1.10: Sample periodic AMC structure, producing an effective LC circuit. At resonance, high reactive impedance causes in-phase reflection. Taken from [9]

wavelength. This resonance can be suppressed, of course, using lossy materials to absorb all backwards radiating energy. In this way, the presence of the groundplane can be negated, resulting in very wide bandwidths [43], but results in a maximum efficiency of 50%. However, losses can be reduced by confining the lossy material to a relatively narrow-band Frequency Selective Surface (FSS), paired with the synergistic design of a superstrate. This was demonstrated in [11] to achieve a 13:1 bandwidth above a groundplane, while maintaining efficiency > 70% across the entire band.

1.3 Contemporary Challenges in UWB Phased Arrays

As we have seen in Section 1.1, systems designers continue to require wider bandwidth and greater scanning volume, while reducing size, weight, and power consumption in modern phased arrays. While much progress has been made in the development of UWB phased arrays over recent decades, several major challenges remain to be addressed. Simultaneously, emerging applications present new requirements to be met.
Beyond the fundamental limitations discussed in Section 1.2, relating bandwidth, scanning range, and thickness, there exist several practical challenges in realizing the UWB arrays needed for many applications. These are highlighted as follows.

**Growing Bandwidth, Shrinking Platforms:**

Just as the demand for bandwidth has exploded over recent years, so too has the proliferation of small and highly mobile platforms which use it. This includes the growth in popularity and capability of Unmanned Aerial Vehicles (UAVs), as well as new interest in small satellites (including CubeSats). While the demand for bandwidth has driven interest in higher frequencies, which are inherently physically smaller, the issue of integration with such small platforms remains very challenging. By way of example we can consider that even at millimeter-wave frequencies, arrays must be low-profile in order to conform with cellphones and other consumer devices, which themselves are only several millimeters thick.

**Efficient Feed Structures:**

Many UWB arrays utilize a balanced element; however, signal distribution is often handled in unbalanced lines (for instance coax or microstrip). This necessitates the integration of compact and wideband baluns to efficiently excite the element. Simultaneously, these feed structures play an important role in impedance matching across frequency and scan angles. However, antenna bandwidth can be undercut by spurious resonances and cross-polarized radiation, both of which are often excited by feed structures. Furthermore, the utilization of split unit cells in the feed network [7, 11] have been demonstrated to contribute losses up to 1 dB when scanning [10].
Effective Fabrication for Millimeter-Wave Arrays:

The challenge of effective feed and radiator design is further complicated by the limited fabrication techniques suitable for millimeter-wave applications. Microfabrication techniques are capable of extremely fine features and tolerances, and would allow direct implementation of antennas above the transceiver. However, the radiating aperture is still much larger than the transceiver die, implying a significant increase in cost for large scale application. Conversely, standard printed multi-layer circuit boards are extremely low cost, but suffer from limited feature sizes. Finally, most current UWB arrays take advantage of assemblies of end-fire radiators, which are not feasible at higher frequencies. Planarization of these arrays represents a significant design challenge.

Accurate and Error Tolerant Characterization:

Characterization of millimeter-wave antennas is challenging due to the sensitivity of such measurements to small errors. This includes physical displacements in repeated measurements, variance in fabricated prototypes, and distortions resulting from electrically large interfaces to such antennas. Whereas specialized, often automated, measurement systems exist to address this issue, these are expensive and of limited availability. It is critical to extend the functionality of existing measurement systems into the millimeter-wave band, to increase accessibility to the broader antennas community.
1.4 Contribution and Organization of this Dissertation

The focus of this dissertation is in addressing the challenges outlined in Section 1.3 through the development of novel designs and methods. In particular, the key contributions are as follows: 1) design of a high efficiency and frequency scalable TCDA feed, 2) development of a planarized UWB antenna suitable for millimeter-wave applications, and 3) novel measurement techniques to improve the accuracy of high frequency antenna characterization.

In Chapter 2, the canonical TCDA design is modified to improve efficiency, by eliminating the need for a miniaturized unit cell. Additionally, modifications are made to relax fabrication feature size requirements in the antenna feed, allowing for scalability to higher frequencies. The presented design is shown to operate across 3.5–18.5 GHz, with scanning as far 70° in E-plane. Critically, this design is compatible with linear scaling up 49 GHz without change to the fabrication techniques. An \(8\times8\) prototype is fabricated and measured to validate the simulation results.

Following, Chapter 3 presents the development of a planarized TCDA and balun to extend operation into the millimeter-wave band. Planar co-fabrication of the entire array is critical in eliminating the need for complex assembly of the array at such small scales. Simultaneously, compatibility with low-cost PCB fabrication enables the potential for true mass-market applicability. Optimized design parameters for operation across 24–72 GHz are given, and the tolerable variation of these values in practice is discussed. The design is validated by fabrication of \(3\times3\) and \(5\times5\) prototype arrays.
Chapter 4 presents novel measurement techniques necessary for accurate characterization of antennas at millimeter-wave frequencies. First, a post-processing algorithm is discussed, which serves to compensate the measured phase response of the array for inevitable small movements occurring during measurement. This is developed mathematically, and is verified through measurements at Ku band (18 GHz).

Finally, Chapter 5 summarizes this research and provides conclusions. Opportunities for continuing studies to build on this work are also discussed.
Chapter 2: Design of a Tightly Coupled Array Operating up to 49 GHz

In this chapter, we will make the first thrust towards a TCDA design suitable for millimeter-wave applications. Initially, we consider the intermediate frequencies of interest for space-borne applications, including the S (2–4 GHz), C (4–8 GHz), X (8–12 GHz), Ku (12–18 GHz), and Ka (27–40 GHz) bands. The NASA Space, Near-Earth, and Deep-Space Networks (SN, NEN, and DSN respectively) fall into these bands, as do many other commercial, government, and imaging applications. Communications to/from and between satellites are the enabling infrastructure of NASA’s scientific mission. High data throughput in these communications, despite an increasingly congested spectral environment, necessitates large bandwidth agility in communication systems for the emerging space architecture [24]. Simultaneously, ongoing reduction of size and weight in satellite systems is a must for all space operations.

In this regard, we note that current satellite antennas are predominately mechanically steered reflectors [55, 56]. These antennas are also typically narrow band and single access. Further, physical movement through a powered gimbal is slow and limited in range, while requiring high power and introducing vibrations which can disturb sensitive instruments. As such, there is high interest in the development of
low-profile, ultra-wideband (UWB), electronically steered arrays. These arrays can reduce weight by integrating multiple functions into a single aperture, while increasing spectral agility for intra-satellite communications, broadband sensing, and cognitive operation.

In particular, we demonstrate a versatile array design capable of a tenfold increase in operating frequency over [7, 11], while maintaining the same fabrication feature sizes. Critically, the extension of capabilities achievable using simple and accessible PCB fabrication makes available a wide range of new antennas and applications in this frequency space, which could otherwise not be realized. Considerations are provided as they relate to material selection for space antennas. Also, multi-physics analysis is provided to validate the design’s suitability for high-power transmit applications. Portions of this chapter have been published in [57].

2.1 Initial Design Approach

As discussed in Section 1.2.3, the early TCDAs from Munk [6] and later from Kasemodel [50] operated at higher frequencies (up to 18 GHz), but were limited to 4:1 and 2:1 bandwidths, respectively. Development of a higher-order matching network, in the form of an integrated Marchand balun, enabled the extension of the TCDA bandwidth to 6:1 [10] and 7:1 [7]. However, this wideband performance is achieved at the expense of increased complexity in the feed.

This can be seen in [10], where a unit cell design is developed for operation in the UHF band. In particular, we observe a miniaturized Marchand balun, consisting of a stripline input and open circuit stub, with a gap in the shielding at the feedpoint of the dipoles. A balanced slotline mode is excited in this gap, both feeding the dipoles and
Figure 2.1: Overview of the TCDA unit cell developed in [10] (left) and detail of the integrated stripline balun (center). Transmission line model of the balun is shown on the right.

propagating in a guided mode to the groundplane, acting as a parallel short circuit stub. These components of the physical layout are mirrored in the transmission line model, also shown in Fig. 2.1.

The model shown in Fig. 2.1 can be combined with the basic model of an antenna above a groundplane given in Fig. 1.3 to form a more general model of a TCDA with integrated balun, as shown in Fig. 2.2. In addition to the balun circuit, groundplane, and free-space load, this includes a short superstrate layer, $Z_{\text{superstrate}}$, as well as the dipole self-inductance $L$ and the capacitive coupling $C$ characteristic of coupled arrays. This model was originally developed in [7], and is useful to understand the basic components of the antenna and feed.
Figure 2.2: Transmission line model of a Tightly Couple Dipole Array and feeding network. The reactance $L$ is caused by the dipoles inherent inductance, and $C$ is a result of the capacitive coupling between neighboring elements in the array.

Of course, linearly scaling of previous designs would readily achieve identical bandwidth and beamforming performance at Ku-band. However, fabrication of the scaled array on low-cost Printed Circuit Board (PCB) is not feasible. By way of example, the array in [10] operates up to 4 GHz, and requires traces as thin as 0.107 mm (4 mil). Thus, scaling the previous design to 18 GHz would require the manufacture of features on the order of $23.78 \, \mu\text{m}$ (0.889 mil), as shown in Fig. 2.3. This far exceeds the capability of PCB techniques, which support a minimum feature size of $76 \, \mu\text{m}$ (3 mil).

In particular, we observe that the smallest feature of the design comprises a high impedance line within the feeding network of the antenna, labeled $Z_1$ in Fig. 2.3. Indeed, the TCDA’s integrated balun requires an approximately 20–200 $\Omega$ (10:1) range in characteristic impedances ($Z_0$) to maintain wideband performance [58]. Instead, we must produce the requisite range of impedances without decreasing the minimum
Figure 2.3: Proportional scaling of the unit cell design from [10] up to the Ku-band, by reducing all dimensions by a factor 0.22. The resulting minimum feature size and substrate thickness cannot be realized using commercial processes.

Feature size. To understand how this may be possible, it is useful to consider the lumped-element model of a transmission line, as shown in Fig. 2.4. In this approximation, each infinitesimally short length of transmission line is treated as an RLC circuit. The line is assumed to be very long, with no reflections present. The characteristic impedance ($Z_0$) of the transmission line can then be calculated as:

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}}. \tag{2.1}$$

Equation 2.1 provides some insight as to how the four primary transmission line properties can be manipulated in order to achieve the desired characteristic impedance. However, certain practical considerations must be taken into account. In particular, $G$ and $R$ are determined by the conductivity of the substrate and conductor, respectively, and are both minimized to reduce losses. Similarly, $L$ is largely
Figure 2.4: Distributed element model of a transmission line. Primary constants of the line \((R, L, C, G)\) are assumed to remain constant for the length of the line.

determined by the substrate material’s permeability \((\mu_r)\) and the conductor surface roughness. These are not practical design parameters to control. Therefore, all that remains is an inverse relationship to the capacitance per unit length, \(C\).

This relationship explains, for example, why widening a microstrip trace (thus increasing shunt capacitance) decreases its characteristic impedance. Similarly, increasing the height above the groundplane reduces \(C\) and therefore increases \(Z_0\). Line width and height are useful for tuning, however are limited in their total range. More significant changes in \(C\) and thus \(Z_0\) can be achieved by altering the configuration of the transmission line. This can be seen in Fig. 2.5, wherein several common PCB transmission line cross-sections are compared. As noted before, the values \(w, \epsilon_r\) and \(h\) can be used to adjust \(Z_0\). However, for the same values of \(w, \epsilon_r\) and \(h\), the four transmission lines shown will exhibit increasing \(Z_0\) (from left to right) on account of the decreasing capacitive surface area between the conductors.

Thus, the wide range of impedances needed in the balun is accomplished by combining several of these transmission line formats. Particularly, the high \(Z_0\) feed entering the balun is implemented as microstrip to enable an increase in the trace width.
Likewise, the very low $Z_0$ open stub can be made more narrow by implementing stripline with thin dielectrics. This is further emphasized by staggering the two arms of the balun on separate layers, such that the grounding arm of the open circuit side is closer to the signal trace (thus increased $C$), while the grounding arm of the input side is further away (decreased $C$). Finally, the high $Z_0$ parallel short circuit is formed by the two legs of the balun, which act as a twinline waveguide. These elements are highlighted in Fig. 2.6, with the dipoles included vertically, in-line with the balun.

Applying these techniques, an initial design is shown in Fig. 2.7. Key design parameters are highlighted, and the optimized values are given in Table 2.1. The design implements four metal layers on two Rogers Durroid 5880 substrates (5 mil, $\epsilon_r = 2.2$), with PolyFlon prepreg (4 mil, $\epsilon_r = 2.2$). These materials and thicknesses are commercially available, and compatible with PCB techniques. Additionally, a 3 mm polyethylene (HDPE, $\epsilon_r = 2.3$) superstrate is included, as indicated in Fig. 2.6.
Figure 2.6: Illustrated Ku-band TCDA unit cell, showing co-planar dipoles, balun, and Wilkinson power divider. Microstrip ($Z_0 \approx 100 \Omega$), stripline ($Z_0 \approx 20 \Omega$), and twinline ($Z_0 \approx 200 \Omega$) traces are used to equalize fabrication feature sizes despite a wide range in impedances.
Figure 2.7: Design parameters of a Ku-band TCDA utilizing the split unit cell approach. The four metal layers are displayed individually, and are combined according to the layer stack seen at the bottom. Optimized values are given in Table 2.1.
Table 2.1: Optimized parameter values of the split unit cell Ku-Band TCDA, shown in Fig. 2.7.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value (mm)</th>
<th>Parameter</th>
<th>Value (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>7.75</td>
<td>B</td>
<td>6.67</td>
</tr>
<tr>
<td>C</td>
<td>3.875</td>
<td>D</td>
<td>0.8</td>
</tr>
<tr>
<td>E</td>
<td>0.8</td>
<td>F</td>
<td>1.0</td>
</tr>
<tr>
<td>G</td>
<td>2.0</td>
<td>H</td>
<td>0.75</td>
</tr>
<tr>
<td>I</td>
<td>3.37</td>
<td>J</td>
<td>0.8</td>
</tr>
<tr>
<td>K</td>
<td>0.305</td>
<td>L</td>
<td>1.12</td>
</tr>
</tbody>
</table>

As in [7,10], the smallest feature observed in Fig. 2.7 is the input trace of the balun. However, the capacitance reducing techniques implemented here result in this feature being only 0.305 mm (12 mil), compared to 0.023 mm (0.889 mil) as would be required without this approach (see again Fig. 2.3). Critically, this increase in size of the smallest feature re-introduces low-cost PCB fabrication as a feasible option for the design and production of UWB antennas in this frequency range.

Using these values, the infinite array VSWR response is obtained via full-wave simulation in Ansys HFSS [59]. The VSWR is plotted for broadside, TE-mode (H-plane) scanning, and TM-mode (E-plane) scanning in Fig. 2.8. We observe that the array operates from 3–18 GHz for VSWR<2 at broadside, and VSWR<3 for scanning to 45° from broadside. In Fig. 2.9, we plot the radiation efficiency of the array for the same scan conditions. Radiation efficiency describes the ratio of radiated power to the total power available to the antenna, and therefore accounts for both reflected power at the input, as well as resistive and dielectric losses. Notably, the broadside efficiency remains above -1 dB (79%) across the entire 3–18 GHz band, and
Figure 2.8: Simulated matching of the infinite-array design shown in Fig. 2.7. We observe operation from 3–18 GHz for VSWR < 2 at broadside and scanning to ±45° in both planes for 3–17 GHz with VSWR < 3.

above -0.5 dB (89%) across 3–17 GHz. Conversely, the lowest radiation efficiency observed in Fig. 2.9 corresponds to the case of 45° scanning in the E-plane, where it reaches as low as -2.5 dB (56%). While this dip in efficiency does correspond to a peak in VSWR, as shown in Fig. 2.8, this alone does not explain the low efficiency. Specifically, while VSWR = 3 implies \( \Gamma = 0.5 \), this only corresponds to a mismatch loss \( (ML = -10 \log(1 - \Gamma^2)) \) of 1.25 dB. Thus the additional losses must have another source, specific to E-plane scanning.

### 2.2 Efficiency Improvement and Resonance Mitigation

The losses observed in Fig. 2.9 may be an instance of “scan blindness” due to surface waves [60], or may result from the resistive element found in the Wilkinson power divider in the feed network [57]. However, simply removing this component
Figure 2.9: Ratio of power radiated to power available, at broadside and 45° scans in the E- and H-planes. Losses up to 2.5 dB are observed for wide E-plane scans, which cannot be explained by mismatch alone.

from the feed introduces destructive resonances in the operating bandwidth. Therefore in order to improve the efficiency of the array, we must investigate more closely both the source of losses in the feeding network, as well as the excitation of spurious resonances.

2.2.1 Split Unit Cell

The pair of dipoles seen in Fig. 2.7 are each $\lambda/4$ in length, and fed in parallel through a Wilkinson power divider to form the $\lambda/2$ unit cell. The use of two, half-sized elements fed in-phase is referred to as a “split” unit cell. This approach borrows from [7, 10, 11], wherein it stems from the need to handle high impedances at the aperture. Manipulation of the element impedance through alteration of the array spacing is best understood using the framework laid out in [61].
Briefly, it is noted that the fields due to an infinite array of radiators are identical to the fields of a single such radiator, inside a waveguide with Perfect Electric Conductor (PEC) and Perfect Magnetic Conductor (PMC) boundaries. This configuration is shown in Fig. 2.10, superimposed with the virtual array resulting from the image currents on the waveguide walls. The virtual array spacing is equal to the waveguide boundaries, \( D_y = a \), and \( D_x = b \). Such a waveguide supports a TEM mode propagating perpendicularly to the plane of the virtual array, and has a purely real wave impedance given by

\[
R_{\text{wg}} = \eta b/a, \quad (2.2)
\]

where \( \eta = \sqrt{\mu/\epsilon} \approx 377 \, \Omega \) is the free space impedance. An antenna receiving from this waveguide acts as a transformer \([61]\), producing a voltage at the terminals proportional to its effective height \((h)\). It therefore encounters a scaled radiation resistance,

\[
R = \eta \frac{b}{a} \left( \frac{h}{b} \right)^2 = \eta h^2/ab. \quad (2.3)
\]

By reciprocity, (2.3) is true of the transmit case as well. We note, this formulation assumes the nominal waveguide is fed from one end, and thus extends in a single direction. Practically, this implies the array is backed by PMC, or otherwise PEC at a distance \( \lambda/4 \). In the case of bidirectional radiation, two such waveguides are fed in parallel, extending in opposite directions, such that the radiation resistance in (2.3) is halved.

Notably, for a wire of length \( l \) with uniform current distribution, \( h = l \). Therefore, a uniform current wire spanning the waveguide length \( b \) logically has an effective length \( h = b \), resulting in a radiation resistance \( R = 377 \, \Omega \). Conversely, a resonant half-wavelength dipole has an effective length \( h = 2l/\pi \); for an array spacing of
Figure 2.10: Schematic layout of a linearly polarized radiator situated inside a waveguide cross-section. The boundary conditions at the PEC and PMC walls match those of an infinite array of identical and aligned radiators.

\[ a = b = 0.5\lambda, \text{ this produces a radiation resistance } R = 153\,\Omega. \] Thus, for the quasi-constant current distribution of a TCDA, we expect an \( R \) value between these two (i.e. \( 153\,\Omega < R_{\text{TCDA}} < 377\,\Omega \)). Empirical experience has found this value to be \( \approx 188\,\Omega \) \[7,10,11\].

As was noted before, this high input impedance to the tightly coupled dipole element is challenging to match to 50\,\Omega. However, from (2.2) we can see immediately that altering the ratio \( b/a \) can decrease this impedance. Significantly increasing \( a \) is not feasible due to the resultant onset of grating lobes in the H-plane. Thus, the split unit cell approach halves the value of \( b \), by placing two quarter-wavelength radiators inside each \( 0.5\lambda \) unit cell. Specifically, halving the length of the periodic cell in the direction of current serves to reduce by half the Floquet mode impedance \[48\] (i.e., \( 94\,\Omega \)). By feeding these two miniaturized cells in parallel, the input impedance
is further reduced to 47 Ω and is easily matched to the 50 Ω system impedance. This formulation is only approximate, and assumes that \( h \) scales linearly, and that the miniaturized elements are not significantly impacted by increased mutual coupling. Nonetheless, it serves as a useful qualitative tool.

Another useful insight stemming from this mode of analysis is an understanding of the array impedance when scanning. As shown in Fig. 2.11, the dimensions of the conceptual waveguide are distorted, as the aperture area is projected into the scanning direction. We can see that the length of the aperture is reduced by a factor \( \cos \theta \), in the scanning direction. Adjusting this to account also for azimuth angle, we can modify (2.2) to state more generally

\[
R_{wg} = \eta \frac{b \sin \phi \cos \theta}{a \cos \phi \cos \theta},
\]

where \( \phi = 0^\circ \) is defined as parallel to the primary current of the radiator (i.e. E-Plane). As such, we observe that the wave impedance is reduced as the array is scanned in the E-plane, but increases as it is scanned in the H-plane (\( \phi = 90^\circ \)).

It should be noted, that because both \( \lambda/4 \) elements of the split unit cell are fed from a common source, the two-element subarray can be treated as a single antenna with phase center between the elements. These subarrays are spaced at \( \lambda/2 \) intervals. Therefore, this approach does not impact the calculation of grating lobes, nor result in an increase in the total number of signal chains in the array, implying no additional backend hardware requirements. Further, while the signal reflected from the two elements remains in-phase (as is the case at broadside), virtually no power is lost in the power divider’s resistors. However, when scanning in the E-plane, these reflections fall out of phase, and losses up to 1 dB have been observed [10]. At higher frequencies, where link budget is already strained, these additional losses are undesirable. With
Figure 2.11: Conceptual diagram of an array radiating at an angle into a waveguide, such as in Fig. 2.10. The waveguide width at broadside \(d_1\) is reduced when projected into the scan direction \(d_2 = d_1 \cos \theta\). This results in an altered impedance, looking into the array.

This in mind, we seek to eliminate the split unit cell and associated power divider, and implement a half-wavelength \((\lambda/2)\) unit cell.

### 2.2.2 Surface Waves

Surface waves can exist in many geometries of dielectric interfaces. In antenna arrays these can be excited at broadside, but are much more commonly observed when scanning (sometimes referred to as “scan blindness”). Though detrimental in the context of surface transmission lines or antenna arrays, surface waves can be used to form purely dielectric waveguides. Indeed, to understand the onset of surface waves in an array setting, it is useful to initially consider a simple waveguide consisting of a grounded dielectric slab, as shown in Fig. 2.12.

The configuration in Fig. 2.12 supports guided TE and TM wave modes, propagating as \(e^{-j\beta z}\). The derivation by Pozar [36] of the TM mode yields the pair of
Figure 2.12: Geometry of a dielectric slab above a groundplane. The dielectric region is assumed to be isotropic, and infinite in \( y \) and \( z \). Orientations of the guided TM and TE modes are shown.

Transcendental equations:

\[
(k_c d)^2 + (h d)^2 = (\epsilon_r - 1)(k_0 d)^2, \tag{2.5}
\]

\[
k_c d \tan(k_c d) = \epsilon_r h d, \tag{2.6}
\]

where \( h \) is the rate of attenuation of the evanescent wave in the air region (thus must be \( > 0 \) for a physical solution), \( k_c \) is the cutoff wavenumber in the dielectric, and \( k_0 \) is the free-space wavenumber (\( k_0 = 2\pi/\lambda_0 \)). Derivation for the TE modes yields a similar pair of equations:

\[
(k_c d)^2 + (h d)^2 = (\epsilon_r - 1)(k_0 d)^2, \tag{2.7}
\]

\[
-k_c d \cot(k_c d) = h d. \tag{2.8}
\]

These equations must be solved numerically and are plotted for reference in Fig. 2.13. Equations (2.5) and (2.7) represent circles in the \( k_c d, h d \) plane, having a radius,

\[
r = \sqrt{\epsilon_r - 1} k_0 d. \tag{2.9}
\]
However, this value takes on a more intuitive meaning when put in terms of the guided wavelength, $\lambda_g$:

$$\lambda_g = \frac{\lambda_0}{\sqrt{\varepsilon_r}} = \frac{2\pi}{k_0\sqrt{\varepsilon_r}},$$

which is rewritten as

$$k_0 = \frac{2\pi}{\lambda_g\sqrt{\varepsilon_r}},$$

and inserted into (2.9) to yield

$$r = \frac{\sqrt{\varepsilon_r - 1}}{\sqrt{\varepsilon_r}} \frac{d}{2\pi \lambda_g}.$$  \hspace{1cm} (2.10)

These circles are plotted for various values of $d/\lambda_g$ in Fig. 2.13. Intersection of a circle with the plotted curves represents a valid solution describing a propagating mode. Where multiple tangent curves intersect a circle, the dielectric slab represented by that circle supports multiple modes simultaneously. The propagation factor of the guided wave ($\beta_{sw}$) can be determined from the value of $k_c$ extracted from Fig. 2.13, by

$$k_c^2 = \frac{\varepsilon_r k_0^2}{\lambda_g} - \beta_{sw}^2.$$  \hspace{1cm} (2.11)

Careful observation of Fig. 2.13 yields several useful insights. Notably as the central TM mode includes the origin, any slab of thickness $d > 0$ and $\varepsilon_r > 1$ supports at least this fundamental mode. As the slab thickness increases, the supported modes progress as: TM$_0$, TE$_1$, TM$_1$, ... TE$_n$, TM$_n$. Further, as the derivative of (2.6) and (2.8) along the $+k_c d$ axis is always positive, and the derivative of the circles (2.5) and (2.7) always negative, the earliest intersection of each new mode occurs as $hd \rightarrow 0$.  

40
Figure 2.13: Contour plots of equations (2.5)–(2.8) in the $k_c d, h d$ plane, for slabs of dielectric constant (a) $\epsilon_r = 2$, (b) $\epsilon_r = 8$. Intersections of the curves with the circles represent solutions to the equation pairs. The solid curves correspond to TM modes, whereas dashed curves correspond to TE modes, and the dotted curves represent dielectric slabs of varying thickness.
As such the cutoff frequencies of any TM$_n$ and TE$_n$ modes can be calculated as:

$$f_{c,TM} = \frac{nc}{2d/\sqrt{\epsilon_r} - \frac{1}{2}}, \quad n = 0, 1, 2, ... \quad (2.12)$$

$$f_{c,TE} = \frac{(2n-1)c}{4d/\sqrt{\epsilon_r} - \frac{1}{2}}, \quad n = 1, 2, 3, ... \quad (2.13)$$

Clearly, surface waves can be mitigated to a certain extent by using thin or low dielectric constant substrate and superstrate materials. However, as was discussed in Section 1.2, even thin UWB arrays such as the TCDA have a thickness above the groundplane of $\approx \lambda_g/2$. It should also be noted that in an array setting, the additional loading of the antenna elements and feed structures will produce a slightly higher effective dielectric constant ($\epsilon_{eff} > \epsilon_r$) [60,62,63]. Thus from (2.12) and (2.13), or alternatively from Fig. 2.13, we expect to contend with two potential surface wave modes: TM$_0$ and TE$_1$.

Critically, the existence of these propagating modes alone does not inherently mean they will become excited in the array. However when excited, energy becomes trapped in the substrate with little or no radiation, resulting in complete reflection at the antenna feed. This condition occurs when the propagation factor of the radiated wave inside the substrate ($\beta_{sub}$) matches the propagation factor of a guided wave mode of the same polarization. For an array geometry such as that shown in Fig. 2.14, the substrate propagation factor is derived in [60] and is calculated as

$$\beta_{sub}^2 = k_x^2 + k_y^2,$$

wherein the tangential wavenumbers $k_x, k_y$ are found to be

$$k_x = \left(\frac{2\pi m}{a} + k_0u\right) \quad \text{(2.14a)}$$

$$k_y = \left(\frac{2\pi n}{b} + k_0v\right) \quad \text{(2.14b)}$$
In (2.14), \( m \) and \( n \) are positive or negative integers corresponding to the Floquet mode (see again Section 2.2.1), \( a, b \) are the element spacing as shown in Fig. 2.14, and \( u, v \) correspond to the scan angle of the array:

\[
\begin{align}
    u &= \sin(\theta) \cos(\phi), \\
    v &= \sin(\theta) \sin(\phi).
\end{align}
\]

Therefore, we can write the condition of surface wave resonance as

\[
\left( \frac{\beta_{sw}}{k_0} \right)^2 = \left( \frac{m\lambda_0}{a} + u \right)^2 + \left( \frac{n\lambda_0}{b} + v \right)^2,
\]

noting again that the polarization condition requires \( u \neq 0 \) for TM modes and \( v \neq 0 \) for TE modes, with the exception \( u = v = 0 \).
As was discussed before, the surface wave propagation factor is dependent on the thickness and dielectric constant of the substrate. Conversely, (2.16) demonstrates that the propagation factor of the antenna fields in the substrate is dependent only on the element spacing and scan angle. The appropriate value of $\beta_{sw}$, can be determined from (2.11) in conjunction with Fig. 2.13. For convenience, the value of $\beta_{sw}$ has been solved numerically as a continuous function of $d$, and is given in Fig. 2.15. Logically, we see that as the substrate thickness increases, $\beta_{sw}$ asymptotically approaches the propagation factor of a plane wave in dielectric, $k_0\sqrt{\epsilon_r}$. We observe also that increasing the dielectric constant brings the TM$_n$ and TE$_{n+1}$ modes closer together. Due to the factor $\epsilon_r$ in (2.6), it is in principle possible for the TE mode propagation factor to exceed that of the TM mode. However, this will only occur in the case of very high dielectric constant (e.g. $\epsilon_r > 30$) and slab thickness (e.g. $d/\lambda_0 > 1$).

Taking as example an array of thickness 0.5$\lambda_g$ with $\epsilon_r = 2$ substrate, Fig. 2.15 yields surface wave propagation factors of 1.3 (TM$_0$) and 1.09 (TE$_1$). In (2.16), these values represent the radii of circles in the $u,v$ plane, as can be seen in Fig. 2.16. The centers of these circles are offset by the Floquet mode numbers ($m,n$) multiplied by the element spacing ($a/\lambda_0$ and $b/\lambda_0$). These surface wave modes will become excited and severely mismatch the array where the plotted circles enter the visible region of the array ($u^2 + v^2 < 1$). Immediately, we see that as $\beta_{sw}$ increases, the onset of surface waves moves inwards towards broadside, occurring at decreasing values of $\theta$. Likewise, increasing the element spacing beyond 0.5$\lambda_0$ draws the center points of the surface wave circles inwards, also exacerbating the onset of these modes.

The visible region in Fig. 2.16 is plotted in greater detail in Fig. 2.17, in order to predict the specific state in which the surface waves will occur. While these will occur
Figure 2.15: Numerical solutions to the surface wave propagation factor ($\beta_{sw}$) as a function of substrate thickness, for (a) $\epsilon_r = 2$, and (b) $\epsilon_r = 8$. The first three TE and TM modes are labeled.
Figure 2.16: Solutions of (2.16) plotted in one quadrant of the $u, v$ plane, for an array spacing of $a = b = 0.5\lambda_0$, on $\epsilon_r = 2$ dielectric of thickness $d = 0.5\lambda_g$. Portions of the TE and TM circles which fall inside the visible region ($u^2 + v^2 < 1$) represent scan conditions at which those surface wave modes will become excited.
Figure 2.17: Detail of the visible region from Fig. 2.16, showing azimuth ($\phi$) and elevation ($\theta$) grids. TM$_0$ and TE$_1$ surface wave modes are plotted (solid and dashed curves, respectively), and will be excited at 44.4° in the E-plane ($\phi = 0^\circ$) and 65.5° in the H-plane ($\phi = 90^\circ$). Results are symmetric about $u$ and $v$ (i.e. for negative values of $\phi, \theta$).

for continuous values of $\phi$, we consider for simplicity the cardinal scan planes, E-plane or TM mode radiation ($\phi = 0^\circ$) and H-plane or TE mode radiation ($\phi = 90^\circ$). We observe the TM$_0$ curve reaches the closest to the origin (broadside) as it intersects the $u$ axis, corresponding to a $\theta = 44.4^\circ$ scan angle in the E-plane. The intersection of the same curve with the $v$ axis will not resonate, as the polarizations do not match. As such, the earliest surface wave resonance in the H-plane occurs with the TE$_1$ mode, at 65.5°.

This trend can be generalized to state that for any thin and low dielectric constant array, the earliest surface waves will occur in the E-plane, at wide scan angles. Further, as the electrical thickness of the substrate in (2.10) and the element spacing
in (2.16) both shrink as the frequency is decreased, we can further state that surface waves will be observed earliest at the high end of the operating bandwidth. We note however, that surface waves at very shallow scan angles, or even broadside, do not require extreme dielectrics to become excited. For an array spaced at $0.5\lambda_0$, broadside surface waves can be excited for a value of $\beta_{sw}/k_0 = 2$, which corresponds to the relatively common dielectric constant of $\epsilon_r \approx 4$. Thus, with the exception of very thin arrays ($d < 0.25\lambda_g$) which are typically narrowband, substrate material selection is the most limiting factor with regard to scanning range in UWB arrays.

### 2.2.3 Cavity Resonance

Unlike surface waves, cavity resonances are primarily excited at broadside and dissipate when scanning. These are caused by a net current in the balanced feed of the dipole, compounded by the presence of a resonant cavity within the geometry of the array. At resonance, this induces strong vertical fields beneath the aperture which ultimately overwhelm the balanced currents of the differential feed. For this reason, it is sometimes referred to as a “common-mode” resonance. This type of resonance is characteristic of unbalanced-fed antennas [34, 51], and will not occur in an ideal, balanced excitation. However even with a balun, this condition cannot be guaranteed, and an imbalance in the currents on the differential feed will produce a net vertical current, as shown in Fig. 2.18.

The unbalanced current causes vertical fields above the groundplane, which under the correct conditions can resonate. In particular, the repeating grounded conductors present in the array baluns form a cavity. At such frequencies that these grounded posts are separated by $\lambda/2$, the energy fed into the array is stored in these resonant
cavities, and is not radiated. This condition is illustrated in Fig. 2.19. Notably, scanning the array introduces a relative phase difference between the neighboring elements, weakening and ultimately preventing this resonance. Thus, cavity resonance is seen only at or near broadside.

As was demonstrated in [34], this resonance can be relatively accurately predicted based on the separation of the grounded posts. In Fig. 2.20 we consider a simple linearly polarized array, as viewed from above. If the dipole feeds are imbalanced, as noted above, a cavity resonance is excited along the diagonal length indicated as $L$. Resonance will occur when this length is a half wavelength:

\[
f_r = c \left( \frac{1}{2L\sqrt{\varepsilon_{\text{eff}}}} \right) = \frac{c}{2\sqrt{D_x^2 + D_y^2\sqrt{\varepsilon_{\text{eff}}}}}.
\]

(2.17)
Figure 2.19: Strong vertical fields at resonance induce common-mode currents in the normally differential feed line. Notably, such resonance will more commonly occur diagonally between rows, rather than in-line as shown here.

We note that the value $D_x$ is slightly less than the element spacing (i.e. $a$ in Fig. 2.14) due to the width of the feed. Likewise, $\epsilon_{eff}$ is the effective dielectric constant, and can be approximated for the setup in Fig. 2.20 by a simple volumetric average:

$$\epsilon_{eff} = 1 + \frac{d}{D_y}(\epsilon_r - 1). \quad (2.18)$$

For the array design shown in Fig. 2.7, the expected cavity resonance can be calculated from (2.17). The resonant length is found to be $L = 7.54$ mm, with $\epsilon_{eff} = 1.05$, resulting in the value $f_r = 20.88$ GHz. Therefore this resonance occurs beyond the operating band of the array. However in eliminating the split unit cell feed network to improve efficiency, as discussed in Section 2.2.1, the E-plane element spacing is doubled, resulting in a decrease in the cavity resonance frequency. As a simple example we consider all the same dimensions as in Fig. 2.7, but with every other balun removed. This results in $L = 9.127$ mm, $\epsilon_{eff} = 1.05$, and thus an expected resonance around $f_r = 15.65$ GHz. Indeed, we can see in Fig. 2.21 that a
Based on (2.17), it is clear that the undesired cavity resonance can be pushed out of band by altering the resonant length, $L$. With the use of a split unit cell, this stemmed naturally from a much smaller distance between neighboring baluns. However, a similar effect can be achieved without a split unit cell by introducing a grounded shorting pin between neighboring elements. These run vertically from near the aperture (though not touching the dipoles) to the groundplane, as shown in Fig. 2.22. The response of the modified array, with the inclusion of shorting pins, is also shown in Fig. 2.21 (second peak), and is seen to increase the resonant frequency. However, the new resonant frequency (16.5 GHz) has only slightly increased and still falls within the passband. This is the case because while $D_x$ in (2.17) has decreased, $D_y$ remains constant. The resonant frequency can be further increased, and pushed out of the band, by additionally increasing the collar dimension noted as
Figure 2.21: Infinite array input reflection, showing initial spurious cavity resonance (dashed peak at 15 GHz). Introduction of a shorting post shifts the resonance (dashed peak at 16.5 GHz), and finally increasing the dimension ‘A’ (see Fig. 2.22) pushes the resonance out of the operating band (solid curve).

‘A’ in Fig. 2.23. Including this modification, the solid curve in Fig. 2.21 demonstrates a resonance-free band, without the use of a split unit cell.

2.3 Optimized Performance and Scalability

With the above considerations, the initial design from Section 2.1 was modified to eliminate the split unit cell and corresponding wilkinson power divider, as well as to mitigate the onset of cavity resonance at broadside and surface waves when scanning. The final design layout is shown in Fig. 2.23, with key design parameters labeled. The optimized values for the parameters shown in Fig. 2.23 are given in Table 2.2. Rogers Duroid 5880 substrates ($\varepsilon_r = 2.2$, 125 $\mu$m or 5 mil) are used for $Sub_{I-II}$ and $Sub_{III-IV}$, with Polyflon bond ($\varepsilon_r = 2.32$, 100 $\mu$m or 4 mil) as $Sub_{II-III}$. A superstrate of High-Density Polyethylene (HDPE, $\varepsilon_r = 2.3$) is included to improve
Figure 2.22: Physical representation of the parameter $D_x$ in Fig. 2.20. Addition of a shorting post significantly reduces this value (labeled $D'_x$), as does adjustment of the dimension ‘A’.

scanning performance, and is marked as ‘X’ in Fig. 2.23. The element spacing, ‘A’, corresponds to 92.5% of the nominal $0.5\lambda$ grating lobe separation, implying some oversampling of the array area. The presented data assumes a single-polarized array, although extension to dual polarizations can be accomplished using the interlocking, dual-offset approach demonstrated in [7,10], without significant re-optimization.

Impedance matching was determined through full-wave simulation of the infinite array at broadside and $\pm 45^\circ$ in the E- and H-planes, and are given in Fig. 2.24. We note that this design achieves a 5.3:1 bandwidth, from 3.5 GHz–18.5 GHz, with VSWR < 2 at broadside and in the E-plane. While scanning in the H-plane, VSWR increases to 2.6, and is thus limited to $\pm 45^\circ$. Conversely, Fig. 2.25 shows that the design preserves up to 88% of the broadside bandwidth (from 4 GHz–17.25 GHz) when scanning down to $\pm 70^\circ$ from boresight in the E-plane, with VSWR < 2.2. This
Figure 2.23: Illustration of the proposed array periodic unit cell. Design parameters are noted as ‘A’−‘X,’ and three transmission line elements are identified. The design as shown has a 90Ω input impedance, and requires tapering to operate in a 50Ω system.
behavior is expected in dipole arrays, based on the variation in the Floquet mode impedance in both planes, as was discussed in Section 2.2.1.

The VSWR values in Figs. 2.24 and 2.25 assume the balun is nominally fed from a 90 Ω source, which may not be available in practice. As such, Fig. 2.24 also shows the simulated performance of the array when fed from a 50 Ω source, with a simple linear taper. Notably for a taper only one wavelength in length at the highest frequency of operation (\(\lambda_{hi}\)), the array performance approaches that of the nominal, 90 Ω feed. Clearly, this additional length can be meandered to minimize wasted space beneath the groundplane, and ultimately is dominated by the size of the physical input port, as will be discussed in the following section.
Figure 2.24: Simulated infinite array VSWR of the design in Fig. 2.23, using the values in Table 2.2. Broadside (red), and ±45° scanning in the E-plane (black) and H-plane (blue) are shown. Solid curves refer to a 90 Ω impedance match, whereas dashed curves are matched to 50 Ω using a taper of length $\lambda_{hi}$.

Figure 2.25: Infinite array VSWR for scans in the E-plane at ±50° (orange), ±60° (green), and ±70° (violet). Notably, the array maintains 88% of the broadside bandwidth down to ±70°, with VSWR < 2.2. Scanning beyond ±45° is achievable only in the E-plane.
Returning to the optimized values in Table 2.2, we note that the smallest design dimension is 178 μm (7 mil). This is quite generous compared to previous TCDAs [7,10,11], as well as the state of the practice for standard, multilayer PCB fabrication (75–100 μm or 3–4 mil). As such, this design is well suited for scaling to higher frequencies. Indeed, this is demonstrated by linearly scaling this design by reducing all dimensions in Table 2.2 by a factor of 2 (except ‘U’, which is set to 4 mil or 100 μm).

Critically, this brings the array’s operation into the X–Ka bands. As shown in Fig. 2.26, the scaled design operates from 7 GHz–37 GHz with VSWR < 2.1 at broadside and ±45° in the E-plane, and VSWR < 3 at ±45° in the H-plane. As expected, the matching behavior of Fig. 2.26 closely resembles that of Fig. 2.24. Minor variations between the two are a result of the nonlinear scaling in substrate dimensions. Particularly, the results shown in Fig. 2.26 account for a layer stack of Duroid 5880 (ε_r = 2.2, 5 mil thick), PolyFlon bond (ε_r = 2.32, 2 mil thick), and Duroid 5880 (5 mil thick).

Continuing this scaling down to the PCB fabrication limit (i.e. 3 mil minimum feature size), the array reaches well into the millimeter-wave band. As shown in Fig. 2.27, similar performance can be achieved at 9 GHz–49 GHz (upper X-band into Millimeter-Wave), by reducing the dimensions in Table 2.2 by a factor of 2.67, and setting ‘U’ to 75 μm (3 mil). This represents the state of the practice for PCB fabrication [64], and accounts for a layer stack of CuFlon (ε_r = 2.1, 4 mil thick), PolyFlon bond (2 mil thick), and CuFlon (4 mil thick). In both Figs. 2.26 and 2.27, the simulated arrays are spaced at 92.5% of the grating lobe separation. Critically, the scalable performance demonstrates the versatility of the present TCDA design.
Figure 2.26: Infinite array impedance match of the proposed Ka-band design. Broadside (red), and \( \pm 45^\circ \) scanning in E (black) and H (blue) planes are shown. This design can be fabricated with a minimum feature size of 4 mil.

Notably, operation up to 49 GHz as observed in Fig. 2.27 represents a nearly tenfold increase in operating frequency over previous works, without alteration of the fabrication method.

### 2.4 Prototype Fabrication

The design operating from 3.5–18.5 GHz was validated by fabrication of an \( 8 \times 8 \) prototype, according to the dimensions given in Table 2.2. The assembly procedure of the array prototype is illustrated in Fig. 2.28. The prototype consists of 8 vertical PCBs, each containing a linear array of 8 identical elements. Each element is fed separately via an edge-launch coax port, accessible beneath the groundplane. These subarray cards are inserted, from below, through slots in the groundplane. Thereafter, dielectric spacers and superstrate can be inserted between the subarray cards.
Each component of the prototype seen in Fig. 2.28 was commercially sourced, and the final array assembled by hand in the laboratory. The layout and composition of the antenna PCB is discussed in Section 2.3. The unit-cell design shown in Fig. 2.23 is supplemented with a meandered microstrip taper to transform the input impedance, and the footprint for coax port, as can be seen in Fig. 2.29. The impedance taper takes up space beneath the groundplane, but is similar in size to the former Wilkinson divider, and ultimately is small in comparison to the coax port itself. The identical unit cell, balun, and feed is repeated for all 64 elements. Finally, the exterior copper layers are plated using the Electroless Nickel Immersion Gold (ENIG) process to prevent oxidation. This is preferable to cheaper solder-coating (HASL) which forms a thicker coating, and can lead to parasitic capacitance at high frequencies.
Figure 2.28: Assembly diagram of the various components of the 8×8 array prototype. Detail of the 1 × 8 subarray assembly is shown below.
In the array prototype, each element is fed from an edge-launch 50Ω coaxial, Sub-Miniature Push-to-connect (SMP) port. The SMP format supports the desired frequency range, and moreover is small enough (4 mm width) to fit between each element. We note, the SMP ports represent a large component of the total size, weight, and cost of the complete array; however, in an application setting it is easily possible to integrate electronics on the same PCB board as the antenna feed, eliminating the need for many coax ports.

Once the coax ports are attached, the antenna PCBs are inserted through the groundplane, which is a laser-cut steel foil (225 μm or 9 mil thickness). A partially assembled array showing the groundplane slots can be seen in Fig. 2.30. The laser
cutting process is low-cost and commercially available at high scale due to its use in the production of stencils. Steel was used as it was readily available in this process, but it is expected higher performance may be achieved by using aluminum, which has higher conductivity ($\sigma_{al} = 36.9 \times 10^6$ SI/m as opposed to $\sigma_{st} = 1.3 \times 10^6$ SI/m). The PCBs are reinforced, and the superstrate held in place, by rigid spacers. These were custom manufactured using Stereo Lithography (SLA, a form of “3D printing”), and were designed to be > 95% air by volume. Dielectric properties of the proprietary resin are unknown, but an effective dielectric constant of the complete structure, $\epsilon_{eff} < 1.05$, is expected based on the predominantly air volume. When better characterization of these materials is available, it is expected that the superstrate can be directly co-fabricated with the spacers, which would simplify the assembly process, and reduce the possibility of error. The assembled array, with superstrate removed, can be seen in Fig. 2.31, and the complete prototype with superstrate is shown in Fig. 2.32.

Antennas intended for space applications require special consideration with regard to the selection of suitable materials. Although a nuanced topic, it is beneficial to highlight two primary concerns: volatile outgassing, and thermal cycling, for which additional detail can be found in [56,65]. Briefly, outgassing is the release of organic and other matter from a material due to vacuum and high temperature conditions, and can cause contamination of optics, solar panels, and thermal radiators. Suitable materials are generally expected to experience < 1% in Total Material Loss (TML) due to outgassing, with < 0.1% Collected Volatile Condensable Material (CVCM) [66]. Likewise, thermal cycling refers to the repeated temperature extremes encountered in space that can be typically in the range -190–160° C [56]. Continual expansion and contraction can cause fracturing of surface metalization, and particularly vias
Figure 2.30: Photograph of the prototype during assembly. Printed subarrays can be seen (black) with shaped dielectric spacers (white), as well as remaining slots in the groundplane. The superstrate is added last, above the spacers.
Figure 2.31: Photograph of the assembled prototype array, shown with superstrate removed. The 3D-printed spacers can be seen in white, and the fabricated antenna boards in black.
Figure 2.32: Photograph of the assembled array and groundplane. The HDPE superstrate is composed of nine separate strips which lie between the antenna boards, as indicated in Fig. 2.28. Future iterations of the array could integrate the superstrate and spacers into a single component to reduce assembly burden and possibility of error.
or interconnects, and is screened based on the through-plane Coefficient of Thermal Expansion (CTE$_z$).

The primary dielectric used in this design is Roger’s Duroid 5880, which has been previously tested for these considerations. It has TML of 0.03% and CVCM < 0.01% [67], with a through-plane CTE of 237 ppm/$^\circ$C [68]. The CTE is considered high and may pose a long term risk to the durability of vias. Alternate substrate materials include Duroid 5880LZ which has similar dielectric properties, but reduced CTE$_z$ of 41.5 ppm/$^\circ$C, or ceramic based materials such as Rogers 6002 ($\epsilon_r$ = 2.94, and CTE$_z$ = 24 ppm/$^\circ$C).

2.5 Measurement

The array prototype was measured at the NASA Glenn Research Center far-field range (see Fig. 2.33). For measurement, the prototype was inserted in a ground plane extending at least $2\lambda_{low}$ around the periphery of the array (where $\lambda_{low}$ indicates the wavelength at the lowest frequency of operation). Measurements were conducted with the antenna in a receive configuration, following the Unit Excitation Active Element Pattern (UEAEP) method [69]. In this method, individual ports are measured sequentially with all others terminated at a matched load, with the beam patterns generated via post-processing. Finally, the system setup was calibrated against a standard gain horn, with attenuation in the SMA-SMP converter compensated according to the datasheet [70].

Measured and simulated beam patterns in the E and H planes at 5.2 GHz, 8.4 GHz, 13 GHz, and 17.5 GHz are presented in Fig. 2.34. These patterns assume a uniform amplitude excitation and are normalized to show the sidelobe level, which is observed
to fall under -10 dB in all cases, and even under -15 dB at the higher frequencies. The measured and simulated results are in close agreement. Deterioration of the beam pattern is observed in the E-plane at lower frequencies, and is expected due to the small size of the array (1.1λ and 1.8λ at 5.2 GHz and 8.4 GHz, respectively). Ripple in the beam pattern is caused by reflection at the edge of the array, of the currents travelling along the coupled dipoles. For this reason, the H-plane patterns remain smooth. We note that this effect is minimized as the array size is increased.

Measured co- and cross-polarized gain (Ludwig’s third definition [71]) is plotted alongside the calculated reference directivity [72] in Fig. 2.35. We note that the measured gain tracks the reference directivity within 2 dB. We also note that the cross polarized gain is -30 dB in the upper half of the band, but peaks to -15 dB
Figure 2.34: Measured scanning patterns of the prototype array in the E-plane (left, red) and H-plane (right, blue), for 5.2 GHz, 8.4 GHz, 13 GHz, and 17.5 GHz (top to bottom). Simulated patterns (dashed black) are shown in each plot for reference.
around 7.5 GHz. This peak in the cross-pol is unexpected for this type of array at broadside.

Upon further investigation, it was found that weak contact at the cutouts in the groundplane (through which the antenna PCBs are inserted) allowed the long cutout to radiate as a low frequency slot antenna (see Fig. 2.36). As the electric fields in this type of antenna are orthogonal to the length of the slot, the resultant radiation is cross-polarized to the primary fields of the dipoles. Introducing this nonideality into the model, it is shown in Fig. 2.37 that the measured phenomenon can be accurately recreated in simulation. Further we demonstrate that by breaking up the long slot, and thereby increasing the slot-resonance frequency, the peak cross-pol is reduced by as much as 10 dB and 18 dB at 7.5 GHz and 3 GHz, respectively.
Figure 2.36: Illustration of a continuous slot cut into the groundplane for inserting the antenna PCB (top). This structure can radiate when excited by the array feed. The suggested implementation (bottom) breaks up the long slot to prevent in-band radiation. The electric field orientation of co- and cross-polarized radiation is indicated for reference.

Figure 2.37: Measured peak in cross-polarized gain (dashed) vs. simulations that include the long slot in groundplane (crosses) show good agreement. Concurrently, we show that by breaking up the slot (circles) we observe 10 dB improvement at 7.5 GHz, and 18 dB improvement at 3 GHz.
2.6 Thermal Analysis

For bidirectional networks, we must consider transmit and receive arrays. Indeed, satellites in Low Earth Orbit (LEO) and Geostationary Orbit (GEO) can be expected to transmit on the order of 200–500 W to close a link [56]. This scenario requires additional considerations as Ohmic and dielectric losses in the antenna are manifested as heat. In the vacuum, this energy is not easily dissipated and can lead to excessive thermal buildup. For this reason, multi-physics simulations of the fabricated array design were carried out in COMSOL. As this analysis assumes a large (>100 element) array, we are concerned with RF input powers up to 2 W per element. Heat is generated from dielectric losses in the substrate and superstrate (\(\tan\delta = 0.0009\) and 0.00031, respectively) and Ohmic loss due to the finite conductivity of copper (\(\sigma = 5.96 \times 10^7\) S/m). For our analysis, exterior copper layers were 35 \(\mu\)m thick (1 oz/ft\(^2\)) and interior layers were 17 \(\mu\)m thick (0.5 oz/ft\(^2\)). But as the primary current carrying surfaces reside on the exterior layers, increasing the interior copper thickness was found to have negligible effect.

It is useful to first observe the temperature distribution of the array assuming even convection on all surfaces (i.e. in-atmosphere). This allows us to observe the relative intensity of local heat sources, as is shown in Fig. 2.38 on the left, for a convective heat transfer coefficient \(h_0 = 5\) W/(m\(^2\)K). We observe that the primary heat sources fall along the signal traces below the groundplane and at the input to the balun, where field strength and current are at a maximum.

However in space we cannot expect convective cooling. Therefore, all boundaries are considered thermally insulated with black-body radiation according to the Stefan-Boltzmann law. We also include a weak heat sink with total thermal resistance of
Figure 2.38: Steady state thermal distributions (white-hot convention) of an infinite array, assuming 1 W RF input power at 17.5 GHz. High atmosphere conditions (even convection across structure with $h_0 = 5$) is shown on the left. Also, thermal distribution in vacuum conditions (no convection) with a $8.3\degree K/mW$ heat sink below the groundplane is shown on the right.
8.3°K/mW, attached to the substrate beneath the groundplane (to minimize impact on the antenna’s radiation), which represents heat transfer to the satellite bus. The steady-state thermal distribution for a single frequency excitation in this scenario is shown in Fig. 2.38, on the right. We note that as the antenna balun has large, continuous metal structures near the primary heat sources, the generated heat is efficiently conducted to the heat sink. Nonetheless, thermal accumulation occurs in the superstrate where conductance is limited, causing an inversion of the relative temperature profile as compared to the case of even convection.

While the relative thermal distribution is consistent across various frequencies of operation, the peak and average temperatures vary according to the power spectral distribution, as shown in Fig. 2.39. Likewise, the peak observed steady-state temperature was found to correlate linearly with input power, as seen in Fig. 2.40. These thermal distributions are not arrived at instantaneously, however. The transient response of the unit cell for the first 300 s of continuous operation is shown in Fig. 2.41.

2.7 Conclusion

In this chapter, we develop several TCDA designs to achieve greater than 5:1 bandwidths up to the Ku- and Ka-bands. These frequency bands represent an important intermediate step towards millimeter-wave operation, and are of high relevance to NASA’s space network architecture. Importantly, this was accomplished while maintaining tolerances for low-cost PCB fabrication techniques. The first design presented operates from 3.5–18.5 GHz, with VSWR < 2 at broadside and ±45° in the E-plane and VSWR < 2.6 at ±45° in the H-plane. It was shown that this design maintains 88% of the broadside bandwidth for scans as low as ±70° in the E-plane (VSWR < 2.2).
Figure 2.39: Observed thermal behavior during simulated high-power transmit operation, showing peak and volumetric average temperature change ($\Delta T$) above ambient for 1 W RF input power at individual frequencies.

Figure 2.40: Peak localized temperature change ($\Delta T$) above ambient as the RF input power is varied. The array was operated at 17.5 GHz, corresponding to maximal losses, as shown in Fig. 2.39.
Figure 2.41: Transient response of the unit cell showing instantaneous peak temperature, assessed at 17.5 GHz for 1 W RF input power.

Scalability of this design is demonstrated with further designs which operate from 7–37 GHz and 9–49 GHz, both for VSWR < 2.1 at broadside.

The Ku-band design was validated through fabrication of an 8 × 8 prototype. Broadside gain and scanning patterns were presented. These measurements showed close agreement with simulated results. Preliminary screening of materials for space applications was also reviewed. It is noted that the design substrate (Duroid 5880) is suitable, on account of low outgassing and thermal expansion characteristics. Further, multi-physics simulations of the array in transmit mode showed favorable thermal distribution at up to 2 watts of RF input power per element. The resulting peak temperature was only 40° C above the ambient environment.
Chapter 3: Planar Millimeter-Wave Tightly Coupled Arrays

In this chapter, we will further the concepts developed in Chapter 2 in order to achieve wideband operation supporting the many emerging applications across the millimeter-wave band. Many millimeter-wave antenna arrays have been developed in the literature, primarily operating at the Industrial, Scientific, Medical (ISM) band centered at 60 GHz. Antenna-on-Chip (AoC) arrays [73–75] integrate the array directly above the transceiver circuitry. These are very compact, but suffer from low efficiency (10–50%) due to the high loss in the silicon substrate. Conversely, Antenna-in-Package (AiP) arrays utilize low loss substrates to house the radiating elements, typically with a flip-chip interface to the transceiver die. These can be realized in Low- and High-Temperature Cofired Ceramics (LTCC [76,77] and HTCC [78]) as well as standard Printed Circuit Board (PCB) [79,80].

However, all of the aforementioned arrays utilize simple, narrow-band elements. Given the cost and space required for these arrays, it is not feasible to include multiple narrow-band arrays to cover different functions or bands of operation. For this reason we endeavor to develop broadband millimeter-wave arrays, which will allow the consolidation of many different functions into a shared aperture.

A key challenge to be addressed in this chapter will be the planarization of TCDA arrays, to enable their fabrication at these small scales. This is a result of two
factors: limited feature sizes and tolerances, and minimal fidelity in constructing complex through-plane structures. In this chapter, we present a novel and greatly simplified feed structure for TCDA elements. For the first time, this enables the design of planar, UWB arrays for millimeter-wave applications, which are compatible with low-cost PCB fabrication. An optimized design is presented which provides continuous coverage from 24–72 GHz. This design is validated through fabrication and measurement of several array prototypes.

3.1 Design Challenges

3.1.1 Planar Implementation

Whereas early TCDA [6, 50] were inherently planar in design, recent TCDA [7, 10, 11] have migrated to a non-planar approach. For clarity, we note that herein planar refers to the co-orientation of the fabrication plane with the array plane, as shown in Fig. 3.1. The use of vertical PCBs in [7, 10, 11] is a result of the fact that it is much easier to accommodate complex feeding circuits within the plane of the PCB. Indeed improvements in bandwidth, size, and scanning over earlier designs have been driven by innovation in the design of the feeding network, including the balun and other matching components. With these improvements, however, comes significant complexity. Simultaneously, vertically oriented PCBs provide a predominately air substrate, largely mitigating surface waves.

However, unlike a planar printed array, vertical PCBs require final physical assembly of separate components to form a complete array. This adds complexity and cost to the fabrication process, and moreover introduces sources of error and inconsistency in the resulting arrays. In moving to millimeter-wave frequencies this trend cannot
Figure 3.1: Illustration of planar (left) and non-planar (right) arrays. In both cases, a groundplane resides below the elements in the $x, y$ plane, and broadside is in $z$-direction. Notably in the planar case, the entire array can be formed from a single PCB, whereas in the non-planar array several PCBs must be combined to form the array.

continue, as subassembly at this scale is not feasible. However, planar fabrication of introduces significant challenges in the design and implementation of UWB arrays. As can be seen in Fig. 3.1, planar radiating elements (for instance coupled dipoles) are relatively easily implemented within this framework. However implementation of the feeding network, which is critical to the wideband operation of the array, is significantly more challenging. This is because planar fabrication techniques such as PCB have little fidelity for complex through-plane structures. As such, the primary building block for the feeding network must be simple metallized vias. For the same reason, through-plane radiating elements such as tapered slots cannot be implemented. As a result, existing planarized UWB arrays must utilize simplified balun structures which significantly limit bandwidth [50,51].
3.1.2 Material and Process Selection

It is important to realize that in planar arrays the volume of the antenna is buried in dielectric, making material parameters an important consideration. This includes loss tangent as well as dielectric constant, which plays a large role in determining the onset of surface waves (discussed in greater detail in Section 2.2.2). Furthermore, material selection also relates to the choice of fabrication process.

Low Temperature Co-fired Ceramic (LTCC), for instance, is a popular process for millimeter-wave antennas, for its low loss and relatively fine feature sizes. However, the ceramic tapes used in this method have high dielectric constant ($\epsilon_r > 6$), which can result in surface waves even at broadside. This can be mitigated across a narrow band by making the antenna thin or with EBG structures, but is not suitable for wideband applications. Similarly, Antenna-on-Chip (AoC) designs have been explored, which place the antenna directly on the silicon wafer housing the transceiver circuitry. The silicon substrate, however, has a high dielectric constant ($\epsilon_r \approx 11$) and high loss tangent at high frequencies ($\tan\delta > 0.015$), resulting in very low efficiency for these antennas. This has recently been mitigated by isolating the antenna from the silicon and using low-loss polymers as a substrate. Similarly, purely metal-air structures have been demonstrated using micro-fabrication techniques. However, it remains that AoC and micro-fabrication techniques are very expensive.

Conversely, Printed Circuit Board (PCB) fabrication is almost universally available, and thus can be sourced at very low cost. Likewise, PCB fabrication is compatible with a wide range of materials, including low loss and low dielectric constant polytetrafluoroethylene (PTFE or Teflon, $\epsilon_r = 2.2$). For these reasons, it is desirable to enable PCB fabrication for antenna designs. However, PCB fabrication is also
much more limited than LTCC and micro-fabrication, particularly for high frequency and planar antennas.

3.1.3 Limitations of PCB Processing

Basic design rules for advanced PCB processes are summarized in Fig. 3.2, with values given in Table 3.1. We observe that the minimum feature size (positive or negative) within the plane is limited to 76 $\mu$m (3 mil), which at 90 GHz translates to 2.25% of the wavelength. However, signal paths travelling through-plane must be formed from vias, and are further limited. While laser drilled holes can reach a similar diameter as the traces (i.e. 3 mil, 76 $\mu$m), they are limited to very thin layers. The comparatively thick core substrate must be mechanically drilled, resulting in a minimum diameter of 6 mil (152.4 $\mu$m). These further require annular catch and landing pads for the metallization process, which increase the surface diameter to 12 mil (304.8 $\mu$m). Likewise, in-plane traces are required to maintain a 6 mil clearance from all via edges. Finally, the minimum edge-to-edge via pitch is 10 mil (254 $\mu$m), or 7.6% of the wavelength at 90 GHz.

The severe challenge inherent in the available fabrication tolerances is illustrated by briefly considering some scaling examples. Taking for example [7], and applying the linear scaling principle to shift its operation up to a target frequency of 90 GHz would require fabrication of features down to 5.25 $\mu$m (0.2 mil) in size. Even if we consider the improved high frequency design described in Section 2, when scaled to 90 GHz would require features of 41.5 $\mu$m (1.63 mil). While clearly these dimensions may be achievable using on-wafer, micro-fabrication techniques, they are a far cry from the PCB capabilities outlined above.
Figure 3.2: State of the practice Printed Circuit Board fabrication limits, shown from edge. Key limiting parameters are labeled, and provided in the Table 3.1

Table 3.1: Minimum feature sizes for PCB fabrication, corresponding to the parameters shown in Fig. 3.2

<table>
<thead>
<tr>
<th>Dimension</th>
<th>µm</th>
<th>mil</th>
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<tbody>
<tr>
<td>A</td>
<td>76.2</td>
<td>3</td>
</tr>
<tr>
<td>B</td>
<td>76.2</td>
<td>3</td>
</tr>
<tr>
<td>C</td>
<td>152.4</td>
<td>6</td>
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<td>D</td>
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<td>12</td>
</tr>
<tr>
<td>F</td>
<td>254</td>
<td>10</td>
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</tbody>
</table>
3.2 Design Approach

The goal of this design is to develop a low-cost, multifunctional array for the many emerging applications in the millimeter-wave spectrum. Accounting for the difference in scale of the antenna array (>100 mm²) and transceiver dies (≈ 10 mm²), we will approach this design as an Antenna-in-Package application. This assumes the transceiver and antenna are bonded using a flip-chip approach, and implies that the antenna materials and processing are not dependant on the transceiver. The input to the antenna is assumed to be a 50 Ω unbalanced line, such as microstrip or co-planar waveguide. To be mass-market compatible, and scalable to high gain applications, the design accounts for fabrication using standard PCB processes. Moreover, largely turnkey preparation is expected, implying no complex assembly after fabrication.

3.2.1 Development of Three-Pin Balun

Given the limitations outlined above, the simplest feeding approach may be a direct or unbalanced feed such as [51]. This has been shown to be a viable option, however requires extensive matching circuits beneath the groundplane to compensate for the reactive loading of the dipoles. Likewise the feeding network, including balun, can also be made planar to take advantage of the improved in-plane tolerances [50]. However, it is known that a planar balun is limited in size by the element spacing, and thus struggles to achieve a wide impedance ratio, leading to a limited bandwidth (>2:1 in [50]).

Instead we will seek to develop a more compact and simplified balun, using only very basic transmission lines which can be formed from combinations of vias (for instance twin-wire as opposed to coax or stripline). Ultimately, we expect the antenna
and feeding network to resemble the circuit model shown in Fig. 3.3, and so begin by simply imitating these same circuit elements. For this analysis we will take as a starting point a tightly coupled dipole array fed from an ideal port directly across the dipoles, as shown in Fig. 3.4. This assumes an infinite array, with a conducting wall along the H-plane of the array per the discussion in Section 2.2.3 As we would expect, the purely differential feed demonstrates a wideband response. However, producing such a feed from the unbalanced input beneath the groundplane is the primary challenge.

Implementing the shunt short circuit seen in Fig. 3.3 is relatively trivial, requiring only two shorting pins from the dipoles to ground, as shown in Fig. 3.5. Notably, this parallel inductance inverts the impedance behavior, drawing the lowest frequencies to the top of the Smith chart, while bringing the high frequencies roughly to the center. By inspection, we expect that a series open circuit element would now push the low
Figure 3.4: Model of an infinite array of tightly coupled dipoles above a groundplane fed from an ideal gap source, shown in black, (left) and $S_{11}$ response from 20–90 GHz (right). This model assumes a conducting wall along the H-plane, and will serve as the starting point for the design of the feed.

We might attempt to imitate the balun structures in [7,10,11] by bringing the feed up along one of the shorting pins, and down the opposite. In this case the open circuit would comprise a partial via, along the lines of Fig. 3.6. Some degree of tuning can be accomplished by varying the length of the via, but little can be done to alter the characteristic impedance, as the via spacing is bounded by the fabrication process. We note that this is effective to a certain extent, but ceases to function at the high end of the band. This is a result of the weak coupling of the twin-wire transmission line, between the open circuit and the shorting pin, which allows the signal to decouple into a propagating mode along the substrate instead.
Figure 3.5: Two conducting vias are added to the model from Fig. 3.4, shorting the dipoles to ground (left). This forms the high impedance shunt shorted stub seen in Fig. 3.3, inverting the $S_{11}$ response of antenna (right).

Figure 3.6: Series open-circuit implemented in the style of [7,10,11], using a via opposite the shorting pins (left). VSWR is plotted to better show frequency-dependent behavior (right). Notably, the open circuit is effective in matching the low frequencies, but becomes detuned at high frequencies (grating lobe onset frequency is 90 GHz in this simulation).
Figure 3.7: The open circuit is implemented as a patch on top of the opposite dipole. The $S_{11}$ plot shows all portions of the band are well matched. The strong coupling of the patch, compared to the pin in Fig. 3.6 mitigates detuning at higher frequencies.

To improve the high end response, we replace the stub open circuit with a patch parallel to the dipole, as shown in Fig. 3.7. This is situated above the dipole to avoid the shorting pins. As thin laminates can be used, this patch can be brought very close (i.e. 25 $\mu$m or 1 mil) to the dipole to achieve the desired low impedance. We note that in Fig. 3.7 the capacitive patches between dipole tips are also brought on top of the antenna; this is simply to reduce the number of layers, and does not meaningfully impact the antenna performance.

The low impedance patch has the desired effect of maintaining a good impedance match at the highest portions of the band. However, the antenna is still fed from an ideal port, near the aperture. We must still consider how this feed point can be translated to an accessible point beneath the groundplane. The envisioned feed is shown in Fig. 3.8, wherein a thin microstrip trace continues along the dipole opposite
Figure 3.8: Conceptual model to route the antenna feed towards the groundplane. This requires that the short microstrip trace above the dipole can match the impedance of the antenna ($\approx Z_0 = 100\Omega$).

the patch, and once clear of the shorting pin, a via drops through the dipole towards the groundplane. However, we see in Fig. 3.9 that the antenna rapidly becomes detuned, even before reaching the feed via. This is because the feed line should have roughly $Z_0 = 100\Omega$ to match the impedance at the aperture. At a distance of 1 mil from the surface of the dipole, the microstrip trace would have to be 0.6 mil in width to reach this impedance. We note this introduces a trade-off between the impedances of the open-circuit element and microstrip feed, which have opposite requirements with regard to the separation of the dipole and feed layers.

Finally, as seen in Fig. 3.8, the input feed is brought below the groundplane with a third via. This via, together with the nearby shorting pin (orange in Fig. 3.8) forms a high-impedance twinwire transmission line. The element is fed from pin A. We observe that this structure forms two transmission lines opposite to the center
Figure 3.9: Analysis of the impact of the microstrip trace in the feeding network. Due to the fabrication limitations, the trace width cannot be less than 3 mil. However, due to the proximity of the dipole, the trace $Z_0$ is low. This causes detuning of the element feed, even in the short length of the microstrip transition (right, $L = 0–10$ mil).
conductor (pin B, which is shared between both transmission lines). This is better understood by considering a transmission line model of the feed formed by these vias and open patch, as is shown in Fig. 3.10, with $Z_{\text{ant}}$ unchanged from Fig. 3.3. The impedance $Z_{\text{feed}}$ is then primarily determined by the separation $d_{\text{feed}}$, and likewise $Z_{\text{short}}$ is determined by $d_{\text{short}}$. Careful comparison shows that this circuit contains the same fundamental elements of the Marchand balun illustrated in Fig. 3.3. In principal, the additional design parameters of the balun would allow us to expand the bandwidth of the isolated dipoles (Fig. 3.4). However, as the limited fabrication tolerances severely limits flexibility in the selection of these parameters, we expect some reduction in the bandwidth.

At this stage, the aperture has been matched to a feedline accessible by the back-end. We note that the current feed assumes a nominal 75 Ω input. However, from [57] we expect that this can be matched to a 50 Ω system relatively easily, with a short taper. This can be implemented in stripline or co-planar waveguide, beneath the groundplane.

### 3.2.2 Addition of a Conducting H-Wall

The challenge of mitigating spurious cavity resonance within the volume of the array, as introduced in Section 2.2.3, is exacerbated in the planar TCDA design by the need for a continuous dielectric substrate. The fields at resonance are illustrated in Fig. 3.11. We consider the relation:

$$f_r = \frac{c}{2D \sqrt{\epsilon_{\text{eff}}}},$$  \hspace{1cm} (3.1)

wherein $f_r$ is the frequency of resonance, $D$ is the distance between the grounded conductors, and $\epsilon_r$ is the relative dielectric constant of the substrate. Substituting
Figure 3.10: Side view of the three-pin balun shown in Fig. 3.8 (left), as well as equivalent transmission line model (right). This circuit can be understood by observing the similarity to the balun circuit shown in Fig. 3.3.

the values $D_1 = \sqrt{(D_x - d_{\text{feed}})^2 + D_y^2} = 2.2$ mm and $\epsilon_r = 2.1$, we calculate $f_r$ as 46 GHz. This value agrees closely with simulated results, as observed in the central peak in Fig. 3.12.

To eliminate this resonance, we must force $f_r$ to occur outside the band of interest. However, we note that as a result of the higher effective dielectric constant, the resonance observed in Fig. 3.12 occurs much lower in the relative operating band than it did in the previous design (compare to Fig. 2.21). As a result, the addition of shorting pins from the dipole edges to the groundplane, as was successfully implemented in the previous design, is not sufficient to eliminate the resonance. We see this in the right-most peak in Fig. 3.12. This is because the resonant cavity length forms diagonally, between the shorting pin and the grounded conductor of a neighboring element (see Fig. 3.13). This diagonal path would have an approximate length $D_2 = 1.78$ mm, resulting in $f_r = 58$ GHz, which agrees closely with the simulated result.
Figure 3.11: Illustration of the electric field distribution excited by the vertical currents of the input feed (dashed arrows) during cavity resonance.

Figure 3.12: Simulated VSWR demonstrating a destructive cavity resonance in the unmodified (central peak), alternate substrate (left peak), and shorting pin (right peak) designs. The resonance can be mitigated with a continuous conducting “H-wall”.

91
Figure 3.13: Top view of a $2 \times 2$ array, indicating the orientation of the resonant length $D$ in (3.1) of the spurious cavity mode. Based on the lengths $D_1$ (unmodified) and $D_2$ (with shorting pin), we expect that the addition of a shorting pin will only slightly increase the resonant frequency. Conversely, inclusion of an H-wall will significantly increase the resonant frequency.
Conversely, we may try to lower $f_r$ by using a substrate of higher $\epsilon_r$. Substituting $\epsilon_r = 6$ into (3.1), we see $f_r$ is indeed lowered to 28 GHz, corresponding to the leftmost peak in Fig. 3.12. Likewise, we can use (3.1) to calculate a sufficiently high $\epsilon_r$ to eliminate the resonance completely (i.e. set $f_r = 18$ GHz), yielding $\epsilon_r > 16$. While the calculated value is theoretically achievable, it is not practical and would cause detrimental surface waves when the array is scanned.

The occurrence of this resonance can be pushed higher, and ultimately out of the band, by replacing the single shorting pin by a continuous conducting wall perpendicular to the primary dipole current (solid curve, Fig. 3.12), referred to as an H-wall. Though ideally a solid and infinitesimally thin conducting wall, the nominal H-wall can be approximated by a series of connected shorting pins, as shown in Fig. 3.14. We note that the vias used to form the H-wall must be mutually connected near the tips to avoid monopole radiation. Finally, the same effect can be realized with the inclusion of an orthogonal polarization, in the dual-offset layout described in [10]. This layout requires a single shorting pin at the intersection of the two polarizations.

### 3.2.3 Sample Design for 5G Frequencies

The ever-growing need for high bandwidth mobile communications has led to exploration of the millimeter-wave spectrum. Unfortunately, allocated frequencies for millimeter-wave communications are widely distributed. Table 3.2 shows the existing unlicensed Industrial, Scientific, Medical (ISM) bands as well as the recently allocated Fifth Generation (5G) mobile bands, in and around the millimeter-wave spectrum [81]. Together, these allocations span nearly 50 GHz. Given the required cost and space, it is not feasible to include many narrow-band arrays to cover the 5G
and ISM bands outlined in Table 3.2. Instead, it is necessary to consolidate all six millimeter-wave 5G and ISM bands into a common, multi-functional aperture. Applying the concepts introduced above, we present a planar, UWB array which supports all the allocated 5G bands by providing continuous coverage from 24–72 GHz.

Table 3.2: Allocated Millimeter-Wave 5G and ISM Bands

<table>
<thead>
<tr>
<th>Allocation</th>
<th>Frequency (GHz)</th>
<th>Allocation</th>
<th>Frequency (GHz)</th>
</tr>
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<tr>
<td>ISM</td>
<td>24–25</td>
<td>5G</td>
<td>38.6–40</td>
</tr>
<tr>
<td>5G</td>
<td>27.5–28.35</td>
<td>ISM</td>
<td>57–64</td>
</tr>
<tr>
<td>5G</td>
<td>37–38.6</td>
<td>5G</td>
<td>64–71</td>
</tr>
</tbody>
</table>

The design parameters of the unit cell are shown in Fig. 3.15, with the optimized values given in Table 3.3. Notably, the values given in Table 3.3 fall within the acceptable values for PCB fabrication, detailed in Table 3.1. Likewise, the layout uses
Figure 3.15: Key design parameters of the nominal unit cell given in Fig. 3.14. Optimized values of these parameters are given in Table 3.3

Commercially available Duroid 5880 ($\varepsilon_r = 2.2$) substrates for the core and superstrate, with Rogers 2929 prepreg ($\varepsilon_r = 2.9$). Full-wave simulations of an infinite array of the proposed unit cell are shown in Fig. 3.16, for broadside, and $\pm 45^\circ$ scans in the E- and H-planes. The array is assumed to be fed from a 50 $\Omega$ unbalanced source beneath the groundplane. We note that the array operates across the desired bands, from 24–72 GHz, for VSWR $< 2.2$ at broadside, and VSWR $< 3$ for scans up to $45^\circ$ in the E- and H-planes.
Table 3.3: Optimized values of the parameters shown in Fig. 3.15.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>mm</th>
<th>mil</th>
<th>Parameter</th>
<th>mm</th>
<th>mil</th>
<th>Parameter</th>
<th>mm</th>
<th>mil</th>
</tr>
</thead>
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<tr>
<td>A</td>
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<td>4</td>
<td>B</td>
<td>0.152</td>
<td>6</td>
<td>C</td>
<td>0.076</td>
<td>3</td>
</tr>
<tr>
<td>D</td>
<td>0.406</td>
<td>16</td>
<td>E</td>
<td>0.305</td>
<td>12</td>
<td>F</td>
<td>2.06</td>
<td>81</td>
</tr>
<tr>
<td>G</td>
<td>0.1</td>
<td>4</td>
<td>H</td>
<td>0.457</td>
<td>18</td>
<td>I</td>
<td>0.152</td>
<td>6</td>
</tr>
<tr>
<td>J</td>
<td>0.406</td>
<td>16</td>
<td>K</td>
<td>0.457</td>
<td>18</td>
<td>L</td>
<td>0.635</td>
<td>25</td>
</tr>
<tr>
<td>M</td>
<td>0.61</td>
<td>24</td>
<td>N</td>
<td>0.1</td>
<td>4</td>
<td>O</td>
<td>0.914</td>
<td>36</td>
</tr>
</tbody>
</table>

Figure 3.16: Infinite array VSWR of the array shown in Fig. 3.15, at broadside (black) and 45° scanning in the E- (red) and H- (blue) planes. The 5G and ISM bands from Table 3.2 are highlighted for reference, and are observed to fall in the passband of the array.
3.3 Prototype Fabrication

The array design shown in Fig. 3.15 was validated through the fabrication of several prototypes. Several $3 \times 3$ and $5 \times 5$ arrays, as well as calibration standards and test articles were integrated into a single PCB panel for fabrication. As the material stackup and design features fall within common fabrication capabilities, the fabrication was outsourced to a commercial vendor. Unlike the test articles discussed in [7, 10] as well as in Section 2.4 of this work, the prototypes of this array can not be used to measure multiple elements sequentially. Instead, each fabricated coupon, containing a fully populated array, only activates a single element for measurement. The remaining elements are terminated with a $50 \Omega$ resistor, which serves as a matched load.

The ability to feed only a single element for measurements is due to the small size of the array elements, which prevents connectorizing each element, as in previous designs. Instead, the active element feed is routed away from the array via microstrip, to an edge-launch 1.85 mm coaxial cable port. This layout is illustrated in Fig. 3.17 for the case of a $3 \times 3$ array, with the center element fed. Measurement of other elements within the array is accomplished by fabricating otherwise identical coupons, with the desired element connected. This approach relies on the repeatability of the fabrication process between coupons. To simplify de-embedding and insertion loss compensation, the feed and coax footprint remains identical, and the array is shifted to align the appropriate element.

While the coupon shown in Fig. 3.17 is large compared to the array itself, we note that this arrangement is required for measurements only. In an application setting, it is expected that the transceiver circuitry would be bonded directly to the
Figure 3.17: Sample layout of a $3 \times 3$ array measurement coupon with the center element fed, shown from top, cutaway view (left) and bottom view (right). Approximate dimensions are included for reference.

antenna inputs, via flip-chip solder bumps. As such, the surrounding groundplane and long input feed from the coax port are not needed, leaving a compact antenna array package. A fabricated $5 \times 5$ array prototype in the expected application packaging is shown in Fig. 3.18, with a U.S. penny for scale (0.75 in diameter). However, the prototype shown in Fig. 3.18 is only useful for visual reference. Characterization of the antenna performance requires measurement coupons having the elements individually interfaced, as is shown in Fig. 3.17.

Several challenges were encountered in producing the needed test articles. Indeed the initial fabrication run failed, due to an inability of the metallization solution to adhere to the Duroid substrate (Teflon-based material) in the long vias running from the dipoles to the groundplane. It was suggested by the PCB vendor to intentionally
Figure 3.18: Prototype 5 × 5 array (approximately 12 × 12 × 1.7 mm³) without measurement fixtures, alongside a U.S. penny (0.75 in diameter). This coupon is not used for measurement, but is representative of the array as it would be applied in practice. Black surface seen is the superstrate material (Duroid 5880), and thin tan layers are the prepreg (Rogers 2929).

deviate from the design rules, to increase the via diameter to 8 mil to improve adhesion. All copper layers and the via centerpoints were kept identical. This solution was successful and the coupons could be fabricated. Unfortunately, the final nonplated drilling stage, used to make the mounting holes for the coax port, was inadvertently skipped (see Fig. 3.19). The coupons were sent back to drill the holes in re-work. However, as the individual coupons had been de-panelized prior to the initial shipment, the re-work would be performed without fiducial markers. This resulted in high inaccuracies and, moreover, inconsistency between coupons (see Fig. 3.19).

Simulations were conducted to assess the impact of misalignment on the coax-to-microstrip transition, which may arise due to the inaccurate mounting holes. As seen in Fig. 3.20, this is a very sensitive parameter, with as little as ±10 mil (0.254 mm) displacement causing several dB of loss, particularly at the low end of the band. Further, we note that at ±15 mil (0.38 mm) offset, the coax launch pin is no longer touching the microstrip trace, resulting in near total reflection ($S_{21} < -30$ dB). As
Figure 3.19: (a) Initial fabrication delivered by vendor lacked mounting holes for the coax port. (b) Holes were drilled as rework, but suffered low accuracy due to lack of fiducial markers on de-panelized boards. As this was anticipated, hole diameter was increased to allow for misalignment.

this was anticipated, the hole diameter was increased to account for the potential offset. Final alignment of the port within the range allowed by the mounting holes was done by hand.

Similarly, the routing used to separate coupons from the fabrication panel was found to be inconsistent, resulting in varied coupon outlines. This is also a potential source of error at the edge-launch coax port, where the data sheet calls for the footprint to sit flush with the substrate edge. As can be seen in Fig. 3.21 however, some coupons have up to 1 mm of excess dielectric beyond the footprint edge. Simulations were conducted to understand the effect of this additional dielectric, shown in Fig. 3.22. Similar to misalignment of the port (Fig. 3.20), offsets of the port due to extra dielectric cause several dB of losses at relatively small values (20 mil or 0.51 mm). Notably, slightly negative offsets (i.e. cutting away part of the footprint) have very
Figure 3.20: Simulated $S_{21}$ of the coax to microstrip transition, with the coax port shifted laterally by 0–10 mil (0–0.254 mm). We note at 15 mil offset the signal is completely reflected (not shown in figure).

little effect. This implies the excess dielectric can be filed down with relatively low risk of filing away too much.

After filing the excess dielectric and aligning the port by hand, the mounted coax port can be seen in Fig. 3.23. The feed pin of the coax transition can be seen in the center of the port. The coupon is also shown next to a ruler for scale. Finally, four coupons were selected for measurement based on visual inspection for fabrication errors. Initially, only $3 \times 3$ prototypes were considered, to accelerate the process and validate the measurement system. High frequency 0201 packaged 50Ω resistors, serving as matched terminations, were soldered to the array elements surrounding the actively fed element. These were tested for good connection at DC. As the soldering process was carried out by hand, the effective yield was roughly 50%
Figure 3.21: Excess substrate found at the edge of the coax footprint is a result of de-panelization. The thickness of the substrate boundary varies between coupons, and must be removed to avoid losses.

Figure 3.22: Simulated $S_{21}$ of the coax to microstrip transition, for various values of the excess dielectric shown in Fig. 3.21. Negative values imply too much has been removed, cutting into the coax footprint. We note at 30 mil, $S_{21} < -10$ dB (not shown in figure).
Figure 3.23: (a) Mounted 1.85 mm coax port showing alignment of launch pin and microstrip trace. The port is bolted in place, not soldered, and thus can be moved or removed as needed. (b) Picture of a $3 \times 3$ coupon with coax port mounted; ruler included for scale.

per cycle. Thus despite several soldering cycles, two elements of 36 in the selected measurement coupons still failed the DC test. The positions of these elements are shown in Fig. 3.25.

3.4 Measurement

Far-field measurements of the of the four selected array prototypes were conducted at the NASA Glenn Research Center millimeter-wave range. A conceptual schematic of the measurement setup is shown in Fig. 3.26a. The Antenna Under Test (AUT) is placed at port 2, where it receives the signal radiated by a source antenna at port 1. The AUT can rotated about the vertical axis to capture scanning information in one
plane (which plane depends on the orientation of the AUT). Cross-polarized gain is measured by rotating the source antenna. This setup as implemented at the NASA Glenn Research Center is shown in Fig. 3.26b.

Pictures of a source antenna, and mounted test coupon are shown in Fig. 3.27. Due to the wide bandwidth of the array, each coupon was measured in three separate bands. Open-Ended Waveguide (OEWG) sources were used for the Ka-band (20–40 GHz) and U-band (40–60 GHz), and a standard gain horn source was used in the V-band (50–75 GHz). Gain calibration, accounting for both cable losses and free space path loss, was determined by measurement of identical OEWG or horn antennas at the transmit and receive ports. A normalization factor can be determined based on the measured $S_{21}$ and the known gain of these antennas. Finally, insertion loss of the measurement fixture of the coupon, including losses in the coax port and long
Figure 3.25: Locations of the active elements as well as failed connections after soldering of 50 Ω resistors.
Figure 3.26: (a) Schematic of the far-field test setup used for gain characterization of the fabricated prototypes. (b) Picture of the millimeter-wave range at the NASA Glenn Research Center, with key elements labelled. Open-ended waveguides can be seen mounted at both the source and receive ports.
Figure 3.27: (a) Picture of the U-band open-ended waveguide source mounted at port 1. (b) Picture of one coupon mounted at port 2. The coax cable feeding the antenna is run inside the mounting post; the entire post rotates during measurement to change the angle of incidence.

input microstrip trace, are determined by a simple “through” $S_{21}$ measurement, as shown in Fig. 3.28.

Measured gain of the individual active element in each of the four coupons is shown in Fig. 3.29 for E-plane elements, and Fig. 3.30 for H-plane elements (see again Fig. 3.25 for coupon orientation details). These are compared to simulated values, using a semi-finite (3 × infinite) model. Notably, the measurements track the simulated values very closely, though some deviations are observed at the low end of the band for both center elements (Figs. 3.29a and 3.30a). These nulls may result from resonances of the larger coupon and feedline, which are not accounted for in the semi-finite simulation (see Fig. 3.31). To verify this, the full 3 × 3 array and
feed fixture was simulated. Due to the high computational cost of this simulation, it was carried out only for the E0 element, and with a more course frequency sweep. These results are shown in Fig. 3.32. We observe that the addition of the extended substrate and feedline does introduce low-frequency nulls in the broadside pattern, though these are shifted slightly in frequency between the simulated and measured results.

Additionally, we observe that the nulls observed in the broadside gain are a result of the changing angular pattern of the element over frequency, as is shown in Fig. 3.33 for the null at 52 GHz. At higher frequencies, this dual-lobe pattern changes into three lobes, as we see in the 56 GHz pattern. These effects are also reflected in simulation (circles in Fig. 3.33). The discontinuous nature of $3 \times 3$ array frequency response is a result of its small size. As the array size increases it more closely approximates the infinite array assumed during the design process, resulting in reduced ripple in the element pattern and frequency response.
Figure 3.29: Measured embedded element gain at broadside (blue) compared to simulated values (red) for two measurement coupons, with the array oriented horizontally. Edges of the measurement bands are indicated (dashed lines).
Figure 3.30: Measured embedded element gain at broadside (blue) compared to simulated values (red) for two measurement coupons, oriented vertically. Edges of the measurement bands are indicated (dashed lines).
Figure 3.31: (a) Semi-finite simulation model which assumes a periodic boundary along one side of the array to reduce computational burden. This was the model used for the results shown in Figs. 3.29 and 3.30. (b) Complete model including extended substrate and feedline of the measurement coupon (shown from below).

Figure 3.32: Measurement data of the center element compared against $1 \times 3$ and $3 \times 3$ simulations (models shown in Fig. 3.31). As expected, the increased accuracy of the $3 \times 3$ simulation introduces nulls that are more noticeable at the low frequencies and which are also observed in the measured results.
Figure 3.33: Measured (solid) and simulated (circles) element patterns at 52 GHz and 56 GHz show close agreement. We observe that dips in the broadside gain (see Fig. 3.32) are a result of changes in the element pattern over frequency.

### 3.5 Conclusion

There is a need for many emerging applications to have multi-functional access to a wide range of the millimeter-wave spectrum. Current antennas at these frequencies are narrow-band and expensive to produce in large quantity. Simultaneously, UWB Tightly Coupled Arrays could not be designed at these frequencies due to the challenge inherent in miniaturizing the complex feed structures.

In this chapter, we present a novel and greatly simplified feed structure for TCDA elements, which enables the design of planar, UWB arrays for millimeter-wave frequencies. Moreover, these designs are compatible with low-cost and low-loss PCB fabrication. This is validated through the design, fabrication, and measurement of a TCDA array operating across 24–72 GHz.
Chapter 4: Measurement Techniques for Millimeter-Wave Antennas

Far-field characterization of array antennas can require many sequential measurements. In preparing such measurements, the necessary antenna repositioning and rotation can introduce small displacements. These displacements are not as important at low frequencies, but can be a significant fraction of a wavelength at the Ku-, Ka-, and millimeter-wave bands. Therefore, it is important to correct for them to obtain accurate array measurements. In this chapter we will develop a mathematical model describing the expected phase response of each antenna element for a given physical offset. This error model is applied to Ku-band measurements of an 8×8 dipole array. It is demonstrated that small movements of the antenna during measurement can be corrected via the proposed post-processing technique. Portions of this chapter have been published in [82].

4.1 Displacement Correction in Sequential Far-Field Measurements

Antenna performance is commonly characterized using far-field measurements, typically in an anechoic chamber. The antenna under test (AUT) is illuminated by
a source antenna, located at a distance such that the incident wave can be considered planar [83]. Either the AUT is then rotated in place [84] or the source is swept across an arc [85], and the gain and phase are measured at each incidence angle. For aperture antennas, the center of rotation should coincide with the phase center of the antenna; the same applies to antenna arrays if they include an appropriate beamforming network. However, for low cost measurements, arrays can be characterized independently of the beamforming backend. This can be accomplished using the Unit Excitation Active Element Pattern (UEAEP) technique [69, 86]. Briefly, UEAEP can be thought of as sequential digital beamforming, wherein each array element is characterized individually. Subsequently, the contribution of all array elements is combined via post-processing. Critically, this approach includes the full effects of mutual coupling within the array, making it more accurate than a simple embedded element pattern.

Individual characterization of the array elements implies repeatedly connecting and disconnecting elements throughout the measurement process. With constant handling across numerous measurements, small displacements of the AUT can occur throughout the characterization. We note that in a chamber environment without specialized equipment to achieve an accurate frame of reference, millimeter-scale positional errors are effectively impossible to observe. Physical displacement in the measurement setup introduces errors in the characterization of the antenna phase response. Specifically, phase misalignment impacts the phase of the element patterns, leading to distortion of the combined beam pattern. Although these errors are negligible at lower frequencies, above 8 GHz offsets as small as 1 mm represent a significant fraction of the wavelength. While this topic has been discussed in the
context of near-field measurements [87–90], it has not yet been addressed for far-field measurements.

In this section, a post-processing technique is described to identify and compensate for variable displacement of the AUT in sequential far-field measurements, such as may occur during UEAEP characterization. Notably, this technique is not intended to replace appropriate calibration and equipment, or otherwise good experimental practice. The concept will be derived from a mathematical model of the measurement setup, and validated using measured data of an 8×8 array operating in the Ku-band (i.e., 18 GHz).

4.2 Mathematical Model

In this analysis, we assume an AUT in a receive configuration on a rotating platform, as in Fig. 4.1. The fixed transmit antenna is situated in the far field (> $2D^2/\lambda$, where $D$ is the largest antenna dimension and $\lambda$ is the wavelength), such that the incident wavefront may be considered planar [83]. When an arbitrary offset is introduced in the array’s position, the element paths are altered. From [69], an exact representation of the radiated fields in the rotation plane of either array in Fig. 4.1 is given by

$$E(\theta) = \sum_{i=1}^{N} V_i g_i(\theta),$$  \hspace{1cm} (4.1)

where $g_i(\theta)$ is the $i$th complex element pattern, and $V_i$ is the complex excitation voltage. Similarly, the measured phase at each element can be expressed as

$$\phi_i^M(\theta) = \angle g_i(\theta).$$  \hspace{1cm} (4.2)

From this phase we can easily calculate that element’s relative position with respect to the fixed transmitter (much like a radar). Given the circular motion of the element,
Figure 4.1: Measurement setup of a two-element array (the Antenna Under Test) with an incident plane wave from a transmitting horn. The array is physically displaced from the center of rotation by offsets $x$ and $y$. We note that $x$ and $y$ align with the coordinate axes $\hat{x}$ and $\hat{y}$ when the array is oriented to broadside.

we expect it to follow the path

$$D_{Tx}(\theta) = \frac{\phi_i^M(\theta)}{k} = d_i \sin \theta + D_0,$$  \hspace{1cm} (4.3)

where $\theta \in [-90^\circ, 90^\circ]$. In (4.3), $k = 2\pi/\lambda$ is the wavenumber, and $d_i$ is the radial location of the element with respect to the center of rotation. The distance between the phase centers of the transmit and test antennas, $D_0$, is unknown due to aliasing of the phase and can be disregarded.

A useful way to visualize the measured phase is to examine the Cartesian plot of the calculated positions of each element versus $\theta$. Fig. 4.2 gives plots of $D_{TX}$ for an 8 element, $0.5\lambda$ spaced linear array. As expected, we observe that the phase differential between elements of the array is maximized at endfire ($\theta = \pm90^\circ$). At broadside, the relative phase difference between elements converges to zero as they are in-phase with respect to the incident wavefront.
Figure 4.2: Relative positions of eight elements of a linear array, traveling from $-90^\circ$ to $90^\circ$. We note at broadside, all eight elements are in-phase (zero relative separation) whereas at $\pm 90^\circ$ they are out of phase ($0.5 \lambda$ separation).

Even with an arbitrary offset from the center of rotation, the AUT is swept through $180^\circ$ during measurement, and thus the measured phase will still follow a sinusoidal pattern. From inspection of Fig. 4.1 we can determine that the erroneous phase information would follow the polar equations:

$$\phi_i^e(\theta) = kr_i \sin \theta_i$$  \hspace{1cm} (4.4)

$$r_i = \sqrt{(d_i + x)^2 + y^2}$$  \hspace{1cm} (4.5)

$$\theta_i = \tan^{-1}\left(\frac{y}{x + d_i}\right) + \theta_0 + \theta,$$  \hspace{1cm} (4.6)

where $x, y,$ and $\theta_0$ are offsets of the array in the $\hat{x}, \hat{y},$ and $\hat{\theta}$ directions, respectively. As an example, Fig. 4.2 shows phase distortions due to these offsets, by substituting $\phi_i^M$ in (4.3) with $\phi_i^e$ from (4.4).
For clarity, we observe the impact of $x, y,$ and $\theta$ offsets individually, in Fig. 4.3. Intuitively, a displacement along $\hat{x}$ (up or down in Fig. 4.1 when $\theta' = 0$) has no impact on the measured phase at broadside, but causes the array to appear closer or further away at endfire (see Fig. 4.3a). Similarly, displacement along the $\hat{y}$ direction has no impact at endfire, but causes phase coherence of all elements to occur at some other relative phase (Fig. 4.3b). A $\theta$ offset shifts the broadside angle, and will reduce the angular range of the measurement by $|\theta|$ on both sides (Fig. 4.3c).

While (4.4) is phrased in terms of $x, y$ translations, it also applies to rotations in the spherical coordinate frame which are simply projected onto the scanning plane. However spherical rotations, particularly about the $\hat{y}$ axis, can cause polarization misalignment and deviation from the desired scanning plane. The formulation given in (4.4) assumes that the measurement offsets do not contribute significantly to the free space path loss (i.e., $|g_i(\theta)| \approx |g_i^*(\theta)|$). For very large offsets, additional gain calibrations may be necessary. Further, to maintain the assumption of a planar wavefront, the maximum expected offset should be included in $D$ of the far-field calculation ($2D^2/\lambda$).

If these conditions are met, then as long as these offsets are static and remain constant for the measurements of each element, they will have no impact on the calculated beam pattern. This can be understood by observing the relative phase between neighboring elements. For simplicity, we define $\zeta$ as

$$\zeta_i = \tan^{-1}\left(\frac{y}{x + d_i}\right),$$

and the effective angle of incidence, $\theta'$, as

$$\theta' = \theta + \theta_0.$$
Figure 4.3: Relative positions of eight elements at some offset from the center of rotation (refer to Fig. 4.1): (a) with $x$-offset only, (b) with $y$-offset only, (c) with $\theta$-offset only, (d) with arbitrary offsets of each element.
Using (4.7) and (4.8) in conjunction with the angle-sum identity for sines, we can rewrite (4.4) as:

\[
\phi^e_i(\theta) = kr_i \sin \theta_i
\]

\[
= kr_i \sin(\zeta_i + \theta')
\]

\[
= kr_i (\sin \zeta_i \cos \theta' + \cos \zeta_i \sin \theta'),
\]

(4.9)

Further, using (4.5) we can write the expressions

\[
\sin \zeta_i = \sin \left( \tan^{-1} \left( \frac{y}{x + d_i} \right) \right) = \frac{y}{r_i},
\]

(4.10)

\[
\cos \zeta_i = \cos \left( \tan^{-1} \left( \frac{y}{x + d_i} \right) \right) = \frac{x + d_i}{r_i},
\]

(4.11)

such that (4.9) becomes:

\[
\phi^e_i(\theta) = kr_i \left( \frac{y}{r_i} \cos \theta' + \frac{x + d_i}{r_i} \sin \theta' \right)
\]

\[
= ky \cos \theta' + k(x + d_i) \sin \theta'.
\]

(4.12)

for the \(i\)th element. Therefore, the relative phase between two elements displaced from the center of rotation is:

\[
\Delta \phi^e(\theta) = \phi^e_i(\theta) - \phi^e_{i-1}(\theta)
\]

\[
= k(y - y) \cos \theta' + k(x + d_i - x - d_{i-1}) \sin \theta'
\]

\[
\therefore \Delta \phi^e(\theta) = k \Delta d \sin \theta'.
\]

(4.13)

We observe that (4.13) reduces to the ideal case illustrated in Fig. 4.2 \((x = y = 0)\). This demonstrates that static offset has no impact on the measured beam pattern, an intuitive conclusion.

Static phase offsets in arrays have been well studied [91–93]. Here, our challenge is that the values of the \(x\) and \(y\) offsets change between the array element measurements.
Figure 4.4: Computed array factors of isotropic radiators with ideal (solid) and distorted (dashed) characterizations, as shown in Fig. 4.2 and Fig. 4.3d, respectively. Notably, even small phase errors ($\sigma = 0.125\lambda$) contribute to significantly increased sidelobe levels and beam skew.

An example of such element trajectories is shown in Fig. 4.3d. For these trajectories, we assumed a normal distribution of $x$ and $y$ with $\sigma = \lambda/8$ (2.08 mm at 18 GHz). Using the phase response due to the offset trajectories, we proceed to calculate the array factor. The resulting pattern is shown in Fig. 4.4, for both the ideal (Fig. 4.2) and distorted trajectories (Fig. 4.3d). We observe that small errors in the positions of the elements can significantly degrade the accuracy of the array characterization. It is important to understand that these errors are not in the array itself, rather they are due to our inability to repeat the measurements for each element, as dictated by the UEAEP approach.
4.3 Implementation & Results

The aforementioned phase correction scheme is applied in the measurement of an 8×8 configuration of dipoles, operating in the Ku-band. Details of the antenna design can be found in Section 2.3. Following the UEAEP approach [69], far-field patterns of the individual elements are measured sequentially, with non-active ports terminated at a matched load. In as much as possible, the array was centered on the rotation platform, and was not moved between measurements. Nonetheless phase compensation is necessary to extract accurate patterns of the array under these conditions.

For pattern corrections, the raw element phase data, $\phi_i^M$, are first unwrapped using

$$
\phi_i^M(\theta') = \begin{cases} 
\phi_i^M(\theta') - 2\pi & \phi_i^M(\theta') > \phi_i^M(\theta' - \Delta \theta) + \pi \\
\phi_i^M(\theta') + 2\pi & \phi_i^M(\theta') < \phi_i^M(\theta' - \Delta \theta) - \pi \\
\phi_i^M(\theta') & \text{otherwise}
\end{cases}
$$

where $\phi_i^M(\theta' - \Delta \theta)$ is the measurement immediately preceding $\phi_i^M(\theta')$. The need for phase unwrapping implies that the angle step of the measurement setup ($\Delta \theta$) should be selected such that no element is moved more than 0.5λ in a single step, and thus avoid aliasing of the phase (i.e. $kr_i \sin(\Delta \theta) < \pi$). To meaningfully assess the data, we perform a minimum least-squares error regression of each element using the representation:

$$
R_i(\theta) = A_i \sin(\theta + B_i) + C_i
$$

(4.14)

The regression is weighted more heavily for the points occurring within ±60°, where received power is higher. The factor $C_i$ is an artifact from unwrapping, and is eliminated, resulting in a mean of zero across $[−180^\circ, 180^\circ]$. However, the mean over the observation region $[−90^\circ, 90^\circ]$ may be nonzero. Eight measured array elements, with corresponding best-fit sine curves are shown in Fig. 4.5.
We seek to compare the extracted parameters $A$ and $B$ from (4.14) against the constructs $r_i$ and $\theta_i$ from (4.5) and (4.6) in order to determine the values of $x$, $y$, and $\theta_0$. However, the polar equations defining each sine curve are individually under-determined, necessitating some additional assumption. Two options could be: 1) the element positions at grazing are known (namely, $R_i(\pm 90^\circ) = \mp d_i$), or 2) some offsets remain constant for more than one measurement.

The first assumption would unambiguously force the fitted curves into the ideal form, but assumes we have knowledge of the position of elements within the array, ignoring any errors in the array’s physical construction. It is important to note that the appropriate choice is specific to the particular measurement or fabrication scenario. Based on the conditions under which this data was collected, we will assume herein that $y$ is unique to each element, while $x$ changes slowly (effectively constant.
for each row of elements), and $\theta_0$ is constant for the entire array. As such, we can determine $\theta_0$ by finding the grazing angle. With this, (4.5) and (4.6) can be solved for $y$ and the quantity $(x + d_i)$. Finally, we resolve $x$ and $d_i$ by averaging a row of elements at the grazing angle ($x = 0$ implies the elements are evenly distributed across the center of rotation.)

Considering the data set shown in Fig. 4.5, after compensating for measurement errors we arrive at the final element paths, shown in Fig. 4.6. Notably, the corrected curves are similar to the ideal response (Fig. 4.2), but far from perfect. To illustrate the importance of this error extraction, we construct normalized beam patterns of the array using raw, corrected, and ideal phase data (Fig. 4.7). Immediately, we see that without removing measurement errors, the array characterization is completely inaccurate. After error correction, the measured results closely follow the expected pattern.

As this technique is conducted completely in post-processing, it gives an interesting measure of process accuracy. In Fig. 4.8, we plot the derived X- and Y-offset values for the measured elements. We note our assumptions proved accurate, as $y$ varies rapidly while $x$ remains relatively constant. Notably, the positional errors are on the order of $\pm 5$ mm, which would be effectively impossible to observe in situ without specialized equipment.
Figure 4.6: Element trajectories after removing measurement errors. Measurement data is shown in black, and fitted curves used for processing in blue.

Figure 4.7: Comparison of normalized array patterns using measured gains with ideal phase response (red), raw measured data (black), and error-corrected data (blue). Notably, the error-corrected data closely follows the expected, ideal response.
Figure 4.8: Derived measurement offsets in \( y \) (solid lines) and \( x \) (dashed lines). As expected for this measurement setup, \( x \) offsets remain relatively constant, while \( y \) offsets change with each measurement.
5.1 Summary of this Work

The goal of this research has been to develop UWB antenna arrays, which allow us to consolidate many separate functions into a single aperture. In particular, we have emphasized the myriad emerging applications in the millimeter-wave spectrum. In Chapter 1, we explored these applications, as well as existing UWB technologies and their limitations. Tightly coupled arrays, which utilize intentional capacitive coupling between neighboring elements to counteract the inductance of the groundplane, are found to provide the most bandwidth and scanning performance in a low profile setting. However, existing designs rely on a complex feed which cannot scale to millimeter-wave frequencies, and current measurement techniques suffer from low accuracy at these higher frequencies.

In Chapter 2 sources of inefficiency are investigated, including surface waves, cavity resonance, and Ohmic losses. The TCDA feed is modified to improve efficiency by eliminating the need for a split unit cell and corresponding power divider, and mitigating resultant resonances. Additionally, modifications are made to increase the size of the smallest features in the antenna feed, without increasing the form factor of the antenna. This is demonstrated to allow the new design to scale up 49 GHz,
representing a roughly $10 \times$ increase over the high frequency limit of previous designs. An $8 \times 8$ prototype operating over 3.5–18.5 GHz is fabricated and measured to validate the design. Infinite array simulations show VSWR < 2 across this band at broadside, with scanning to $\pm 45^\circ$ in the H-plane (VSWR < 2.6) and as far 70$^\circ$ in E-plane (VSWR < 2).

In Chapter 3 we present the development of a planarized TCDA and balun to extend operation into the millimeter-wave band. At millimeter-wave frequencies, planar co-fabrication of the entire array is critical to achieving repeatable fabrication, by eliminating the need for complex assembly at such small scales. Simultaneously, compatibility with low-cost PCB processes enables the potential for true mass-market applicability. The limitations of PCB fabrication are discussed, and a simplified balun is developed to overcome them. Optimized parameters are given for a sample design covering the allocated 5G and ISM bands across 24–72 GHz, with VSWR < 2.2 at broadside, and VSWR < 3 for $\pm 45^\circ$ scans in the E- and H-planes. The design is validated by fabrication of $3 \times 3$ and $5 \times 5$ prototype arrays through a commercial PCB vendor.

Finally, in Chapter 4 we develop a novel measurement technique, necessary for accurate characterization of antennas at or near millimeter-wave frequencies. This comprises a post-processing algorithm which serves to compensate the measured phase response of the array for millimeter-scale movements occurring during sequential measurements. This is developed mathematically, and is verified through measurements at Ku band (18 GHz).
5.2 Opportunities for Future Work

In this work new TCDA designs were developed, extending their applicability up to the millimeter-wave band. However, several opportunities for continued research exist, both in the design of UWB antennas, as well as in the development of backend systems to support these wide bandwidths.

5.2.1 Scalability Beyond 70 GHz

The primary limitation of the design presented in Section 3.2 is in the spacing of the balun vias. In particular, the in-line arrangement combined with limited edge-to-edge pitch (see again Table 3.1) results in a bounded unit cell size of > 2.03mm. Due to the grating lobe, this implies a limit of < 73.8 GHz (assuming λ/2 spacing). However, higher frequencies can be reached for the same fabrication process by altering the feed. In particular, we consider adding an offset to the feed via, as shown in Fig. 5.1a. This allows for a more compact unit cell, resulting in a reduced bound of > 1.58mm, or theoretically up to 94.8 GHz operation.

However, the offset layout results in elevated cross-polarized radiation at high frequencies. This is due to the asymmetry of the feed, which introduces fields orthogonal to the primary dipole current. Simulated co- and cross-polarized gain is plotted in Fig. 5.2. Notably, we observe that co- and cross-polarized radiation are nearly equal at the high end of the band, implying fields diagonally polarized with respect to the orientation of the dipoles. This can be resolved by use of a symmetrical double feed, as is shown in Fig. 5.1b. Indeed, it is expected that this would also serve to improve bandwidth by lowering the feed impedance. This approach still requires
Figure 5.1: Miniaturized unit cell design utilizing an offset feed via, scalable to 90 GHz (left). Asymmetry in the offset feed results in high cross-polarized gain at high frequencies, which can be reduced by duplicating the feed via to form a symmetric unit cell (right, shown without H-wall for clarity).
Figure 5.2: Co- and cross-polarized gain (Ludwig’s third definition) of the offset feed design shown in Fig. 5.1a. Notably, co- and cross-polarized gain are nearly equal at high frequencies, implying diagonal polarization.

Further optimization. Applying these techniques in conjunction with alternate fabrication techniques, such as microfabrication, could enable UWB antennas operating in the sub-millimeter wave band (> 100 GHz) or terahertz regime.

5.2.2 UWB Components and Systems

Ultimately, an UWB antenna is only the first component of a complete system, and the full utilization of such an antenna requires an equally wideband RF front-end. Several challenges are encountered when considering the transceiver electronics of an UWB phased array, in particular: 1) band-limited components, and 2) oversampling of the array at low frequencies. These may be overcome at a systems level through the use of frequency diplexing to channelize the wide bandwidth of the antenna. In this way, band limited components such as LNAs or PAs can be included in parallel to
support the extended bandwidth, while frequency dependent sub-arrays can be combined to reduce the burden of oversampling at low frequencies, without compromising high frequency performance. These concepts are shown in Fig. 5.3.

5.2.3 Hardware Reduction in High Gain Phased Arrays

While the dynamic capabilities of phased arrays could benefit virtually every wireless application, they are currently found predominately in defense applications. This is due to the high cost associated with independently controlling hundreds or thousands of antenna elements. Continued efforts to reduce the cost of all components within these systems, as well as novel systems level solutions to reduce the hardware burden, are critical to bringing this beneficial technology to widespread use.
Figure 5.3: Application of frequency diplexing to channelize the UWB antenna output, in order to: (a) overcome band-limited components, (b) reduce oversampling of low frequencies by forming frequency dependent subarrays.
Bibliography


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