

University of Central Florida

Electronic Theses and Dissertations, 2004-2019

2009

# Phase Shaping In The Infrared By Planar Quasi-periodic Surfaces Comprised Of Sub-wavelength Elements

James Ginn University of Central Florida

Part of the Electrical and Electronics Commons Find similar works at: https://stars.library.ucf.edu/etd University of Central Florida Libraries http://library.ucf.edu

This Doctoral Dissertation (Open Access) is brought to you for free and open access by STARS. It has been accepted for inclusion in Electronic Theses and Dissertations, 2004-2019 by an authorized administrator of STARS. For more information, please contact STARS@ucf.edu.

#### **STARS Citation**

Ginn, James, "Phase Shaping In The Infrared By Planar Quasi-periodic Surfaces Comprised Of Subwavelength Elements" (2009). *Electronic Theses and Dissertations, 2004-2019.* 4007. https://stars.library.ucf.edu/etd/4007



## PHASE SHAPING IN THE INFRARED BY PLANAR QUASI-PERIODIC SURFACES COMPRISED OF SUB-WAVELENGTH ELEMENTS

by

JAMES C. GINN, III B.S. University of Central Florida, 2004 M.S. University of Central Florida, 2006

A dissertation submitted in partial fulfillment of the requirements for the degree of Doctor of Philosophy in the School of Electrical Engineering and Computer Science in the College of Engineering and Computer Science at the University of Central Florida Orlando, Florida

Summer Term 2009

Major Professor: Glenn D. Boreman

© 2009 James Ginn.

### ABSTRACT

Reflectarrays are passive quasi-periodic sub-wavelength antenna arrays designed for discrete reflected phase manipulation at each individual antenna element making up the array. By spatially varying the phase response of the antenna array, reflectarrays allow a planar surface to impress a non-planar phasefront upon re-radiation. Such devices have become commonplace at radio frequencies. In this dissertation, they are demonstrated in the infrared for the first time - at frequencies as high as 194 THz. Relevant aspects of computational electromagnetic modeling are explored, to yield design procedures optimized for these high frequencies. Modeling is also utilized to demonstrate the phase response of a generalized metallic patch resonator in terms of its dependence on element dimensions, surrounding materials, angle of incidence, and frequency. The impact of realistic dispersion of the real and imaginary parts of the metallic permittivity on the magnitude and bandwidth of the resonance behavior is thoroughly investigated. Several single-phase reflectarrays are fabricated and measurement techniques are developed for evaluating these surfaces. In all of these cases, there is excellent agreement between the computational model results and the measured device characteristics. With accurate modeling and measurement, it is possible to proceed to explore some specific device architectures appropriate for focusing reflectarrays, including binary-phase and phase-incremental approaches. Image quality aspects of these focusing reflectarrays are considered from geometrical and chromatic-aberration perspectives. The dissertation concludes by briefly considering two additional analogous devices - the transmitarray for tailoring transmissive phase response, and the emitarray for angular control of thermally emitted radiation.

To Mom and Dad, I would not be here if not for you.

### ACKNOWLEDGMENTS

I would like to thank Dr. Glenn Boreman, for whom I am grateful for the opportunity to participate in the IR Systems Lab and who has always encouraged and aided in my research. I would also like to thank Dr. Brian Lail, for his constant guidance and support beginning when I was an undergraduate freshman learning the principles of Electrical Engineering.

I would like to acknowledge Dr. Javier Alda, for agreeing to sit an office with three graduate students during his sabbatical and providing a voice of reason during our many discussions. I would also like to acknowledge Dr. Parveen Wahid, for agreeing to serve on my committee and providing insight during the development of my dissertation.

This research was funded in part by Lockheed Martin Missile and Fire Control in Orlando, Florida. I would like to specifically recognize Edit Braunstein, Darren Zinn, Gene Tener, and Thomas Haberfelde for their help.

I would like to thank the other members of the FSS team, Dr. Jeff Tharp and Dave Shelton, for providing an invaluable source of support and insight through out the entire thesis process.

I would like to acknowledge several members of the IR Systems Lab and UCF Engineering that provided help during my dissertation: Dr. Todd Du Bosq, Wilson Caba, Li Chen, Dr. Ivan Divliansky, Bill Franklin, Dr. Ray Folks, Peter Krenz, Dr. Jose LopezAlonso, Dr. Tasneem Mandviwala, Rajesh Paryani, Dr. Chris Middlebrook, Dr. Charles Middleton, Dan Mullally, John Reeder, Sid Pandey, Brian Slovick, Ismael Quijano, and Guy Zummo. I am honored to have worked with all of you.

I cannot forget my friends and family, for always being there for me and providing me with a constant source of personal inspiration.

Mom and Dad, I can never thank you enough.

I must thank my fiancée, Elayne, for supporting me throughout the dissertation process with her love, patience, and humor.

# TABLE OF CONTENTS

LIST OF FIGURES	i
LIST OF TABLES	X
LIST OF SYMBOLS/ABBREVIATIONS	X
CHAPTER 1: INTRODUCTION	1
1.1 Electromagnetic Phase Shaping, Interference, and Array Behavior	1
1.2 State of the Art Resonant Phase Shaping Elements and Arrays	5
1.2.1 Resonant Polarizers	0
1.2.2 Phased Arrays	2
1.2.3 Reflectarrays 1:	5
1.3 Thesis	8
1.4 Prior Publication and Financial Support Disclosure	9
CHAPTER 2: THE INFRARED RESONANT PHASED ELEMENT	0
2.1 Metal-Dielectric Resonance Modes	0
2.2 Relating Resonance Properties to Phase	2
2.3 Chromatic Dependence of Resonance Modes	3
2.4 Angular Dependence of Resonance Modes	5
2.5 Computational Electromagnetic Modeling of Infrared Phased Elements	7
2.5.1 Finite Element Model	9
2.5.2 Periodic MoM Model	3
2.6 Fabrication of Infrared Phased Elements	5
CHAPTER 3: METAMATERIAL RESONANCE DAMPING 4	0
3.1 Infrared Metamaterial Dependence on Dispersive Materials	0

3.1.1 Material Characterization	41
3.1.2 Implementation of Dispersion in Full-Wave Simulation	42
3.1.3 Example Results: Square Loop FSS on ZrO <sub>2</sub>	44
3.1.4 Example Results: Square Loop FSS on a Polymer	
3.1.5 IR Material Dispersion: Conclusions and Applications	
3.2 Theoretical Foundation of Metamaterial Damping Theory	49
3.2.1 Thin-Film Resistor Impedance at Infrared Frequencies	50
3.2.2 Equivalent-Circuit Resonator Derivation	54
3.3 Simulated Validation of the Metamaterial Damping Theory	57
3.3.1 Real Frequency-Independent Conductivity	58
3.3.2 Complex Frequency-Independent Impedance Surfaces	59
3.3.3 Complex Frequency-Independent Conductivity	60
3.3.4 Measured Metal-Film Properties	
3.4 Measured Validation of the Metamaterial Damping Theory	66
3.5 Implications of Metamaterial Damping Theory	68
3.5.1 Impact of Relaxation Time on Damping and Film Reactance	68
3.5.2 Bandwidth Considerations	71
3.5.3 Phase Considerations	
CHAPTER 4: THE INFRARED REFLECTARRAY	
4.1 Fundamental Theory	73
4.1.1 The Variable-Patch Reflectarray Element	73
4.1.2 Waveguide Equivalent Circuit Model	75
4.1.3 Improving Bandwidth	79

4.1.4 Achieving Phase Variation Greater Than 360°	
4.1.5 The Dielectric Reflectarray	
4.1.6 Circularly Polarized Reflection	
4.1.7 Focusing Reflectarray Layout Development	
4.1.8 Additional Fabrication Requirements	
4.2 LWIR Reflectarray	
4.2.1 Initial Element Development	
4.2.2 Low Loss Element Development	
4.2.3 Continuous Reflectarray	107
4.2.4 Binary Focusing Reflectarray	109
4.2.5 Two-Element Focusing Reflectarray	
4.2.6 Eight-Element Focusing Reflectarray	
4.3 MWIR Reflectarray Element Development	
4.4 NIR Reflectarray Element Development	
4.5 Aberration Behavior	
4.5.1 Angular	
4.5.2 Chromatic	
CHAPTER 5: THE INFRARED TRANSMITARRAY	
5.1 Fundamental Theory	
5.2 Single Layer Transmitarray	
5.3 Feasibility and Alternatives To The Infrared Transmitarray	
CHAPTER 6: THE INFRARED EMITARRAY	
6.1 Fundamental Theory	

6.1.1 Relating Thermal Performance to Electromagnetic Performance	. 132
6.1.2 Phasing Thermal Emission	. 133
6.2 Polarized Emission	. 134
6.2.1 Linearly Polarized Emission by Asymmetric Resonant Elements	. 135
6.2.2 Circularly Polarized Emission by a Compact Multi-Layer Structure	. 137
6.2.3 Circularly Polarized Emission by a Single-Layer Structure	. 140
6.3 Mutually Coupling Induced, Long-Range Coherent Emission: The Emitarray	. 144
CHAPTER 7: CONCLUSIONS	. 147
7.1 Summary	. 147
7.2 Future Work	. 149
APPENDIX A: PHASE CHARACTERIZATION IN THE INFRARED	. 152
A.1 The Variable Angle Spectral Ellipsometer	. 153
A.1.1 The Variable Angle Spectral Ellipsometer Fundamentals	. 153
A.1.2 IR-VASE Characterization of the Optical Properties of Thin Films	. 154
A.1.2.1 Sample Preparation	. 154
A.1.2.2 Minimizing Backside Reflections	. 155
A.1.2.3 Ensuring Unique/Realistic Solutions When Modeling	. 156
A.1.2.4 Miscellaneous Tips	. 157
A.1.3 IR-VASE Characterization of Phased Devices	. 158
A.2 The Twyman-Green Interferometer	. 158
A.2.1 Identifying Phase Shaping Elements in Interferograms	. 162
A.2.2 Analyzing Interferograms for Phase Information	. 164
A.2.3 Interferogram Analysis Post Processing	. 166

A.2.4 Interferogram Analysis using MATLAB	168
A.2.5 Interferometry of Transmissive Devices	
A.3 At-Focus Spot Imaging	171
A.4 The Infrared Microscope	172
APPENDIX B: GDSII TOOLBOX	174
B.1 About the GDSII Toolbox	
B.2 Using the GDSII Toolbox	
B.3 Creating New GDSII Toolbox Modules	
B.4 Limitations of the GDSII Toolbox	
LIST OF REFERENCES	

# LIST OF FIGURES

Figure 1: Young's sketch of the interfering wavefronts of two slit sources
Figure 2: Parabolic reflector with reflected planar wavefront
Figure 3: Soret FZP layout
Figure 4: Meanderline polarizer schematic
Figure 5: Phased array with spherical radiated wavefront
Figure 6: Planar reflectarray mimicking the reflected wavefront of a parabolic reflector.
Figure 7: Current modes on two microstrip antenna elements in an infinite array at
resonance (standing wave present)
Figure 8: The near-field electric field pattern from the two microstrip antenna elements in
Figure 7 demonstrating coupling
Figure 9: Band gap behavior of an arbitrary resonant, groundplane-backed antenna array.
Figure 10: Phase behavior of an arbitrary resonant, groundplane-backed antenna array. 24
Figure 11: Summary of design rule of thumbs for resonant, groundplane-backed antenna
array
Figure 12: Reflectivity vs. incident angle for an example groundplane-backed antenna at
28.28 THz
Figure 13: Phase vs. incident angle for an example groundplane-backed antenna at 28.28
THz
Figure 14: IAWM waveguide cross section with electric field polarization
Eigung 15: Madal lawayt fan EEM transmissive daeien

Figure 16: Model layout for FEM reflective design.	. 32
Figure 17: Example of layer stack-up in Ansoft Designer.	. 34
Figure 18: Model layout for periodic MoM modeling	. 35
Figure 19: Standard substrate coating process	. 36
Figure 20: Resist coating and e-beam exposure.	. 38
Figure 21: Resist development to reveal pattern	. 38
Figure 22: Metal evaporation and lift-off process.	. 39
Figure 23: Implementation of frequency dependent modeling	. 43
Figure 24: SEM image of fabricated square loop FSS on ZrO <sub>2</sub> .	. 44
Figure 26: Measured, frequency independent PMM, and frequency dependent PMM	
results for square loop FSS on ZrO <sub>2</sub> .	. 45
Figure 27: Frequency independent PMM and frequency dependent PMM results for	
square loop on plastic	. 47
Figure 28: Frequency dependent index of refraction (n,k) of polymer dielectric from	
ellipsometer	. 48
Figure 29: Resonant circuit equivalent model for an infinite array of dipole metamater	ial
elements.	. 55
Figure 30: Dimensions of cross metamaterial element	. 57
Figure 31: Modeled transmission of the cross element with four frequency independent	ıt
conductivities.	. 59
Figure 32: Modeled transmission of the cross element with four frequency-independent	ıt
complex sheet resistances.	. 60

Figure 33: Modeled transmission of the cross element with four frequency-independent	nt
complex permittivities.	62
Figure 34: Measured real part of permittivity for gold, aluminum, and nickel	64
Figure 35: Measured loss tangent for gold, aluminum, and nickel	64
Figure 36: Calculated skin depth for gold, aluminum, and nickel	65
Figure 37: Modeled transmission of four cross metamaterial arrays made of three	
different metals and PEC.	65
Figure 38: SEM micrograph of a portion of the gold cross metamaterial array	67
Figure 39: Measured (solid) and modeled (dotted) transmission of three cross	
metamaterial arrays made up of three different metals	67
Figure 40: The complex impedance of two bulk metals with relaxation time of 10 or 1	00
fs in the IR and THz	70
Figure 41: Variable-patch reflectarray schematic	74
Figure 42: Reflectarray transmission line equivalent.	76
Figure 43: Smith chart plot of $\Gamma_{in}$ demonstrating reflectarray phase transition by varying	<u>1g</u>
patch size	78
Figure 44: Example multi-band reflectarray layout	80
Figure 45: Stack-up for a multi-layer reflectarray.	81
Figure 46: Dipole reflectarray with greater than 360 degrees of phase delay.	83
Figure 47: Dielectric reflectarray at 28.3 THz.	84
Figure 48: Modeled phase performance of dielectric reflectarray	85
Figure 49: Modeled reflection and phase properties of a circularly polarized reflectarra	ay
element	86

Figure 50: Ideal focusing reflectarray layout	
Figure 51: 10.6 $\mu$ m interferogram of a typical 380 $\mu$ m thick prime-grade silicon wafer	
exhibiting significant surface curvature	90
Figure 52: Custom e-beam holder for thick substrates	91
Figure 53: Proof of concept reflectarray layout	92
Figure 54: SEM image of metalized dose matrix.	93
Figure 55: Visible image of metalized dose matrix	94
Figure 56: Fabricated strip reflectarray with reference sizes	95
Figure 57: SEM micrograph of one of the stripes of the fabricated reflectarray	95
Figure 58: Visible micrograph of one of the stripes of the fabricated reflectarray	96
Figure 59: Interferogram of coated wafer	97
Figure 60: Interferograms of fabricated reflectarrays	97
Figure 61: Smoothed phase response results for first reflectarray	99
Figure 62: Smoothed phase response results for second reflectarray 1	00
Figure 63: Modeled (solid line) vs. measured (dark squares) results for initial element	
development1	02
Figure 64: New layout prototype on a silicon wafer imaged below previous prototype o	n
an optical flat 1	03
Figure 65: Modeled reflectarray reflectivity with varying standoff layer film thickness.	
	04
Figure 66: Modeled results for new layout (black line) vs. old layout (grey line) 1	05
Figure 67: Improved interferometer characterization capability demonstrating zoom,	
better edge transition resolution, and uniform illumination 1	06

Figure 68: Improved efficiency of new layout with reflectarray regions denoted by red
box
Figure 69: Square loop reflectarray element with reflectarray regions denoted by red box.
Figure 70: Improved phase selection by mixing elements
Figure 71: GDSII layout of continuous reflectarray
Figure 72: Continuous phase variation reflectarray device measured at 10.6 µm 108
Figure 73: Visible micrograph of (a) reflectarray rings with layout schematic and (b)
patch elements in the rings
Figure 74: (a) Reflected beam profiles of reflectarray at optimal focus and (b) reflectarray
outside of optimal focus
Figure 75: 8 element graded reflectarray
Figure 76: 3.39 $\mu$ m interferometer setup using portions of the 10.6 $\mu$ m optical train 113
Figure 77: Four stripes of the $3.39\mu m$ reflectarray prototype demonstrating unique fringe
shifting at each stripe114
Figure 78: Measured and modeled results for the 3.39 µm reflectarray prototype 114
Figure 79: 1.55 µm interferometer source and testing setup
Figure 80: Two stripes of 1.55µm reflectarray prototype demonstrating unique fringe
shifting at each stripe
Figure 81: Measured and modeled results for the 1.55µm reflectarray prototype 117
Figure 82: Images of binary reflectarray aberration behavior: (left) image of blur spot
with detector placed inside focus and (right) detector placed outside focus
demonstrating astigmatism

Figure 83: Future design method for aberration corrected reflectarray
Figure 84: Planar transmitarray focusing a planar wavefront 125
Figure 85: Square patch transmitarray modeled performance
Figure 86: Patch slot transmitarray modeled performance
Figure 87: Cross slot transmitarray modeled performance
Figure 88: UCF Pegasus logo with dipole orientations
Figure 89: Images of the linearly polarized Pegasus with the camera's linear polarizer (a)
horizontally oriented and (b) vertically oriented
Figure 90: Images of the linearly polarized Pegasus with the camera's linear polarizer not
present
Figure 91: State-of-the-art circular polarizer for emitted radiation
Figure 92: Compact, multi-layer circular polarizer for emitted radiation 140
Figure 93: Asymmetric tripole
Figure 94: Modeled emission and phase properties of asymmetric tripole 142
Figure 95: Visible micrograph of four-quadrant checkerboard of asymmetric tripoles. 143
Figure 96: Two thermal images (side by side) of checkerboard asymmetric tripole with
linear polarizer filter present in front of the camera in two orthogonal directions. 143
Figure 97: Two thermal images (side by side) of checkerboard asymmetric tripole with
circular polarizer in front of the camera and in two orthogonal directions
Figure 98: Thermal image of reflectarray at focus
Figure 99: Thermal image of reflectarray at focus with slight tilt added to the hot plate.
Figure 100: J. A. Woollam IR-VASE

		٠	٠
X	V	1	1

Figure 101: Twyman-Green interferometer schematic [120]	9
Figure 102: Measured interferogram of a nearly flat surface	2
Figure 103: Measured interferogram of a region of phase shifting elements	3
Figure 104: Measured interferogram of two arrays of phase shaping elements with one	
array having a phase variation greater than 180 degrees	4
Figure 105: Ideal interferogram of two phase-shaping arrays	5
Figure 106: Ideal interferogram of two phase-shaping arrays with tilt added	8
Figure 107: AFSI test setup 17	1
Figure 108: Infrared microscope	3

# LIST OF TABLES

Table 1: Comparison of runtime for a square loop FSS using 100 frequency points 46
Table 2: Frequency-independent sheet resistances used in modeling and modeled
resonance frequency and bandwidth
Table 3: Frequency independent permittivities used in modeling and derived loss tangent
and skin depth values61
Table 4: Calculated bandwidths for the three measured-metal designs in Figure 37 72
Table 5: Measured relative phase shift vs. reflectarray patch size.    101
Table 6: Measured emitted power from a dipole array heated to 100 °C 136

## LIST OF SYMBOLS/ABBREVIATIONS

AF	Array Factor
AFSI	At-Focus Spot Imaging
Al	Aluminum
AMC	Artificial Magnetic Conductor
Au	Gold
BCB	B-staged Bisbenzocyclobutene
CEM	Computational Electromagnetic Modeling
dB	Logarithmic Power Ratio
dc	Direct Current
deg	Degree
<i>E</i> <sub><i>r</i></sub>	Dielectric Constant
e-beam	Electron Beam
EBG	Electromagnetic Band Gap
eFSS	Emission Frequency Selective Surface
<i>F/</i> #	F-number
FD	Frequency Dependent
FEM	Finite Element Method
FLAPS <sup>TM</sup>	Flat Parabolic Surface
FSS	Frequency Selective Surface
fs	Femtosecond (10 <sup>-15</sup> second)
FWHP	Full Width Half Power
FZP	Fresnel Zone Plate

GDSII	Graphic Data System II Database
GIMP	GNU Image Manipulation Program
GRIN	Graded Index
GUI	Graphical User Interface
НТ	High Tension
HeNe	
IAWM	Infinite Array Waveguid Method
IR	Infrared
IR-VASE	Infrared Variable Angle Spectral Ellipsometer
IPA	Isopropyl Alcohol
kV	Kilovolt (10 <sup>3</sup> Volt)
LWIR	Long-Wave Infrared (8 – 15 µm)
m	Meter
μm	Micrometer (10 <sup>-6</sup> Meter)
μC	Microcoulomb (10 <sup>-6</sup> Coulomb)
mm	Millimeter (10 <sup>-3</sup> Meter)
Mn	Maganese
MoM	Method of Moments
MPIE	Mixed-Potential Integral Equation
MSE	
MWIR	Mid-Wave Infrared (3 – 8 μm)
NIR	Near Infrared (0.75 – 3 μm)
OPD	Optical Path Difference

UI	User Interface
Ni	Nickel
nm	Nanometer (10 <sup>-9</sup> Meter)
nA	Nanoamp (10 <sup>-9</sup> Ampere)
PEC	Perfect Electric Conductor
РМС	Perfect Magnetic Conductor
PML	Perfectly Matched Layer
PMM	Periodic Method of Moments
PtSi	Platinum Silicide
PVD	Physical Vapor Deposition
px	pixel
rf	Radio Frequency
RGB	Red Green Blue (Color Model)
S	Siemens
SEM	Scanning Electron Microscope
SIC	Solver Independent Code
$\delta_{skin}$	Skin Depth
SSC	Solver Specific Code
$tan(\delta)$	Loss Tangent
THz	Terahertz (10 <sup>12</sup> Hertz)
Ti	Titanium
UAV	Unmanned Aerial Vehicle
VBScript	Visual Basic Script

ZDMAC	N,N-Dimethylacetamide
ZEP RD	Xylene (o-, m-, p- mixed)
ZEP Resist	Methyl Styrene / Chloromethyl Acrylate Copolymer
ZrO <sub>2</sub>	Zirconium Dioxide

### **CHAPTER 1: INTRODUCTION**

### 1.1 Electromagnetic Phase Shaping, Interference, and Array Behavior

Time-harmonic electric fields have been classically described by four terms: position or direction (x, y, z or a similar coordinate system), time (t), magnitude (E), and sinusoidal frequency (f) [1]:

$$\widetilde{E}(x, y, z; t) = \operatorname{Re}[\vec{E}(x, y, z)e^{j2\pi f t}]$$
(1)

From this expression, and knowledge of the medium containing or confining the electric field, several other electromagnetic quantities can be derived using Maxwell's equations [2] and the wave equation, including the corresponding magnetic field, electric flux density, magnetic flux density, and electric current density. These relations and terms are the fundamental basis of a wide range of science and engineering disciplines including wave propagation, antenna theory, and circuits.

Interference arises when two propagating time-harmonic fields interact. Consider a propagating electric field of the form:

$$\widetilde{E}_1(z) = (a\hat{x} + b\hat{y})E_0 e^{j\gamma z} e^{j2\pi ft}$$
(2)

If a second, identical propagating field is delayed by an arbitrary time delay  $t_0$ , the new field can be expressed by:

$$\widetilde{E}_{2}(z) = (a\hat{x} + b\hat{y})E_{0}e^{j\gamma z}e^{j2\pi f(t-t_{0})}$$
(3)

which, when superimposed with the first field, yields:

$$\widetilde{E}(z) = (a\hat{x} + b\hat{y})E_0 e^{j\gamma z} e^{j2\pi ft} [1 + e^{j\theta}]$$
(4)

where  $\theta$  is equal to  $-2\pi f t_0$  and represents the phase delay between the two propagating waves. The magnitude of the resulting electric field (neglecting position and time) is equal to:

$$|\widetilde{E}| = E_0[1 + \cos(\theta)] \tag{5}$$

Thus, when the phase delay is  $2n\pi$  (n = 0, ±1, ±2...), the fields will constructively interfere completely and double the magnitude of the resulting propagating field. When the phase delay is  $2n\pi + \pi$  (n = 0, ±1, ±2...), the fields will destructively interfere completely and no propagating field will be present. The specific case of interference over time described is known as temporal interference [3] or interference by division of wave amplitude [4].

A more practical application of electromagnetic interference is the interference that arises between two coherent point sources, also known as spatial interference or interference by division of wavefront [4]. In spatial interference, the optical path difference between the two sources result in a spatially varying relative phase difference between the two coherent radiated fields and, subsequently, a predictable change in field magnitude as a function of position. Thomas Young carried out what is generally regarded as the bestknown demonstration of spatial interference [5]. In his double slit experiment, Young inadvertently verified the dual nature of photons as particles and waves by shining light through two diffractive slits and imaging the resulting periodic dark and bright interference bands on a planar screen placed behind the slits. The experiment was successful due to the slits behaving as coherently fed Huygens' sources [6], which radiated the necessary spherical wavefronts that gave rise to the periodic interference pattern observed (Figure 1). His experiment has been modified and repeated for a multitude of other elementary particles (electrons, protons, large molecules [7], etc.) and for more than two interfering slits.



Figure 1: Young's sketch of the interfering wavefronts of two slit sources. Public domain image.

Electromagnetic phase shaping, for the purpose of selective spatial interference, is of significant interest to antenna and optics designers. Through a geometrical arrangement of radiating sources with or without a selective feeding phase delay, it is possible to give rise to wavefront shaping or beam steering due to interference processes alone, without the need of utilizing physical path differences like as is the case with refractive surfaces. Following the array development procedure outlined by Balanis [8], the electric field radiated by the superimposition of an N x N planar array of identical, coherent, non-coupled sources can be expressed by:

$$E_{array} = E_{element} \times AF \tag{6}$$

where  $E_{element}$  is the electric field contribution of each source, AF is an expression that describes the array layout, and  $E_{array}$  is resulting electric field radiated by the array. The

normalized array factor for a periodic phased array in the x-y plane can be represented by:

$$AF_{n} = \left[\frac{1}{N} \frac{\sin(\frac{N}{2}\Psi_{x})}{\sin(\frac{\Psi_{x}}{2})}\right] \left[\frac{1}{N} \frac{\sin(\frac{N}{2}\Psi_{y})}{\sin(\frac{\Psi_{y}}{2})}\right]$$
(7)

$$\Psi_x = kd\sin\theta\cos\phi + \beta_x \tag{8}$$

$$\Psi_{y} = kd\sin\theta\sin\phi + \beta_{y} \tag{9}$$

where *d* is the distance between each element, *k* is the wave number, and  $\beta$  is the progressive phase difference of the excitation field between each element. Assuming the sources radiate with the same field magnitude, altering the progressive phase difference between each element will define the direction of the main lobe of the array's radiation ( $\theta_0$ ,  $\Phi_0$ ), or:

$$\beta_x = -kd\sin\theta_0\cos\phi_0\tag{10}$$

$$\beta_y = -kd\sin\theta_0\sin\phi_0 \tag{11}$$

From these relations, it is clear that the altering the spacing and phase delay of the elements making up a planar, periodic arrangement of sources can give rise to selective radiation directivity and wavefront shaping. Thus, the goal of this dissertation is to investigate the use of passive, sub-wavelength resonant elements to achieve selective wavefront shaping in the infrared. All aspects of phase manipulation will be considered – reflection, transmission, and emission.

#### **1.2 State of the Art Resonant Phase Shaping Elements and Arrays**

Beamforming antennas are a sub-class of antennas comprising of an array or quasi-array of antenna elements characterized by their ability to collectively exhibit a narrow radiated beamwidth or a focusing wavefront [9]. Beamforming antennas are highly directive antennas that find heavy usage at microwave frequencies where omni-directional antennas would be undesirable, such as in satellite communication systems, line of sight transmission systems, and scannable detectors. At higher frequencies, including visible and infrared (IR) frequencies, beamforming antennas are often deployed as focusing elements, directional detectors, and collimators.

Reflector antennas (Figure 2) are one of the oldest and simplest known beamforming antenna devices and have been deployed, in various forms, across the entire electromagnetic spectrum. In its most basic configuration, the reflector antenna is simply a reflective surface of arbitrary geometry impinged upon by electromagnetic radiation. Thus, the planar mirror could be viewed as the earliest developed reflector antenna. From the planar mirror, additional reflector antenna geometries have been developed at visible frequencies, such as the focusing, spherical mirror. Reflector antenna development at microwave frequencies did not began to mature until World War II to meet the needs of the emerging field of radar [10]. Now sold commercially for use in receiving telecommunication and television signals, radio frequency (rf) reflector antennas continue to increase in popularity as an integral part of most satellite communication systems [11]. The main advantage for reflector antennas, aside from simplistic design procedure, is their potential for high radiation efficiency. Across the electromagnetic spectrum, several high conductivity materials are available that allow for low-loss upon reflection by the reflector antenna. Without active components, reflector antennas are well suited for high power applications and rarely need maintenance. In addition, with the capability of collimation, reflector antennas exhibit high gain – often exceeding 60 dB at rf frequencies [11].



Figure 2: Parabolic reflector with reflected planar wavefront.

Even with its numerous advantages, the reflector antenna exhibits several distinctive limitations. In the traditional spherical and parabolic configurations, the reflector is inherently bulky. Unlike planar antennas, the parabolic reflector, traditionally physically thicker to maintain proper shape, prohibits folding for transport. Resulting from the fact that wavefront modification occurs due to physical path length differences, reflector antennas cannot be made conformal, which restricts deployment on mobile structures, such as vehicles or aircraft, where drag may become a concern. At shorter wavelengths, the cost of fabrication will increase because of enhanced sensitivity to height variations on the antenna surface. In addition, diffraction at the edge of the reflector becomes a concern at short wavelengths. Additionally, utility stacking of reflector antennas and other devices, such as frequency selective surfaces (FSS), is difficult or completely impossible.

To overcome the bulky nature of refractive optics, diffractive surfaces have seen widespread interest in development due to their planar nature. Classical diffractive focusing surfaces, commonly known as Fresnel zone plates or FZP, consist of a series of concentric rings spaced to transform normally incident planar radiation into a spherical wave through diffraction from the concentric grating and the resulting far-field interference [6]. Soret proposed the earliest FZP comprised of repeated regions of high transmission and low transmission (Figure 3) [12]. These regions were spaced by 180 degrees of relative phase to ensure that only the components of the incident wavefront that led to constructive interference were transmitted and the destructive components were reflected. Wood later suggested that greater throughput could be achieved if the opaque zones would instead be retarded by 180 degrees allowing them to interfere constructively [12]. Wood also noted that the binary nature of the early FZP would lead to multiple, unwanted higher order foci [12]. To reduce the power in the higher-order foci, numerous phase corrected zone plates have been developed with repeated rings with relative phase difference smaller than 180 degrees [13]. The smaller relative phase steps result in smaller wavefront errors and, subsequently better image quality.



Figure 3: Soret FZP layout. Black segments denote regions of high reflectivity.

There are several problems associated with FZPs. Their best known issue is their extremely narrowband operation due to the wavelength dependence of their concentric ring layout. Another common pitfall with FZP design is achieving phase variations of less than 180 degrees for improved image quality. One method for reducing undesirable foci, while maintaining a planar surface, has been to employ phase correction through index grading [6]. In a graded-index FZP, the number of discrete phase states in the structure is increased through the progressive variation of the index of refraction of a dielectric material contained in each zone of the FZP. Unfortunately, this approach is difficult to implement in the infrared and terahertz band of the spectrum owing to the lack of practical, low-loss dielectric materials of varying refractive index. Instead, infrared designs have required the implementation of non-planar dielectric kinoform FZPs [14], which utilize discrete height grading for phase correction or shaping across the zones of

the diffractive element. The need for three-dimensional lithography in kinoform FZPs greatly increases the cost and complexity of these devices when compared to binary FZPs.

Thus, it is of high interest to develop a new method for designing focusing elements that do not require bulky refractive elements or difficult-to-fabricate kinofrom FZPs. One such method is to employ resonant antenna elements to replicate the behavior of the phase corrected FZP. Through a variety of methods, resonant elements can be tuned to exhibit a desired relative phase delay upon radiation, while maintaining entirely planar surfaces. The resonant behavior of the antenna elements also opens the door for subwavelength phase control for integrated aberration correction in the infrared. The next three sections will discuss some state-of-the-art phased antenna structures, both passive and active, currently utilized in the rf and infrared bands.

It should also be noted that passive resonant elements have additional uses outside of phase shaping. Resonant planar antenna devices have been used for years as spectral filters, either as frequency selective surfaces (FSS) [15] or electronic band gap devices (EBG) [16]. Through a phenomenon discussed later in the dissertation, FSS can also be developed to exhibit engineered emission properties when heated. Recent research has expanded into investigating the usage of these surfaces to exhibit negative index of refraction, known as negative index materials or NIM [17], to exhibit extremely low indices of refraction, known as near-zero materials or NZM [18], or to exhibit perfect magnetic conductivity, known as artificial magnetic conductors or AMC [19].

#### **1.2.1 Resonant Polarizers**

One of the simplest planar, phased element devices that is well documented in the literature is the resonant polarizer. These surfaces have been utilized in the microwave portion of the spectrum as engineered quarter-wave plates or linear-to-circular polarizers [20]. Polarization conversion is achieved by introducing an asymmetry in the resonant element to delay one orthogonal component of the linearly polarized incident radiation by 90 degrees, while maintaining equal power in each component, which results in reradiated circular polarization. Because of this, the name resonant polarizer is a bit of a misnomer as only a capacitive, inductive, or physical path delay is necessary to achieve the desired circular polarization. The surfaces; however, still exhibit a preferred band of operation and share a physical structure analogous to most frequency selective surfaces. Several element designs have been developed for the purpose of linear-to-circular polarization conversion including the meanderline polarizer [21], the twist reflector [22], and the waffle-grid polarizer [23].

Of the three mentioned polarizers, the meanderline polarizer is the most applicable to the research presented in the dissertation. First published in the 1970s [24], the meanderline polarizer consists of a periodic, sub-wavelength stepped metal grating (Figure 4). Decomposing a linearly polarized impinging field at 45 degrees relative to the direction of a meanderline, yields two equal magnitude, orthogonal fields about the horizontal and vertical axis of the meanderline. The horizontal component of the meanderline will experience a capacitive delay due to the inter-element coupling between each meanderline, while the vertical component will experience an inductive delay due to the

physical length of the meanderline. When these capacitive and inductive leads are selected properly, it becomes possible to introduce a 90 degree phase delay between the two orthogonal states with the intent of inducing circular polarization. Furthermore, because each state is dominated by one reactive component and the cross-polarization of the meanderline is low, minimal loss will be experienced since the meanderline is unable to resonate and will not form a band gap.



Figure 4: Meanderline polarizer schematic. Horizontal in this figure is defined from the left to right on the page and vertical is defined from top to bottom of the page.

Resonant polarizers have only recently been developed to operate in the infrared. Work by Dr. Jeff Tharp at the Infrared Systems Lab has verified the operation of a meanderline polarizer in the mid-wave infrared  $(3 - 5 \ \mu m)$  [25] and the long-wave infrared  $(8 - 12 \ \mu m)$  [26] [27]. Meanderline surfaces are desirable in the infrared for their low footprint, high design flexibility, and low fabrication cost compared to similarly behaved birefringent crystals In addition, the surface exhibits excellent bandwidth, acceptable angular dispersion, and the potential for reflective phase shaping [28].

### 1.2.2 Phased Arrays

Phased arrays (Figure 5) share several similarities with reflector antennas in terms of their ability to control radiated wavefronts and exhibit high directivity. Phased arrays are traditionally active, planar devices, which, by introducing a progressive feed phase delay between neighboring elements, will radiate only in a specified direction or with a specific wavefront [29]. The simplest phased array consists of a large array of horn antennas fed by a single source, but with each antenna connected to the source by a different length of waveguide. The waveguide length difference between the aperture antennas introduces a phase delay in the radiated fields of each horn, which, in turn, changes the array's farfield pattern to add constructively in only a specific direction or exhibit an arbitrary wavefront. Planar, microstrip phased arrays behave in exactly same way, with progressive phase difference introduced by waveguide length or lumped elements.


Figure 5: Phased array with spherical radiated wavefront.

From the array relations developed in the previous section, it is clear that antenna's directivity depends on both the inter-array element spacing and the directivity of an element of the array. Thus, phased arrays can be designed to achieve high directivities comparable to the directivity of a reflector antenna. Additionally, the utilization of a non-uniform array arrangement allows for further control of the radiated far-field pattern and is one method to introduce non-planar wavefronts.

The significant difference between the reflector antenna and the phased array is that the phased array is not dependent on a physical height difference to alter the radiated wavefront. This allows the phased array to be significantly thinner than the reflector antenna and allows the antenna to be designed for conformal deployment with correction for height differences occurring in the progressive phase delay [30]. Phased arrays can also be tuned by introducing a progressive phase delay at each radiating element. These

devices can be designed to radiate more complicated wavefronts than conventional reflective or refractive surfaces.

Unlike reflector antennas, phased arrays exhibit many of the limitations common to most planar antennas. Often inefficient with low power excitation, phased arrays cannot handle high power sources without physical breakdown [8]. Additionally, phased arrays are inherently active devices and, thus, cannot be utilized as an intermediate focusing element. Analysis of phased arrays, especially when accounting for coupling or when using non-uniform arrays, is far more complicated than reflector antennas and will often demand the use of a numerical analysis.

Similar to the resonant polarizer, phased arrays are another newly developed infrared technology [31]. Phased arrays are desirable in the infrared as a potential method to achieve lensless imaging – utilizing the directionality of the phased array to achieve high directivity and directionality without a need for a lens [32]. Presently, phased arrays in the infrared have to rely on antenna-coupled bolometers [33] for detection, although alternative mechanisms could be employed including metal-oxide-metal [34] or Schottky [35] diodes. Phased arrays are also an important tool for simultaneously measuring phase and magnitude in the infrared using square-law sensors [36] and near-field probing [37]. Currently, tunable infrared phased arrays are also under development allowing for electronically-controllable field of view detectors.

## 1.2.3 Reflectarrays

The reflectarray antenna, in its most simple form, is passive, planar microstrip antenna array designed for reflected beamforming. By varying the re-radiated phase response across the surface through variation of local geometry of the array elements, reflectarrays allow a planar surface to radiate a wavefront of arbitrary shape upon reflection (Figure 6). Similar to the phased array described in 1.2.2, the far-field pattern of the reflectarray can be found by summing the far-field pattern re-radiated by each of the individual elements in the array. The reflectarray also has a very small physical footprint, can be conformal, allows fabrication using traditional lithography techniques, and grants the possibility of utility stacking [38]. As a passive device, the reflectarray inherits many of the advantageous characteristics of the reflector antenna including simplified design using ray tracing and relatively low-loss operation. The combination of these beneficial characteristics makes reflectarrays desirable for use in a multitude of antenna systems [39].



Figure 6: Planar reflectarray mimicking the reflected wavefront of a parabolic reflector.

From a conceptual standpoint, the individual microstrip elements making up the reflectarray can be viewed as direct reflector/phased-array hybrid. When radiation impinges on an element in the reflectarray the phase delay introduced upon reflection relative to the neighboring elements will depend on the spacing difference between each element (path difference) and the dimensional differences between the elements (equivalent to a feed phase difference). Most reflectarray designs have fixed element periodicity for simplicity, and instead simply varied the dimensions of the array element to change the surface impedance at the point of the element to introduce the desired phase shift in the same way that the phased array would vary waveguide lengths or the reflector would vary curvature to create the desired phase delay.

The first reflectarray design published in 1963 was not a microstrip array, but instead relied on an externally illuminated series of stacked rectangular waveguides to impress reflected phase responses [40]. Nearly a decade later, in the late 1970s, initial work began

to adapt reflectarray concepts to exploit the emerging field of microstrip technology giving birth to the modern reflectarray [41]. The majority of these early microstrip reflectarrays utilized printed dipole or crossed dipole array layouts for phase modification due to their relative ease in characterization and fabrication. Around the 1990s, more complicated and efficient element geometries emerged with the creation of ring [42], variable patch [43], and stub-tuned reflectarrays [44]. State of the art reflectarray research has begun to focus on reflectarray bandwidth improvement, polarization control, and offset feed configurations.

Several commercially available reflectarrays have already been released to the market. Malibu Research, based in Camarillo, California, currently sells a reflectarray known as FLAPS<sup>TM</sup> (Flat Parabolic Surface) for use in rf and millimeter radar applications [45] -[46]. The FLAPS<sup>TM</sup> is an integral component of many new microwave devices including the newly deployed crowd control rf gun. Because the FLAPS<sup>TM</sup> is designed to be foldable, it allows the rf gun to be rapidly deployed on military or civilian vehicles, unlike earlier designs that utilized bulky reflectors. ILC Dover based in Frederica, Delaware markets an inflatable reflectarray as a way to lower payload weight when deploying in Ka and X band satellite applications [47]. Additionally, TRLabs in Canada offers a tunable reflectarray design for beamsteering and offset feeding applications [48]. No prior research has been carried out to expand reflectarray operation into the infrared.

#### **<u>1.3 Thesis</u>**

Even with the advent of high-resolution fabrication and improvements in material characterization at high frequencies, very little research has been carried out into using so-called metamaterial surfaces for phase shaping in the infrared, aside from polarization control and detector beam steering. Compared to traditional optical elements, resonant-antenna-based wavefront shaping surfaces can be cheaper to fabricate, have a smaller physical footprint, and allow for direct stacking of multiple planar elements, (e.g. filters and polarizers) for additional weight and volume reductions. These IR resonant focusing element devices could also provide additional degrees of freedom not previously available in conventional polished and diffractive IR-optical surfaces for correction of monochromatic and chromatic aberrations.

It is the intended goal of this dissertation to establish practical methods for development and characterization of passive, phased elements in the infrared. Procedures for designing, analyzing, and modeling planar resonant elements in the infrared are described. Fabrication of these devices using electron beam lithography is outlined with an emphasis on reducing future fabrication costs. Finally, the dissertation provides detailed instruction for testing final and intermediate versions of these devices using interferometry and at-focus imaging. In all, three classes of resonant, phased surfaces are considered: reflectarrays, transmitarrays, and emitarrays.

# **<u>1.4 Prior Publication and Financial Support Disclosure</u>**

Portions of this dissertation were originally published in [49]. Portions of Chapter 3 were originally published in [50] and [51]. Portions of Chapter 4 were originally published in [52] – [56]. This work was supported in part by the Lockheed Martin Corporation and in part by the Florida High Tech Corridor Council.

# **CHAPTER 2: THE INFRARED RESONANT PHASED ELEMENT**

#### 2.1 Metal-Dielectric Resonance Modes

The performance and behavior of planar, resonant antenna elements have been described in detail in the literature (i.e. [6], [8], [15], etc.). Simply speaking, if a metal film deposited on the surface of a dielectric is illuminated by an incident electromagnetic wave, currents will begin to be excited on the surface of the metal film. If the length of the film, along the direction of the polarization of the incident radiation, were to be less than a wavelength, a standing-wave current mode can be formed on the film (Figure 7). The currents excited on this film will also interact with currents on other neighboring elements because of fringing fields excited at the edge of the film, resulting in electromagnetic coupling (Figure 8). The resulting standing wave gives rise to a reactive storage mechanism, with the peak storage occurring when the standing wave has current nulls at the two ends of the metal film and a maximum at the center. At this wavelength, the antenna is considered to have its primary resonance. The overall spectral performance of these surfaces can therefore be related to that of a semiconductor band gap, where the center of the band gap of the antenna element is at the point of peak energy storage. The metal surface may also have multiple resonant wavelengths, depending on its geometry, which will result in multiple absorption bands.



Figure 7: Current modes on two microstrip antenna elements in an infinite array at resonance (standing wave present).

Incident radiation is polarized from the right to the left of the page.



Figure 8: The near-field electric field pattern from the two microstrip antenna elements in Figure 7 demonstrating coupling. Incident radiation is polarized from the right to the left of the page.

Several factors define the shape of an antenna's primary band gap. The geometry of the metal film will determine the current path that must be taken to establish resonance and will determine the antenna's center wavelength and bandwidth. The index of the dielectric material the antenna is fabricated on will influence the effective wavelength of the incident radiation on the top surface of the metal film. Inter-element spacing will determine the degree of coupling observed between elements. Inclusion of a groundplane will allow for the excitation of microstrip-type modes and suppression of the backside re-

radiated lobe. Because of the planar nature of these surfaces, microstrip type antennas are also strongly anisotropic.

# **2.2 Relating Resonance Properties to Phase**

For any electromagnetic resonator, it is possible to directly relate the surface's spectral reflectivity to its phase delay upon reflection. From Cauchy's integral theorem and the residue theorem, the reflectivity of a surface (R) can be directly related to the phase delay of that surface ( $\theta$ ) by the Kramers-Kroenig relation [57]:

$$\theta(\omega) = \frac{\omega}{\pi} \int_{0}^{\infty} \frac{\ln(\frac{R(\omega')}{R(\omega)})}{\omega'^{2} - \omega^{2}} d\omega'$$
(12)

where  $\omega$  in this case is the desired angular frequency. Several important properties can be determined from this relationship. First, for phase shaping to occur, a change in reflectivity must occur. Thus, this requires all reflective phase shaping surfaces to exhibit some loss to operate. Second, the phase delay introduced by a surface is determined by the reflectivity of the surface at all frequencies, not just at the frequency of interest for phase shaping. This allows for multi-band metamaterials, such as slot-loaded elements, to shape phase differently than a single band resonator, even in the case where the one of the additional resonance bands falls outside the design frequency. Finally, a band gap type of resonance will result in a relatively symmetrical dip in reflectivity with a specified bandwidth. This symmetrical drop in resonance will result in a non-linear decrease in phase upon reflection with a smaller phase shift at shorter wavelengths and a larger phase shift at longer wavelengths. This is commonly referred to by the saying "bumps (in reflectivity) makes wiggles (in phase)." A similar type of relationship, from

reciprocity, will later be demonstrated for transmission devices, and it can be seen that for a single layer, the frontside and backside lobes of a resonating antenna element will be in phase.

# **2.3 Chromatic Dependence of Resonance Modes**

The chromatic dependence of the resonance mode for a resonating antenna element array can be found directly from the Kramers-Kroenig relation. Figure 9 demonstrates the band gap behavior of an arbitrary resonant metamaterial element with a groundplane. The resonance behavior is relatively symmetrical with the center frequency around 25 THz. The phase upon reflection for this surface is presented in Figure 10. As mentioned before, the phase decreases with an increase in resonance frequency or shorter wavelength. The specific phase and reflection magnitude depends on the resonant structure used and cannot be easily predicted analytically. Modeling of these structures will be discussed later in this chapter.



Figure 9: Band gap behavior of an arbitrary resonant, groundplane-backed antenna array.



Figure 10: Phase behavior of an arbitrary resonant, groundplane-backed antenna array.

While it is not possible to look at a metamaterial element and immediately predict its performance without modeling, several important rules of thumb can be utilized. For a groundplane-backed element, changing the width, length (along the direction of the incident radiation polarization), or the dielectric material isolating the antenna structure

from the groundplane will primarily result in a change in the center frequency of the element. Altering the thickness of the dielectric standoff layer or altering the trace material making up the antenna element will primarily change the bandwidth and depth of notch of the resonant antenna element. These rules are summarized in Figure 11 and all of these rules will be explored in greater detail throughout the dissertation, especially in Chapter 3.



Figure 11: Summary of design rule of thumbs for resonant, groundplane-backed antenna array.

# 2.4 Angular Dependence of Resonance Modes

The angular dependence of a resonant antenna element in an array is not quite as simple to explain as its chromatic behavior. For most of the dissertation, reflected phase and magnitude will assumed to be measured or modeled at normal incidence, unless otherwise specified. The angular performance of the same metamaterial from the previous section is presented in Figure 12 and Figure 13. Up to about 22.5 degrees (45degree cone angle) the antennas performance is relatively unchanged relative to its performance at normal incidence. At larger angles, the two polarization states (TE and TM) no longer behave similarly and significant changes in phase relative to angle are observed. While difficult to predict without the use of modeling, the decrease in phase delay can be explained by the effective size of the antenna elements increasing with increasing angle of incidence. The separation of the two polarization states is due the to TM wave no longer being polarized along the direction of the resonant element and, thus, experiencing difficulty exciting a cavity mode and suffering interference from excited surface currents.



Figure 12: Reflectivity vs. incident angle for an example groundplane-backed antenna at 28.28 THz. In the plot, the TE mode is solid and TM mode is dotted.



Figure 13: Phase vs. incident angle for an example groundplane-backed antenna at 28.28 THz. In the plot, the TE mode is solid and TM mode is dotted.

# **2.5 Computational Electromagnetic Modeling of Infrared Phased Elements**

For the highest design accuracy before fabrication, most phased element designers employ some form of computational electromagnetic modeling (CEM) [58]. CEM takes into account system non-idealities, such as lossy materials or surface coupling, which are difficult to incorporate into circuit or transmission-line equivalent models without a significant increase in complexity. In the context of the dissertation, two independent modeling approaches have been considered and consulted: the infinite array FEM model and the periodic MoM model.

The finite element method (FEM) is a numerical technique by which a three dimensional model, representing the designed antenna, is discretized into a sub-domain in which the fields are represented by local interpolation functions. Matrix equations are derived from

a global assembly of the finite elements in the mesh while enforcing boundary conditions. Solving the matrix equations yields local fields at nodes throughout the mesh, which may be interpolated to arbitrary locations. Post processing computations produce far-field information and scattering parameters. Ansoft HFSS, employed in the dissertation for FEM modeling, specifically generates tetrahedral meshing elements and utilizes a threedimensional design interface for model boundary definition [59]. Ansoft HFSS also includes numerous features beneficial to modeling phased devices, such as automated plotting, wave ports for quasi-plane wave excitation, and parametric solution sweeps. HFSS is also desirable for modeling in the infrared because the software includes the capability of incorporating dispersive materials.

While FEM is suitable for three-dimensional geometries, the Method of Moments (MoM) numerical technique lends itself to planar structures by, traditionally, only meshing the surface of the radiating structures. The reduced discretization required for MoM leads to shorter solution times when compared to the same system modeled using FEM. Considering the structures developed in this dissertation are entirely planar, the shorter simulation times and spatial simplifications make MoM solvers desirable. By meshing the trace surfaces of a design using a given number of polygons for a set frequency, the method of moments technique is used to solve the mixed-potential integral equation (MPIE) and calculate the surface current everywhere on the mesh. Ansoft Designer, the MoM solver utilized in the dissertation, uses a zero-order normal element basis functions are applied to the MPIE to obtain a matrix equation, which is then solved to find surface

current. From the surface current, Designer calculates the S-parameters and the radiated fields. Much like Ansoft HFSS, Designer also includes support for finding far fields, plane wave excitation, and parametric solving.

### 2.5.1 Finite Element Model

Of the two models, the FEM model will always have the potential to be the most accurate, although there is the risk of over-simplification when selecting excitation and bounding methods [61]. For all of the devices modeled in the dissertation, the infinite array waveguide method (IAWM) was identified the best tradeoff between accuracy and solution time. The IAWM utilizes an ideal waveguide geometry to approximate a plane wave excitation and enforces the symmetrical mutual coupling that would be expected with an infinite planar array [62]. Assuming a waveguide with a square cross-section, two opposing faces along the cross-section of the rectangular waveguide are assigned an ideal perfect magnetic conductor (PMC) surface and the other surface is allowed to be the default HFSS boundary, which is a perfect electric conductor (PEC). The primary mode of this waveguide is TE<sub>0</sub>, which mimics a quasi-TEM wave (Figure 14). Placing the element to be tested in this waveguide and exciting or bounding the two open apertures of the waveguide allows for phase characterization of the element using the scattering matrix and preserves symmetric mutual coupling through image theory.



Figure 14: IAWM waveguide cross section with electric field polarization.

Creation of a transmissive FEM model (Figure 15) initially follows the standard procedure used in the formulation of any planar, passive HFSS antenna model. The foundation of the model is a three-dimensional rectangle representing the substrate with dielectric properties equal to the as-fabricated values. Computationally, it is not possible to model the full extent of the substrate. Thus, the substrate layer should be approximated to be a quarter-wavelength in thickness at the longest solved wavelength to ensure no unwanted coupling between the resonant structure and the feeding structure on the bottom face of the substrate at that wavelength. The quarter-wavelength height will also ensure accurate results at shorter wavelengths because the electrical distance between the radiating structure and the waveport increases inversely with wavelength. On the top of the substrate, the geometry that represents the resonant element is constructed. It is critical that this element have a finite thickness, accurate material properties, and the ability to solve for fields within the structure. A three-dimensional air box wraps around the element to bound the air side of the model space with the bottom face of the air box

touching the top face of the substrate layer. PEC and PMC boundaries are assigned on the sides of the air box and substrate, as defined in Figure 14. Finally, the top of the air box and the bottom of the substrate are terminated with waveports having a wave impedance matched to the surface in contact. Solving the model occurs by sweeping the frequency or the dimensions of the resonant element and calculating the change in phase of the scattering matrix (typically  $S_{11}$  for reflection and  $S_{21}$  for transmission).



Figure 15: Model layout for FEM transmissive design.

Some additional post processing may be required for the transmissive design. Because the substrate is terminated in a perfectly matched waveport, it may be necessary to introduce additional loss in the calculated magnitude of transmission to account for the backside reflection from the other side of the substrate. Similarly, additional steps must be taken to account for substrate Fabry-Perot resonances, if present. For frequency swept designs, de-embedding will be required to normalize the linear offset in phase associated with the effective change in path length due to the change in wavelength. It should be noted that this simulation is only valid for on-axis illumination.

Reflective modeling in HFSS is fundamentally similar to transmissive modeling. The layout of the unit cell is identical except the lower waveport has been replaced with a finite conductivity boundary to simulate the groundplane and the substrate is equal to the thickness of the desired standoff layer (Figure 16). Determination of the relative phase delay upon reflection can be found by calculating the phase of  $S_{11}$  at the top waveport. No additional post-processing is required for the substrate since it is accurately modeled as finite. De-embedding is still necessary to account for the change in wavelength at different frequencies.



Figure 16: Model layout for FEM reflective design. The groundplane in the actual model does not need to have a finite thickness, but a finite thickness has been added to the image for clarity.

The modeling approaches presented in this section are by no means complete or the only way to model phased structures in HFSS. Another method to model the structures that was used in the early development of the dissertation is to employ an actual plane wave excitation. Starting with the previous model, the top boundary of the air box is replaced with a perfectly matched layer (PML) boundary "referenced to FSS" and the model is typically excited by an external uniform plane wave instead of a waveport. Instead of PEC and PMC boundaries, the sidewalls of the model will have Master/Slave boundaries. Master/Slave boundaries are linked boundaries where field patterns exiting one boundary are forced to match patterns entering on the other boundary. This field enforcement makes it possible to account for non-normal excitation, while still maintaining a specific incident wavefront. Determination of phasing can still be found from the far-field scattering matrix. Version 11 of HFSS also provides the option for using Floquet Ports, which are similar to waveports in terms of setup, but provide the ability for off-axis simulation.

# 2.5.2 Periodic MoM Model

While less accurate than HFSS due to the lack of meshing in the entire solution space, the solve time improvements from Designer were found to be valuable and the results from these models were accurate for groundplane backed designs. Due to the relative thickness of the substrates used in infrared fabrication, Ansoft Designer is not suitable for modeling transmissive designs. In the model, there is no need to specify the three dimensional properties of the phased device, but a separate layer stack-up dialog is utilized (Figure 17). Through the layer stack-up, Designer defines thickness and material properties for

the standoff layer, groundplane, and trace layers. A separate interface is used to draw the actual resonant element (Figure 18). Placing structures on a "signal layer" effectively deposits metal on that layer. Placing structures on a "metalized signal layer" etches metal from the layer, which is assumed to initially be of infinite extent. To take into account coupling between neighboring elements, the edges of the model have periodic boundaries defined with size equal to the periodicity of the array. Excitation is a plane wave at prescribed angle of incidence. Far-field phase is found by varying element dimensions or frequency and calculating the reflected phase change. No additional post processing is typically needed.

	Name	Туре	Material	Drag Mode	Thickness	Lower Elevation	Upper Elevation	Roughnes
-	Trace	signal	gold	bottom ali	90nm	1.2um	1.29um	Oum
	Standoff La	dielectric	Zr02	1	1.2um	Oum	1.2um	
	Ground Pla	metalizedsig	pec	middle ali	Oum	Oum	Oum	Oum

Figure 17: Example of layer stack-up in Ansoft Designer.



Figure 18: Model layout for periodic MoM modeling.

# 2.6 Fabrication of Infrared Phased Elements

Fabrication of all of the nano-scale devices developed in the dissertation was achieved following a standard electron beam (e-beam) lift-off process. Substrates used include standard silicon wafers or silicon dioxide optical flats. Before fabrication of the patterned surface, groundplane and standoff layers were deposited on the substrate (Figure 19). Groundplanes, if necessary to the design, were either a thin film of gold (Au) or aluminum (Al). Both films were formed using physical vapor deposition (PVD). Gold films were deposited using e-beam evaporation and necessitated an adhesion layer of 5 to 10 nm of titanium (Ti), which was also evaporated using e-beam evaporation. Aluminum films were deposited using thermal evaporation and did not require an adhesion layer. The standoff layer, if necessary to the design, consisted of either zirconium dioxide (ZrO<sub>2</sub>) or Dow Corporation's Cyclotene 3022-35. The ZrO<sub>2</sub> films were deposited using ion assisted e-beam evaporation by an external company, Evaporated Coatings, Inc. in

Willow Grove, PA. Cyclotene was deposited at the IR Systems Lab and is based on Bstaged bisbenzocyclobutene (BCB), a spin on polymer. Processing of BCB films, as well as spin speeds, are outlined in [63].



Figure 19: Standard substrate coating process.

Pattern writing of the desired surface followed a standardized e-beam fabrication procedure developed by Charles Middleton, a member of the IR Systems Laboratory at the University of Central Florida. The process described here is for a single layer of resist and a more complicated, bi-layer procedure that was also used for earlier devices is outlined in [49]. E-beam lithography utilizes a vector scanning electron beam to write the desired device pattern into an e-beam sensitive resist made up of large chains of polymers deposited on the device's standoff layer. In the case of a positive resist, when exposed to the electron beam, the polymers in the resist break apart and the exposed region can be removed from the surface of the standoff layer using a chemical developer. The size and sharpness of the exposed pattern is controlled by the e-beam's electron beam current, dose, and size, which required the use of dose matrixes for characterization prior to fabrication. E-beam lithography is described in detail in [64]. With the desired regions of the standoff layer exposed, it is possible to deposit the metal making up the pattern using a conventional PVD process. Because the evaporation process deposits metal uniformly across the wafer, a final lift off step is necessary to remove undesirable metal deposited on the resist and the remaining resist itself.

Before pattern writing can begin, the substrate, including any coatings, must be cleaned to remove any organic debris or large particles on the surface that may lead to unwanted contamination. The substrate was spun at 6000 RPM for one minute and was sprayed for 10 seconds each with acetone, methanol, and isopropyl alcohol (IPA), in that order. The substrate was then placed on a hotplate at 180 °C for dehydration baking. After three minutes, the substrate was removed and blown for ten seconds using nitrogen to remove any particles that may have accumulated on the surface of the wafer during baking. ZEP 520A-7, a positive resist, was spun on the surface of the substrate at 3000 RPM for 80 seconds and baked for four minutes at 180 °C for a layer thickness of about 300 nm. The resist-coated wafer was then loaded into the UCF/CREOL Leica EBPG5000+ Electron Beam System, the e-beam system used for pattern writing of the all of the designs, for vacuum pump down. Beam current for all designs was 25 nA, accelerating voltage was 50 kV, and dose was 100  $\mu$ C/cm<sup>2</sup>. Pattern writing for the devices typically lasted for one to four hours, upon which the exposed wafer was removed from chamber, ready for development. The resist coating and exposure steps are outlined in Figure 20. The ZEP layer was developed and the exposed sections removed by bathing the wafer in ZEP RD developer for 90 seconds. Development was then stopped by an IPA rinse and the wafer was dried using nitrogen (Figure 21).



Spun on Resist

E-beam Exposure

Figure 20: Resist coating and e-beam exposure.



ZEP Development Figure 21: Resist development to reveal pattern.

With the desired pattern now present in the resist, the developed wafer was then loaded into an evaporator. Similar to deposition process used in forming the groundplane, a thin-film of metal (either titanium, nickel (Ni), gold, or aluminum) was deposited to form the resonant elements. A final lift-off process was employed to remove the unwanted deposited metal and un-developed resist. The metal layer on top of resist was first removed by gently rolling scotch tape across the entire wafer. Most metals will simply roll-off with the tape due to poor adhesion between the film and the resist layer. The ZEP layer was then lifted-off in a methylene chloride or ZEP Remover (ZDMAC) ultrasound bath and, subsequently, residual solvent was removed with an IPA rinse. For further cleaning and removal of unwanted resist with designs not containing a BCB standoff

layer, a 4 minute oxygen plasma etch using Branson Barrel Etcher P2000 was employed. At this point, only the desired pattern remained and the device was ready for testing.



Figure 22: Metal evaporation and lift-off process.

# **CHAPTER 3: METAMATERIAL RESONANCE DAMPING**

#### 3.1 Infrared Metamaterial Dependence on Dispersive Materials

One of the greatest limiting factors of IR metamaterial modeling has been the assumption that materials at IR exhibit electromagnetic properties independent of frequency. Traditionally a valid assumption at rf, the majority of materials utilized in metamaterial fabrication exhibit measurable frequency dependent (FD) optical properties at IR. FD material properties in the infrared are due to several physical phenomenon including molecular vibration, phonon absorption, free carrier loss, and defect scattering [65]. This material dispersion can have a significant impact on the measured performance of the fabricated metamaterial and will degrade agreement between measured and modeled results when assuming static material properties. Furthermore, a large number of commercial electromagnetic solvers used in metamaterial characterization were developed specifically for rf application and, thus, require frequency independent material definitions exclusively or provide only a limited means to account for dispersive materials.

To overcome this limitation, a procedure to account for FD material properties in IR metamaterial modeling was developed. This procedure includes material measurement using an ellipsometer and integration of dispersive materials in existing commercially available full-wave modeling packages. Two example FSS designs are presented showing significant improvements in agreement between modeled and measured results.

## 3.1.1 Material Characterization

Before modeling an IR metamaterial design, materials used for fabrication must be first characterized for their dispersive optical properties. While FD properties for many materials have been previously characterized and published, inconsistency in measurement approaches limit the utility of such results. Published material studies frequently characterize materials only in ideal situations, often at a single frequency, such as within a vacuum, as a bulk composition [66], or using mathematical models [67]. In addition, even if the material is studied in a similar configuration as the FSS to be modeled, variability in deposition techniques, layer intermixing, atmospheric conditions, material composition, and handling can render prior measured data inaccurate for modeling. For the highest possible accuracy when modeling FSS on dispersive materials, optical material properties must be characterized directly as deposited. Specifically, a J.A. Woollam Infrared Variable-Angle Spectroscopic Ellipsometer (IR-VASE) was utilized in the course of the dissertation to measure the optical properties of each material used in fabrication (see Appendix A.1 for more information). Because most commercially available modeling software programs only accept material property definitions as complex dielectric constants, and not index of refraction, the measured data from the ellipsometer had to be converted for direct utilization by using the relationships:

$$\widetilde{n}(\lambda) = \sqrt{\widetilde{\varepsilon}_r(\lambda)} \tag{13}$$

or, if the software only accepts conductivity for metal layers,:

$$\sigma = 4\pi f \varepsilon_0 nk \tag{14}$$

#### **3.1.2 Implementation of Dispersion in Full-Wave Simulation**

To carry out modeling, a MATLAB function was created to utilize the measured FD material properties. The MATLAB function consists of three major components – User Interface (UI), Solver Independent Code (SIC), and Solver Specific Code (SSC). The UI component of the code provides the interface necessary for user input and real time presentation of results. The SIC component interprets the users input, reads FD material properties from the shared network library, and creates result files and directories. The SSC component provides functionality to interface with a specific external electromagnetic solver and to interpret the results generated by the solver. The function's layered approach is desirable as it allows for easy integration of multiple electromagnetic solvers without changing the UI or SIC. Currently, Ohio State University's Periodic Method of Moments (PMM) and Ansoft Designer, both Method of Moments solvers, are supported. It should be noted that, as mentioned previously, Ansoft HFSS already includes support for FD materials.

Solutions for frequency dependent material designs are realized using frequency pointby-point simulation. To improve performance, the MATLAB function is provided with a template specifying initial geometry. Step modeling is achieved by populating the desired template with material properties at each frequency step and calling the necessary solver. In the function's current implementation, PMM setup files, written in FORTRAN, are directly modified at each step, whereas Designer setup files utilize a closed file format and require modification using VBScript to directly interface with the modeling program. Results are then stored for each frequency step in a spreadsheet and the UI is updated in real time. A summary of the program is provided in Figure 23.



Figure 23: Implementation of frequency dependent modeling.

In addition to support for FD materials, the developed MATLAB function further enhances all of the solvers by introducing new capabilities. Most significant of this new functionality, especially from the standpoint of the user, is the fact that parameter input, user interfaces, and results are all presented identically regardless of the chosen solver. Neutral presentation is desirable to lower the learning curve necessary to model, such as the need to learn FORTRAN for PMM or the Ansoft product UI for Designer, and improves post processing and sharing of data between solvers. The function also adds the ability to specify variable parametric sweeps and auto-renders the design in 3-D functionality not available in some commercially available solvers, including PMM. The function also easily integrates optimization.

## 3.1.3 Example Results: Square Loop FSS on ZrO<sub>2</sub>

For verification of the need to account for FD material properties in FSS modeling, a manganese (Mn) square loop FSS on a ZrO<sub>2</sub> standoff layer with an Au groundplane (Figure 24) was fabricated and tested using a spectral radiometer (Figure 25). The same design was modeled using PMM assuming frequency independent materials ( $\varepsilon_r = 3.0272$ ,  $tan(\delta) = 0.023$ ,  $R_s = 40\Omega$ ) and the developed MATLAB function following the process outlined in the previous section. Figure 26 is a plot of the modeled and measured emissivity of the square loop FSS. From the figure, the FD model provides an improved indication of the device's measured behavior over the frequency independent model including a better bandwidth match from 3-6 µm, accurate prediction of the device's reflectivity peak around 7 µm, and improved agreement from 8 to 14 µm. Ansoft Designer yielded similar results.



Figure 24: SEM image of fabricated square loop FSS on ZrO<sub>2</sub>.



Figure 25: Infrared spectral radiometer.



Figure 26: Measured, frequency independent PMM, and frequency dependent PMM results for square loop FSS on ZrO<sub>2</sub>.

In addition to modeling results, run time data for the model from Figure 26 was collected for each program and summarized in Table 1. As expected, the use of frequency dependent materials facilitated by the MATLAB function resulted in an overall increase of run time. The increase can largely be attributed to additional time required to copy the measured permittivity values from the shared drive, extract the results, save the results to a spreadsheet file, generate of the function's UI, and launch and close the desired solver. Overall, the longer runtime is acceptable due to the increase in model accuracy and additional program functionality.

Function of full time for a square roop fob using for frequency points.Frequency IndependentFrequency DependentPMM177s555sDesigner552s1291s

Table 1: Comparison of runtime for a square loop FSS using 100 frequency points.

## 3.1.4 Example Results: Square Loop FSS on a Polymer

From the standpoint of mass-producing an IR metamaterial, non-traditional standoff layers, such as polymers, would be desirable to lower fabrication costs, to reduce fabrication time, and to allow for flexible substrates. Due to their composition, most polymers will exhibit significant frequency dependence and numerous absorption bands at infrared. To evaluate FSS behavior on a polymer dielectric, another square loop FSS was modeled (Figure 27) using both a fixed, lossless permittivity dielectric ( $\varepsilon_r = 1.5$ , tan $\delta = 0$ ) and the complex indices of a sample plastic measured from the IR-VASE (Figure 28). When assuming a fixed permittivity dielectric, the square loop FSS was easily optimized for high emissivity from 5 – 8 µm simply by scaling an existing design. Running the same models using the developed MATLAB function and accounting for the frequency dependence of the plastic, the FSS retains some of its original behavior with the introduction of a high emissivity band between 8 – 9 µm and a sharp dip in emissivity around 7.5 µm. From a design standpoint, this new behavior can significantly change the

potential applications of the FSS by effectively expanding the device's emissivity band and introducing an undesired dip in the middle of that band. Even with the measured optical properties, predicting these new trends before testing is clearly problematic when using only a frequency independent model. By including material frequency dependence, further design optimization can occur with a reasonable expectation of accuracy and, thus, a reduction in the need of costly fabrication and measurement iterations.



Figure 27: Frequency independent PMM and frequency dependent PMM results for square loop on plastic.



Figure 28: Frequency dependent index of refraction (n,k) of polymer dielectric from ellipsometer.

# 3.1.5 IR Material Dispersion: Conclusions and Applications

As demonstrated in the previous sections, material dispersion can have a significant impact on the performance of an IR metamaterial. In the context of the dielectric standoff layer, any absorption bands in the material may lead to a loss of the desired electromagnetic behavior of the metamaterial in that spectral range. Outside of that spectral range, electromagnetic performance can also be altered from the gradual change in the real part of the index, known as normal dispersion. These effects are well known in the literature, even in the rf portion of the spectrum, but the severity of these phenomena at high frequencies demonstrate the need for accurate measurement and modeling in the infrared.
A metamaterial's dependence on the optical properties of the material making up the surface can also be exploited for design flexibility. The rest of the chapter will focus developing a method known as metamaterial damping – a technique that can be used to alter the bandwidth and resonance frequency of an antenna element operating in the infrared. This approach employs the dispersion of the metal making up the metamaterial device, as opposed to the standoff layer, and the inherent unique properties of most noble metals in the infrared to tune the performance of the surface.

## **3.2 Theoretical Foundation of Metamaterial Damping Theory**

One approach to achieve metamaterial optimization or spectral tuning in the infrared without altering the element geometry is to exploit the structure's resonance-behavior sensitivity to its metal-film conductivity. Variation in metal conductivity alters the Ohmic loss in the metamaterial, which yields resonance damping and two phenomena that are important for design optimization: bandwidth expansion and spectral shifting. Resonance damping is a well-understood concept within the physics and engineering community. If a harmonic oscillator encounters a resistive force, the system's oscillations will become damped, occurring at a lower frequency and magnitude compared to the undamped case. Consequently, the decrease in the magnitude of the oscillation must result in an increased resonance bandwidth for the system. For resonant antenna devices, the resistive force typically takes the form of attenuation due to the conversion of electromagnetic energy to heat. Passive, infrared resonant elements are especially sensitive to damping, due to the highly dispersive and lossy nature of metals in the infrared band.

The effect of damping due to the finite conductivity in infrared metamaterials has been published previously, but not heavily explored as a means of metamaterial optimization or tuning. Munk acknowledged that altering the conductivity of an FSS resulted in frequency shifting, attributing this phenomenon to an increase of the real part of the permittivity of the metal film with frequency [66]. Novotny provides a much more detailed analysis of this behavior for visible antennas [68] and found results similar to the ones presented in the following sub-sections, without considering the related resonance-bandwidth issue. Frequency and bandwidth scaling due to damping can also be observed in the results of numerous other papers (such as [69] and [70]), but damping effects have largely been neglected or not properly identified.

#### **3.2.1 Thin-Film Resistor Impedance at Infrared Frequencies**

Before introducing an equivalent resonator model, it is necessary to first develop a method for determining the Ohmic loss of a metamaterial element in the infrared. For the purpose of this derivation, a simple metamaterial unit cell element contained within an infinite length sheet of the same material with thickness h is assumed to consist of a dipole slab of width w and length l. The dipole is excited by a plane wave normally incident on the dipole's surface with an electric-field component of the form:

$$\widetilde{E}(z) = \hat{x} E_0 e^{J \gamma z} \tag{15}$$

where  $\gamma$  is the propagation constant of the plane wave inside the metal film, *l* is orientated in the x-direction, *w* is orientated in the y-direction, and *h* is orientated in the z-direction. The current density excited by the wave as it penetrates the surface is:

$$\widetilde{J}(z) = \hat{x}\sigma E_0 e^{j\gamma z} \tag{16}$$

where  $\sigma$  is the bulk admittivity (dynamic complex conductivity) of the film. The propagation constant can be defined in terms of the optical properties of the metal film:

$$\gamma = \frac{2\pi f(n+jk)}{c} \tag{17}$$

where f is the frequency, n is the index of refraction, k is the extinction coefficient, and c is the speed of light in a vacuum. The relative permeability of the metal in the infrared is assumed to be unity. The total current, I, flowing through the metal film along the length of the dipole can be found by solving the integral:

$$I = w \int_{0}^{h} \widetilde{J}(z) dz \cdot \hat{x}$$
(18)

Neglecting the phase term of the propagation constant, Eq. (18) yields the expression:

$$I = \frac{cw\sigma E_0}{2\pi fk} \left(1 - e^{-\frac{2\pi fkh}{c}}\right)$$
(19)

Eq. (19) can be reduced by substituting in skin depth or:

$$\delta_{skin} = \frac{c}{2\pi f k} \tag{20}$$

To yield:

$$I = w\sigma\delta_{skin}E_0(1 - e^{\frac{-h}{\delta_{skin}}})$$
(21)

Similarly, the voltage, *V*, developed over the length of the slab can be found by:

$$V = E_0 l \tag{22}$$

Finally, the impedance of the film,  $Z_{film}$ , can be determined by Ohm's Law:

$$\frac{V}{I} = Z_{film} = \frac{l}{w\delta_{skin}\sigma(1 - e^{\frac{-h}{\delta_{skin}}})}$$
(23)

Eq. (23) can be recognized as the classical thin-film resistor impedance equation when the ratio of the film thickness to skin depth approaches infinity.

In the microwave portion of the spectrum, it is customary to assume that the admittivity of the metal film is purely real and nearly equal to the dc conductivity. This conclusion is from the Drude-Lorentz model, which relates dc conductivity and metal relaxation time to admittivity [65]:

$$\sigma(f) = \frac{\sigma_0}{\left(2\pi f\tau\right)^2 + 1} (1 + j2\pi f\tau) \tag{24}$$

where  $\tau$  is the metal's relaxation time and  $\sigma_0$  is the metal's bulk dc conductivity. The relaxation time for most noble metals will be on the order of 10 to 100 fs [65], which causes the imaginary term in Eq. (24) to approach zero at low frequencies. In the THz, the imaginary term can no longer be neglected, as the frequency is on the same order as the inverse of the relaxation time. Thus, it is of importance to account for complex admittivity in Eq. (23) by relating the conductivity directly to the film's optical properties, through:

$$\sigma = j2\pi f\varepsilon_0 (1 - (n + jk)^2) \tag{25}$$

where  $\varepsilon_0$  is the free-space permittivity. Substituting Eq. (25) into Eq. (23) yields:

$$Z_{film} = \frac{l}{2\pi f \varepsilon_0 w \delta_{skin} (1 - e^{\frac{-h}{\delta_{skin}}})(2nk + j(k^2 - n^2 + 1))}$$
(26)

Eq. (26) can also be re-written in terms of relative permittivity,  $\varepsilon_r$ , using the relationship:

$$\varepsilon_r' + j\varepsilon_r'' = n^2 - k^2 + j2nk \tag{27}$$

which yields:

$$Z_{film} = \frac{l(\varepsilon_r'' + j(\varepsilon_r' - 1))}{2\pi f \varepsilon_0 w \delta_{skin} (1 - e^{\frac{-h}{\delta_{skin}}})((\varepsilon_r'')^2 + (1 - \varepsilon_r')^2)}$$
(28)

Assuming that  $\varepsilon_r' \ll -1$ , which is typical for metals outside of the visible and ultraviolet portion of the spectrum [65], Eq. (28) can be simplified and separated into complex components:

$$R_{film} \approx \frac{l}{2\pi f \varepsilon_0 w \delta_{skin} (1 - e^{\frac{-h}{\delta_{skin}}})} (\frac{\varepsilon_r''}{((\varepsilon_r')^2 + (\varepsilon_r'')^2)})$$
(29)

$$X_{film} \approx \frac{l}{2\pi f \varepsilon_0 w \delta_{skin} (1 - e^{\frac{-h}{\delta_{skin}}})} (\frac{\varepsilon_r'}{((\varepsilon_r')^2 + (\varepsilon_r'')^2)})$$
(30)

where  $R_{film}$  is resistivity of the film and  $X_{film}$  is the reactivity of the film. From Eqs. (29) and (30), the ratio of the resistive and reactive components of the metal film is nearly equal to the material's loss tangent,  $tan(\delta)$ , or:

$$\frac{R_{film}}{X_{film}} \approx \tan(\delta) \equiv \frac{\varepsilon_r''}{\varepsilon_r'}$$
(31)

Finally, the expressions for  $R_{film}$  and  $X_{film}$  can be further simplified when they are expressed in terms of loss tangent and the real part of permittivity:

$$R_{film} \approx \frac{l}{2\pi f \varepsilon_0 w \delta_{skin} (1 - e^{\frac{-h}{\delta_{skin}}})} (\frac{tan(\delta)}{\varepsilon'_r (1 + tan(\delta)^2)})$$
(32)

$$X_{film} \approx \frac{R_{film}}{tan(\delta)}$$
(33)

The values of  $R_{film}$  and  $X_{film}$  represent the complex impedance of the metal slab presented to the incident radiation. Unlike the situation in the microwave band, the slab is no longer purely resistive. It will exhibit a largely capacitive response, because  $\varepsilon_r'$  is strongly negative in the infrared, with a loss tangent approaching unity.

#### **3.2.2 Equivalent-Circuit Resonator Derivation**

Using the dipole layout presented in the previous section, an equivalent circuit for an infinite array of this element at primary resonance is presented in Figure 29. The resonance properties of the equivalent circuit can be expressed by:

$$\Delta f = \frac{R}{2\pi L} \tag{34}$$

$$f = \frac{1}{2} \sqrt{\frac{1}{\pi^2 LC} - (\Delta f)^2}$$
(35)

where  $\Delta f$  is the full-width at half-power (FWHP) bandwidth of the dipole, R is the effective resistance of the dipole observed by a current excited at the surface of the dipole, L is the effective inductance of the slab, and C is effective capacitance. Inductance and capacitance are predominantly determined by the geometries of the element and the unit cell, with the resistance determined by the loss of the metal making up the dipole. The impedance of the dipole can be determined by summing in series the structural, or resonant, impedance of the element, assuming no damping or metal loss, and the impedance of the metal slab calculated previously. The resistance of the structural resistance,  $R_0$ , of the dipole:

$$R = R_{film} + R_0 \tag{36}$$

The structural resistance is caused by selective loading, substrate loss, or unwanted reradiation and is independent of the metal. Similarly, the inductance of the equivalent circuit can be related to the reactance of the metal film through:

$$L = L_0 - \frac{1}{2\pi f X_{film}} \tag{37}$$

where  $L_0$  is the structural inductance. Altered structural inductance from metal film reactance is likely the same thing reported by the plasmonic community as photon drag [71]. Capacitance is assumed to be independent of the metal film properties and is entirely determined by:

$$C = C_0 \tag{38}$$

where  $C_0$  is the structural capacitance due to mutual coupling.



Figure 29: Resonant circuit equivalent model for an infinite array of dipole metamaterial elements.

Mutual coupling is assumed to only occur along the long direction of the dipole.

Several important conclusions can be made regarding damping in the infrared using the proposed equivalent-circuit model. As expected, increasing the loss of the metal will result in an increase of the dipole's resonance bandwidth from Eq. (34) and this increase in bandwidth will result in an decrease in the resonance frequency from Eq. (35). Interestingly, increasing the magnitude of the reactance of the film will further increase the resonance frequency by decreasing the inductance of the slab from Eqs. (37) and (35). The reactance of the film has a limited impact on the bandwidth of the system relative to the resistance of the film because  $R_0$  will be significantly smaller than  $R_{film}$  for any good metamaterial structure, with most antenna loss occurring due to the metal making up the resonator. This leads to the unique situation where it is possible to have a narrowband design, with a small  $R_{film}$ , experience a significant decrease in the resonance frequency due to having a simultaneously large  $X_{film}$ . The specific conditions required for this to occur, in terms of metal properties, are discussed later.

Unfortunately, there are several limitations inherent to this theoretical approach. It has been shown that metals in the infrared, unlike the microwave portion of the spectrum, exhibit appreciable dispersion. The development of these equivalent-circuit models assumes quasi-static parameter values, which will lead to errors in the predicted bandwidths of the resonant metamaterial. The presented model also does not provide methods for accounting for some non-idealities present in the metal films, such as the anomalous skin effect and film depth profile variations. Most of these limitations can be mitigated by employing some form of full-wave electromagnetic modeling, which will be discussed in the next section. In general, optimization using metamaterial damping is not suitable for every application, since it involves increasing the absorption loss of the metamaterial element. Additional film absorption will lead to unwanted heat loading on the metamaterial element which may lead to structural or performance issues.

## 3.3 Simulated Validation of the Metamaterial Damping Theory

For validation of the proposed equivalent-circuit model, an infinite array of cross FSS elements was modeled at normal incidence using Ansoft HFSS, a finite-element-method electromagnetic solver. Cross elements were chosen due to their wide use in metamaterial structures, their similarity in performance to dipole elements, and their polarization insensitivity at normal incidence. The dimensions of the element considered are presented in Figure 30 and the element's square unit cell size was fixed at 2.5  $\mu$ m. The thickness of the metal film was fixed at 100 nm. The superstrate is taken as vacuum ( $\epsilon_r = 1$ ) and the substrate as lossless silicon ( $\epsilon_r = 11.4$ ). This design was chosen with typical dimensions for resonance near the LWIR (long-wave infrared) CO<sub>2</sub> laser line at 10.6  $\mu$ m or 28.28 THz.



Figure 30: Dimensions of cross metamaterial element.

#### **3.3.1 Real Frequency-Independent Conductivity**

The element under study was first modeled with four frequency-independent real values of conductivity typical for the infrared ( $\sigma = 0.5$ , 1, 5, 10 MS/m). In addition, the modeled performance of the design using a perfect electric conductor (PEC) material ( $\sigma = \infty$  S/m) was found to determine the undamped response. Unless the option to solve inside of a metal film is explicitly enabled, HFSS applies a finite-conductivity boundary to the metal film surface, where the thickness of the film is assumed to be infinite and the admittivity is assumed to be purely real for the calculation of surface impedance [59]. The finite-conductivity boundary surface condition is used to investigate the sensitivity of the crosses' bandwidth and resonance frequency to changes in conductivity without the influence of unwanted radiation propagation through the film or finite skin depth. Modeled results are presented in Figure 31. As expected from Eqs. (23), (32), and (33), the modeled cross metamaterial exhibits increased damping with decreasing conductivity. Because the conductivity is real, no reactive components are present to influence the damping behavior.



Figure 31: Modeled transmission of the cross element with four frequency independent conductivities.

# 3.3.2 Complex Frequency-Independent Impedance Surfaces

To study how the complex impedance of the film impacts resonance behavior and to validate the claims used to develop Eqs. (37) and (38), the cross design was modeled using four complex film impedances. HFSS has the capability to model surface boundaries with complex sheet resistance; however, the surface must have zero thickness, limiting the practicality of this approach. Sheet resistance can be related to the film impedance through

$$Z_{film-sheet} = \frac{w}{l} Z_{film}$$
(39)

The complex sheet resistance values used in modeling are presented in Table 2, as well as modeled center frequency and bandwidth. Each model is labeled with the format ( $R_{film}$ *sheet*,  $X_{film}$ -*sheet*) and the modeled results are presented in Figure 32. As expected, reducing the resistance of the film reduced the bandwidth and increased the resonance frequency. Similarly, increasing the reactance of the metal film resulted in a decrease in the resonance frequency. Slight variations in bandwidth are also observed with changes in reactance, but the bandwidth is significantly more sensitive to the resistance of the film, validating the claim made in the previous section.

resonance requency and sund vidan.						
Model Name	<b>R</b> <sub>film-sheet</sub>	X <sub>film-sheet</sub>	f (THz)	$\Delta f$ (THz)		
(1,1)	1 Ω/□	1 Ω/□	28.10	1.46		
(1,2)	1Ω/□	2 Ω/□	27.46	1.29		
(2,1)	$2 \Omega/\Box$	1Ω/□	27.95	2.82		
(2,2)	2 Ω/□	2 Ω/□	27.32	2.58		

 Table 2: Frequency-independent sheet resistances used in modeling and modeled resonance frequency and bandwidth.



Figure 32: Modeled transmission of the cross element with four frequency-independent complex sheet resistances.

## **3.3.3 Complex Frequency-Independent Conductivity**

It is also important to consider a more realistic model to test the validity of Eqs. (32) and (33). Thus, the cross from Figure 30 was re-modeled using representative frequency

independent complex dielectric parameters and HFSS was instructed to solve inside the metal films. Four individual models were compared to determine the crosses' change in response due to changes in magnitude of  $\varepsilon_{r'}$ ,  $tan(\delta)$ , and  $\delta_{skin}$ . The skin depth can be found directly from permittivity and loss tangent using the relationship:

$$\delta_{skin} = \frac{c}{\pi f \sqrt{2(|\varepsilon_r'| \sqrt{1 + tan(\delta)^2} - \varepsilon_r')}}$$
(40)

The permittivity values used and the derived parameters are listed in Table 3. Each model is labeled using a multiple of permittivity and loss tangent relative to those used in the first model with the format ( $\varepsilon_r'$ ,  $tan(\delta)$ ). The modeled results are presented in Figure 33.

 Table 3: Frequency independent permittivities used in modeling and derived loss tangent and skin depth values.

Model Name	٤ <sub>r</sub> ′	٤ <sub>r</sub> ″	tanð	$\delta_{skin}$ (28.28 THz, nm)
(1, 1)	-1800	2880	-1.6	33.12
(2, 0.5)	-3600	2880	-0.8	26.98
(1, 2)	-1800	5760	-3.2	25.76
(2, 1)	-3600	5760	-1.6	23.42



Figure 33: Modeled transmission of the cross element with four frequency-independent complex permittivities.

The results presented in Figure 33 are consistent with the equivalent-circuit model and Eqs. (29) and (30) – increasing the magnitude of the real part of permittivity decreases bandwidth and increasing the magnitude of the imaginary part of permittivity counteracts damping and shifts the resonance to higher frequencies. One of the more intriguing results is that model (2, 0.5) demonstrates a narrower bandwidth than model (2, 1), even though both models have the same real part of permittivity and model (2, 1) resonates at a higher frequency. This is caused by the loss tangent of the metal also playing a role in determining the resistance of the metal film. Minimizing  $tan(\delta)$  and maximizing the magnitude of  $\varepsilon_{r'}$ , from Eq. (32), will yield a smaller film resistance and, thus, a smaller bandwidth. Minimizing  $tan(\delta)$ ; however, also increases the reactive component of the metal film resulting in a larger resonance frequency. Conversely, (1, 2) has a narrower bandwidth than (1, 1) due to a larger loss tangent, which yields a lower film resistance,

but the smaller  $\varepsilon_r'$  results in a larger overall bandwidth when compared to (2, 0.5) or (2, 1).

# **3.3.4 Measured Metal-Film Properties**

To further test the proposed theory, the cross design was modeled using measured metal permittivities, which were complex and frequency dependent. Three metals were selected for comparison due to their distinct properties in the LWIR: gold for a loss tangent magnitude less than one, aluminum for a loss tangent magnitude greater than one and nickel for a loss tangent magnitude nearly equal to one. Determination of the optical properties of these metals in the infrared was done using films evaporated on a silicon witness sample and the IR-VASE ellipsometer. The measured and derived optical properties are presented in Figure 34, Figure 35, and Figure 36. The modeled resonance behaviors of the crosses using the measured optical properties of the metals are presented in Figure 37.



Figure 34: Measured real part of permittivity for gold, aluminum, and nickel.



Figure 35: Measured loss tangent for gold, aluminum, and nickel.



Figure 36: Calculated skin depth for gold, aluminum, and nickel.



Figure 37: Modeled transmission of four cross metamaterial arrays made of three different metals and PEC.

Once again, the results are consistent with the predictions from the equivalent-circuit model. Nickel demonstrates the most damped response since its magnitude of the real part of permittivity is small and its loss tangent is nearly equal to one. Gold has the largest magnitude of the real part of permittivity of the three metals and, thus, the lowest film loss and therefore exhibits the narrowest bandwidth and deepest notch. Consistent with results from the frequency-independent simulation, aluminum's large loss tangent magnitude decreases the film's reactive component and allows it to resonate at the highest frequency, closest to the undamped design. However, aluminum's small real part of permittivity results in a notch significantly shallower than gold, as well as a larger bandwidth.

## **3.4 Measured Validation of the Metamaterial Damping Theory**

To validate the modeled results, the cross-FSS design was fabricated and measured using the three metals. Fabrication followed the electron-beam lithography process outlined in the previous chapter. Deposition of the metal films was done using electron-beam evaporation for gold and nickel and thermal evaporation for aluminum. The substrate was a 350-µm-thick high-resistivity, double-side-polished silicon wafer. A SEM micrograph of a portion of the fabricated array is presented in Figure 38. Measurement of the transmission of these surfaces was carried out using a Perkin-Elmer micro-FTIR spectrometer with a microscope attachment. Results from this measurement, as well as the modeled results from Figure 37, are compared in Figure 39. Both modeling and profilometer measurements indicated that the thickness of the gold film was approximately 90 nm, while the other films were approximately 100 nm thick.



Figure 38: SEM micrograph of a portion of the gold cross metamaterial array.



Figure 39: Measured (solid) and modeled (dotted) transmission of three cross metamaterial arrays made up of three different metals.

Overall, excellent agreement is observed between modeled and measured results. Ringing is present in all of the measured results due to the Fabry-Perot resonance from the finite thickness of the substrate. This also leads to slight differences between modeled and measured notch depth since the HFSS model neglects the backside reflection. Additional differences can be attributed to the presence of native oxide layers for the nickel and aluminum films as well as the likely presence of residual resist.

# **3.5 Implications of Metamaterial Damping Theory**

Metamaterial damping has a significant impact on a wide variety of infrared resonant devices including antenna-coupled bolometers, waveguides, and FSS. More than any other device; however, phased resonant elements are especially sensitive to damping due to their narrowband performance. The next sections will outline some specific implications of damping theory in terms of material properties, bandwidth, and phase response.

## 3.5.1 Impact of Relaxation Time on Damping and Film Reactance

Based on the circuit model and measured results it is clear, with films sufficiently thick compared to the skin depth, that the loss tangent and the real part of permittivity are the best indicator of metamaterial resonance performance. From Eqs. (24), (25), and (27), loss tangent can be defined completely in terms of metal relaxation time, or:

$$\tan(\delta) \approx \frac{-1}{2\pi f \tau} \tag{41}$$

again assuming that  $|\varepsilon_r'| >> 1$ . Similarly, the real part of permittivity can also be defined in terms of loss tangent:

$$\varepsilon_r' \approx \frac{Ne^2}{m_0 \varepsilon_0 (2\pi f)^2} \left(\frac{-1}{1 + \tan(\delta)^2}\right) \tag{42}$$

With the usual definition of dc conductivity:

$$\sigma_0 = \frac{Ne^2\tau}{m_0} \tag{43}$$

where *N* is the carrier density, *e* is the electronic charge constant, and  $m_0$  is electron rest mass. Substituting Eqs. (41) and (42) into Eqs. (32), (33), and (40), yields:

$$X_{film} \approx \frac{-m_0 2\pi fl}{Ne^2 w \delta_{skin}}$$
(44)

$$R_{film} \approx -X_{film} (2\pi f\tau)^{-1}$$
(45)

$$\delta_{skin} \approx \frac{c\sqrt{1 + (\frac{1}{2\pi f\tau})^2}}{\sqrt{\frac{Ne^2}{2m_0\varepsilon_0}(1 + \sqrt{1 + (\frac{1}{2\pi f\tau})^2})}}$$
(46)

assuming an infinite film thickness for simplicity. From the previous results, to minimize bandwidth damping, it is desirable to maximize relaxation time, while reducing film resistivity. Conversely, to counteract frequency damping by decreasing the reactance of the film, it is desirable to minimize relaxation time, resulting in a larger skin depth. The only other option available to a metamaterial designer is to alter the carrier density which can be achieved by utilizing different metals and not necessarily through film processing. Of the two, metal film relaxation time provides the largest range of variation through the alteration of impurity scattering [72].

It should also be noted that this type of design optimization is limited to frequencies close to the relaxation frequency, or  $(2\pi\tau)^{-1}$ . For most metals, the relaxation frequency falls in the infrared and THz frequency bands. To illustrate, the complex impedance of two

fictitious bulk metals with relaxation times equal to 10 and 100 fs is plotted in Figure 40. Both metals have the same carrier concentration and their dc conductivity is set to  $10^{20}\tau$  S/ms. From the figure and earlier results, it is shown that it is not possible to counteract damping for these two metals in the microwave portion of the spectrum by reducing the reactivity of the metal film because the impedance is already nearly real. Above the relaxation frequency and into the visible, the real part of the impedance for both metals has approached its maximum value and the reactive component becomes dominant and nearly the same for all relaxation times. Thus, in both the microwave and visible, the metal with the highest dc conductivity, and typically the largest relaxation time, will have the narrowest bandwidth and highest resonance frequency.



Figure 40: The complex impedance of two bulk metals with relaxation time of 10 or 100 fs in the IR and THz. Carrier concentration for the two metals is assumed to be the same.

The region of interest for the proposed optimization approach falls in between the relaxation frequencies of the two metals. In this region, the metal with a higher real part

of impedance also has the smaller reactive component. Assuming an appropriate geometry is chosen to exploit the reduced reactivity of the film, according to Eq. (35) or similar expression, it is possible to have a metamaterial structure resonate at a higher frequency with a lossy metal than with a metal with having a larger dc conductivity. The metal film's higher resistivity will still yield a larger bandwidth than the metal with the larger dc conductivity, unless the carrier concentration is significantly different.

### 3.5.2 Bandwidth Considerations

In the development of the equivalent-circuit model, the bandwidth of the metamaterial was defined as the FWHP bandwidth about the resonance frequency. This definition is meaningful for devices where notch depth is significant, such as phased devices like reflectarrays or devices where a strong resonance is necessary to approximate an equivalent material response, such as negative permittivity in NIM metamaterials. For band-reject filter designs, like the cross metamaterial characterized in the previous section, the actual depth of the notch is not necessarily important, as long as the transmission of the reject band is below some arbitrary floor value. Thus, it has been customary in these situations to define the bandwidth as the frequency difference between two 3 dB roll-off points and not relative to the center resonance frequency. Assuming only one rejection band, the 3 dB roll-off points for a band reject filter will occur at the two points where the filter passes 50% of the maximum passband transmitted power.

To investigate the impact of damping on the 3 dB bandwidth, the two sets of bandwidths for the modeled results in Figure 37 were calculated and are presented in Table 4. As

expected from the equivalent-circuit model, the FWHP bandwidth increased as metal loss increased. On the other hand, the results for nickel demonstrate that damping can minimize the 3 dB roll-off bandwidth by reducing the contrast between the loss and pass bands. This result suggests that in infrared filter applications the narrowest bandwidth can be achieved through the use of lossy metals and not necessarily through high conductivity. The trade-off for this approach is higher loss in the passband and the requirement of smaller element dimensions to counteract frequency damping.

Table 4: Calculated bandwidths for the three measured-metal designs in Figure 37.

Metal	f(THz)	$\Delta f$ (THz)	<i>3-dB ∆f</i> (THz)
Gold	29.09	0.77	9.71
Aluminum	29.21	3.23	10.29
Nickel	27.16	3.77	9.50

# 3.5.3 Phase Considerations

The impact of metamaterial damping on phase can be explored through the Kramers-Kroenig relation outlined previously. From the relation, reducing the depth of the notch, and subsequently broadening the resonance, will result in a slower phase transition versus frequency, which can be beneficial to reduce fabrication tolerances. The downside of the slower phase transition is that the range of the possible phase variations will be decreased, a limitation that will be discussed in more detail in the next chapter. Low reactivity metals in the infrared, such as aluminum, are similarly desirable for phased elements due to their minimal frequency shifting compared to an ideal PEC element and their favorable large resonant element size at the designated resonant frequency.

# **CHAPTER 4: THE INFRARED REFLECTARRAY**

#### **4.1 Fundamental Theory**

The first device investigated for phasing in the infrared was the reflectarray. As discussed previously, the reflectarray is classically a planar, microstrip antenna array with a specific layout to give rise to a specific desired wavefront upon re-radiation. Phasing of the antenna elements can be achieved through a multitude of methods including slot-loading, stub-loading, variable unit-cell sizes, and multiple element geometries; however, the phasing mechanism used in the dissertation was largely focused on variable sized elements. This chapter will introduce the methods used in characterizing infrared reflectarray elements, the design techniques necessary to construct a focusing reflectarray, and a preliminary discussion of the aberrated image behavior to be expected from these surfaces.

# 4.1.1 The Variable-Patch Reflectarray Element

As mentioned in the Section 1.2.3, several reflectarray layouts have been designed, fabricated, and tested in the past. Based on this prior research, variable-patch reflectarray designs (Figure 41) have gained popularity as the optimal reflectarray element for efficient, wideband applications. Specifically, variable-patch reflectarrays achieve reflected phase variation by varying the length or width of the patch (or both equally to maintain polarization insensitivity). Changing the dimensions of the patch ideally alters the reactive component of surface impedance at the element's location on the array. Pozar

carried out the initial development of the variable-patch reflectarray [43] and several other individuals and organizations have tested and developed similar designs [73] - [75].



Figure 41: Variable-patch reflectarray schematic.

Variable-patch reflectarrays have numerous advantages when compared to other available reflectarray geometries [76], which make the design desirable for deployment at higher frequencies. Unlike ring element designs, the square patch reflectarray is easily fabricated with a high degree of accuracy using most micro-lithography techniques. The size of the patch scales well with frequency as opposed to reflectarray elements that utilize stubs to generate phase delays. By changing the length of the patch independently from the width, it is possible to impose phase delays in orthogonal directions for polarization selectivity [77], but the conventional variable-patch reflectarray does not inherently exhibit polarization sensitivity unlike some ring designs. Most significantly, variable-patch reflectarrays demonstrate superior bandwidths, approaching 10%, - far larger than almost all other single layer reflectarray designs [78].

#### 4.1.2 Waveguide Equivalent Circuit Model

The simplest approach to characterize the behavior of a variable-patch reflectarray is to break up the patches into individual, isolated unit cells and employ an equivalent circuit approximation analogous to the one developed in the previous chapter. Inherently a resonant structure, radiating patches are best represented as a terminated transmission line (Figure 42). Beginning at the termination of the transmission line, the reflectarray groundplane will behave as a short, which exhibits a reflection coefficient of -1 corresponding to the expected 180 degree phase shift upon reflection by a plane wave impinging on a perfect electric conductor (PEC) surface. The substrate of the reflectarray itself can be modeled as the transmission line by neglecting dielectric loss. Therefore, the standoff layer transmission line will exhibit a characteristic impedance equal to the wave impedance of the substrate material ( $Z_d$ ) and a length equal to the height of the substrate (d). Assuming the width and length of the patch are equal, the patch itself can be represented as a variable inductor [79], with inductance (L) proportional to the ratio of the length of the patch (l) relative to the length of the equally sized unit cell (l<sub>unit-cell</sub>):

$$L \propto \frac{l}{l_{unit-cell}} \tag{47}$$

Thus, when the length of the patch approaches zero, the inductor will behave as an open and when the length of the patch approaches the length of the unit cell, the inductor will behave as a short or a groundplane. The unit cell is finally connected to the open terminals of an infinite waveguide with characteristic impedance equal to the free space wave impedance to represent the air above the reflectarray.



Figure 42: Reflectarray transmission line equivalent.

Determination of phase shifting can be directly calculated by finding the input reflection coefficient at the interface of the substrate transmission line to the air transmission line. Using conventional transmission line calculations, the reflection coefficient ( $\Gamma_{in}$ ) is represented by

$$\Gamma_{in} = \frac{Z_{in} - Z_0}{Z_{in} + Z_0}$$
(48)

where the input impedance (Zin) is equal to

$$Z_{in} = \left(\frac{-j}{2\pi L} + \frac{-j\cot(\frac{2\pi\sqrt{\varepsilon_r d}}{\lambda_0})}{Z_d}\right)^{-1}$$
(49)

where  $\lambda_0$  is the free space wavelength and  $\epsilon_r$  is the real part of the substrate's dielectric constant. Finally the phase response of the reflectarray can be calculated by finding the phase of  $\Gamma_{in}$ , or

$$\angle \Gamma_{in} = \tan^{-1}\left(\frac{\operatorname{Im}(\Gamma_{in})}{\operatorname{Re}(\Gamma_{in})}\right)$$
(50)

The transmission line model allows for several significant conclusions about the general behavior of the variable-patch reflectarray. First, the height of the substrate will determine the extent of the phase shift achievable by the reflectarray by regulating the initial reflectarray phase response when the patch area is not large enough to introduce a significant inductance and only the groundplane is the dominant radiator. This phenomenon can be visualized by plotting the phase response of the reflectarray on a Smith Chart (Figure 43) starting at the dominant groundplane state and tracing out the entire system response as the patch increases in size. When the patch approaches the size of the unit cell, it will begin to behave as a short, independent of the standoff layer and groundplane. Unless the reflectarray element is at the same physical height as the groundplane or at a height corresponding to a multiple of one half the operating wavelength, the phase delay introduced by the groundplane and by the patch dominant states will not be equal and a complete rotation about the Smith Chart will not occur. From this behavior, and taking into account that no dielectric is entirely lossless, it is important to utilize the smallest possible substrate height to achieve nearly 360 degrees of phase shift at a single frequency. Another important conclusion is that the input impedance will always be purely imaginary and, thus,  $\Gamma_{in}$  will always have a magnitude of unity - signifying no losses in the system as expected in the idealized model. Additionally, the input impedance is dependent on the area of the patch, which will result in a non-linear relationship between the phase response of the patch and the length of the patch.



Figure 43: Smith chart plot of  $\Gamma_{in}$  demonstrating reflectarray phase transition by varying patch size.

The bandwidth properties of a variable size patch reflectarray may also be predicted. If the substrate height is reduced to allow for a larger phase response range at the center frequency, the transformed impedance of the groundplane will decrease to a short as wavelength increases – effectively shorting the patch and preventing operation. If wavelength decreases, the effective dielectric height will increase and the phase response range of the reflectarray will diminish. Additionally, operating bandwidth will not be linearly related to the ratio of dielectric height to wavelength.

Although the transmission line equivalent network is beneficial for predicting reflectarray behavior, it can rapidly become complex when accounting for system non-idealities such as dielectric or metal losses (as seen from the previous chapter). The physical dimensions of the patch does not correspond to the actual electrical length of the patch that the incident plane wave will observe due to substrate scaling and fringe fields. The model also cannot take into account surface coupling between neighboring elements without the introduction of a correction factor. Instead, these surfaces must be characterized using numerical modeling for the high degree of accuracy needed in design [80].

# 4.1.3 Improving Bandwidth

One of the greatest limitations of the variable-patch reflectarray is the device's limited bandwidth. Conventional reflective optical components, such as mirrors, have large operating bandwidths dependent entirely on material properties and physical dimensions. As demonstrated through transmission line equivalent circuit, the variable-patch reflectarray's bandwidth is defined by the patch's resonance, which is directly related to the dielectric height and the patch's relative electrical dimensions. Three possible approaches are possible to achieve bandwidth improvement in infrared reflectarrays: developing multi-band layouts, controlling the electromagnetic properties of the materials making up the reflectarray, and stacking reflectarray layers.

The first approach to improving bandwidth involves altering the patch layout to create a multi-band resonance. In this approach, dielectric height is fixed; however, patches are arranged to have multiple resonant frequencies. Design of multi-band reflectarrays are inherently difficult because of the non-uniform element spacing at higher frequencies and the complexity of the design layouts necessary to achieve more than two bands of resonance. Additionally, multi-band reflectarrays are not capable of continuous behavior between the two operating bands – limiting broadband operation. One way to ease these

challenges is to utilize an evolutionary algorithm to design the elements, such as using a genetic algorithm [81] or particle swarm algorithm. One such multi-band element, the slot loaded patch, will be discussed in 4.1.4.



Figure 44: Example multi-band reflectarray layout.

Another approach to improving reflectarray bandwidth requires varying the electromagnetic properties of the materials making up the structure. The easiest means to achieve bandwidth improvement by material variation is through controlled alteration of the substrate's permittivity. By replacing the reflectarray substrate with a piezoelectric, doped material, or photonic material [82] [83] and biasing or illuminating the device, it is possible to tune the substrate permittivity using the bias voltage to shift reflectarray resonance to a specific frequency band. Unlike the multi-band layout, however, the reflectarray will still be limited to a single, narrow band of operation at a fixed applied voltage.

The final approach to improving reflectarray bandwidth is through element stacking. Current research at microwave and millimeter-wave frequencies has demonstrated that stacking of reflectarray layers on top of one another will result in multiple bands of operation and, if desired, these bands can be continuous [84] - [86]. The main limitation of stacked reflectarrays is the increase in material losses due to the introduction of multiple metal layers and the increased complexity in modeling reflectarrays with more than two layers.



Figure 45: Stack-up for a multi-layer reflectarray.

#### 4.1.4 Achieving Phase Variation Greater Than 360°

From the transmission line equivalent circuit, the reflectarray is fundamentally limited to phasing of no more than 360 degrees. Due to the need to prevent the patch from shorting to the groundplane, a more practical limit usually falls around 300 degrees. Thus, if phase steps of less than 60 degrees over the complete unit circle are desired, alternative geometries must be developed. One such method, developed in more detail later, is slot loading. By introducing a slot into the center of the patch, the resonant current modes of the reflectarray are forced to traverse around the slot, effectively increasing their path length and, subsequently, phase delay. Slot elements are also desirable due to their low-loss and additional design flexibility by increasing the number of elements available to

the reflectarray designer given a set of fabrication limits (reflectarray elements and slot loaded reflectarray elements).

From an image quality or aberration correction standpoint it is also desirable to expand the phase delay capability of the reflectarray to even greater than 360 degrees. The simplest method is to fabricate the patch elements on to a stepped structure, similar to a non-planar FZP [87]. The stepped structure physically moves the effective location of the reflectarray location allowing it to better mimic a physical reflector. From a fabrication standpoint, this is not desirable as the reflectarray ceases to be a planar device. Another alternative, at the cost of losing polarization insensitivity, is to utilize slot loaded dipoles in place of patches (Figure 46). By increasing one dimension of the slot loaded dipole to the point it no longer resonates in the desired band of operation, it is possible to increase the electrical length current path of the resonant direction to long enough to yield phase delays surpassing  $2\pi$  [88].



Active Linear State

Figure 46: Dipole reflectarray with greater than 360 degrees of phase delay.

#### **4.1.5 The Dielectric Reflectarray**

As illustrated in Chapter 3, the presence of two metallic surfaces in the reflectarray design can lead to unwanted Ohmic loss. This Ohmic loss is especially undesirable at high frequencies due to Drude dispersion. One method for minimizing Ohmic loss is to construct the variable patch entirely out of a low-loss, but high-index dielectric material. In this type of design, a standoff layer is unnecessary and the phase contrast is determined by the thickness of the dielectric patch sitting directly on the groundplane. Because the surrounding media around the dielectric patch is air, this design has the further benefit of increasing the size of the effective unit cell to reduce fabrication critical dimensions.

To verify the dielectric reflectarray, a simple model was developed at 28.3 THz. In HFSS, a 900 nm thick layer of high-resistivity silicon was modeled on a gold

groundplane with a fixed unit cell of 7  $\mu$ m. The model, with generated fields, is shown in Figure 47. Varying the size of the dielectric patch, results in a phase plot analogous to the resonant antenna element described in Chapter 2 (Figure 48). While promising, the dielectric reflectarray will exhibit significantly greater chromatic variation and angular performance degradation compared to other reflectarray designs. This is due to the fact that the dielectric reflectarray utilizes interference between the top and bottom faces of the patch to achieve a band gap which yields a strong dependence on the dielectric patch's height.



Figure 47: Dielectric reflectarray at 28.3 THz. The brown box is the silicon patch and the concentration of the fields at the patch illustrates its ability to store field energy.


Figure 48: Modeled phase performance of dielectric reflectarray.

### 4.1.6 Circularly Polarized Reflection

As discussed in the development of the variable patch reflectarray, it is possible to introduce an asymmetry into the patch element to separately phase the two orthogonal current modes. Orthogonal phase shaping, assuming equal magnitude excitation in both orthogonal linear polarization states, allows for one polarization state to be delayed by 90 degrees with the intent of yielding circularly polarized re-radiation. Focusing of the circularly polarized radiation is still achievable because the asymmetric patches could still be spatially phased across the surface of the reflectarray, while maintaining the  $\pi/2$  relative phase delay between the two orthogonal polarization states. To verify that a circular polarized element was feasible an asymmetric patch was modeled in Designer (2.1 x 2.8 µm titanium patch, 75 nm thick, 1.17 µm thick BCB standoff layer, aluminum

groundplane, 5 μm square unit cell). Modeled results (Figure 49) show circularly polarized re-radiation at 9.8 μm.



Figure 49: Modeled reflection and phase properties of a circularly polarized reflectarray element.

The gray line corresponds to the difference in reflectivity (field) or  $\Delta R$  and the black line corresponds to the difference in phase upon reflection of the two orthogonal states or  $\Delta \theta$ .

# 4.1.7 Focusing Reflectarray Layout Development

The basic layout of the focusing reflectarray is derived from Figure 50 and [6]. In the figure, O is defined as the center of the reflectarray, F is the desired focal point of the array, f is the focal length (from O to F),  $Z_n$  is the nth zone in the reflectarray,  $\rho_n$  the distance from the  $Z_n$  to focal point, and  $r_n$  is the distance from origin to the outer edge of  $Z_n$ . The maximum phase variation that a polarization insensitive reflectarray element can achieve is typically  $2\pi$ , thus for a total of i unique, equally-spaced phase zones, the relative phase difference between each zone will be:



The focusing reflectarray is specifically designed to correct for the path difference between successive zones by introducing a zone-specific phase delay to allow the reradiated light to arrive in phase. Thus, relative phase can be related to the differences in path between adjacent zones and the focal point:

$$\Delta \theta_{n,n-1} = k \Delta \rho_{n,n-1} \tag{52}$$

(51)

$$\frac{2\pi}{i} = \frac{2\pi}{\lambda} \Delta \rho_{n,n-1} \tag{53}$$

$$\Delta \rho_{n,n-1} = \frac{\lambda}{i} \tag{54}$$

where  $\lambda$  is the center design wavelength and *k* is the propagation constant of the light in the media above the reflectarray. From the triangle OFZ<sub>n</sub> we can relate several terms:

$$\rho_n = \sqrt{f^2 + r_n^2} = f_{\gamma} \sqrt{1 + \frac{r_n^2}{f^2}}$$
(55)

Following the previous assumptions,  $r_n \ll f$ , which reduces the previous equation to:

$$\rho_n \approx f(1 + \frac{r_n^2}{2f^2}) = f + \frac{r_n^2}{2f}$$
(56)

This can then be related to the path difference between the equally spaced-in-phase zones by:

$$\rho_n - f = n\Delta\rho_{n,n-1} = \frac{n\lambda}{i}$$
(57)

Thus:

$$\frac{n\lambda}{i} = \rho_n - f \approx f + \frac{r_n^2}{2f} - f = \frac{r_n^2}{2f}$$
(58)

$$\frac{n\lambda}{i} \approx \frac{r_n^2}{2f} \tag{59}$$

$$r_n^2 \approx \frac{2fn\lambda}{i} \tag{60}$$

$$r_n \approx \sqrt{\frac{2 f n \lambda}{i}} \tag{61}$$

By using this equation, it is now possible to determine the distance from the origin of the array to the outer edge defining each zone. Although the calculations were done in only two dimensions, it is easy to show that the three-dimensional solution can be found by rotating  $OFZ_n$  about the line OF to maintain symmetry.

It is also of interest to determine the number of zones necessary to achieve a focusing design given a known aperture diameter, D. This can be done by substituting D/2 for  $r_n$  in the previous equation and by letting n equal N, the total number of zones:

$$\left(\frac{D}{2}\right)^2 = \frac{2fN\lambda}{i} \tag{62}$$

$$N = \frac{iD^2}{8f\lambda} \tag{63}$$

This can be written in terms of F/#:

$$N = \frac{iD}{8F / \# \lambda} \tag{64}$$

Once the design equations were determined, it was necessary to develop a placement algorithm for creating computer readable layouts used in fabrication of the reflectarray devices. This algorithm is included in the developed GDSII MATLAB Toolbox discussed in Appendix B. The program generates a binary GDSII layout file given layout files for each reflectarray element's geometry, the reflectarray's diameter, the design wavelength, and the desired F/#.

#### 4.1.8 Additional Fabrication Requirements

Fabrication of reflectarray devices is fundamentally the same as outlined in Section 2.6; however, additional steps must be taken to properly handle the substrates used in the construction of a practical reflectarray. Any physical height variations in the reflectarray's substrate will give rise to unwanted phase shaping, which necessitates the use of an optical flat (as demonstrated in Figure 51). The polishing processes used in the

fabrication of visible-grade optical flats typically require large substrate thickness to diameter ratios, on the order of 1:16 or larger. Most e-beam lithography systems, including the Leica EBPG5000+, can only handle thin substrates, no more than a few millimeters thick. To overcome this limitation, a custom adapter was machined that fits into the 5" mask holder of the system and allows the substrate to be recessed into the system by 3.175mm, below where the flat would sit in a standard wafer or piece-part holder. Two grounding clips were added on opposite sides of the recess to provide an electrical path to ground, thus prevent the wafer from charging. An image of this adapter is presented in Figure 52. Otherwise, the fabrication process is unchanged from the one presented in the earlier section.



Figure 51: 10.6 µm interferogram of a typical 380µm thick prime-grade silicon wafer exhibiting significant surface curvature.



Figure 52: Custom e-beam holder for thick substrates.

### **4.2 LWIR Reflectarray**

Initial development of the infrared reflectarray was carried out in the long wave infrared (LWIR). The long wave was chosen over other bands due to the availability of highenergy sources ( $CO_2$  laser) and larger wavelengths of operation for larger resonant element dimensions. The next few sections will discuss the design process starting with initial verification of reflectarray behavior in the infrared through the creation of a focusing device.

### **4.2.1 Initial Element Development**

To test the feasibility of an LWIR focusing reflectarray, a simple two inch, three-stripe reflectarray layout was developed (Figure 53). Three 6.3 mm by 28 mm arrays of

patches, isolated from each other on the dielectric by 8 mm, were deposited on a groundplane backed standoff layer for testing. Each uniform array was made up of a single sized patch element to demonstrate a unique phase shift upon reflection in measurement for comparison to modeled results. In addition, a 3.175 mm, quarter wave flatness (at 632.8 nm), silicon dioxide optical flat was used as the device's substrate to ensure that any phase modification was due to the patches only and not due to a physical defect in the substrate. The standoff layer of these devices was chosen to be 450 nm of  $ZrO_2$  with a 100 nm thick Au groundplane. All of the devices in the initial element development section were fabricated using a bi-layer resist process and all of the metal elements consisted of a 100 nm thick film of Au.



Figure 53: Proof of concept reflectarray layout.

Prior to fabrication of the first proof of concept devices, two initial fabrication runs were carried out on silicon wafers. The purpose of these runs was to characterize the e-beam fabrication process outlined earlier in the dissertation and to verify that 100  $\mu$ C/cm<sup>2</sup> was an appropriate dose for the reflectarray elements. The first sample consisted of a 4 by 4 dose matrix of arrays consisting of alternating rows of 2.98  $\mu$ m, 3.14  $\mu$ m, or 3.24  $\mu$ m size patches with a fixed periodicity of 5.54  $\mu$ m (Figure 54 and Figure 55). The goal of this run was to determine the necessary dose to achieve well-formed patches, verified by imaging of the device using a scanning electron microscope (SEM). The second run consisted of verifying the desired dose from the first fabrication run using a slightly larger array of patches and did not require metallization. With development complete, it was possible to use a visible microscope to observe the exposed pattern in the resist and verify that the predicted dose resulted in well-formed patches.



Figure 54: SEM image of metalized dose matrix.



Figure 55: Visible image of metalized dose matrix.

For the proof of concept, two devices were initially fabricated. The first device used the same patch dimensions as the initial dose matrix for three reflectarray rows - 2.98  $\mu$ m, 3.14  $\mu$ m, and 3.24  $\mu$ m (Figure 56). The patch size values were chosen based on early-modeled values that suggested a phase shift from 180 degrees of 40, 80, and 120 degrees, respectively, but were subsequently better predicted in later models. The three rows in the second device were chosen to provide additional sample points. The patch sizes were measured to be 2.82  $\mu$ m, 2.90  $\mu$ m, and 3.52  $\mu$ m (Figure 56). Images of one of the reflectarray stripes for the first device are presented in Figure 57 and Figure 58.



Figure 56: Fabricated strip reflectarray with reference sizes.



Figure 57: SEM micrograph of one of the stripes of the fabricated reflectarray.



Figure 58: Visible micrograph of one of the stripes of the fabricated reflectarray.

Measured results were interferograms taken of the two prototype devices using a Tywman-Green Interferometer (see Appendix A.2) and these results are presented in Figure 59 and Figure 60. Testing of these two devices were carried out at Lockheed Martin Corporation, Orlando, Florida. The first device tested was a coated optical flat. From the measured results, the optical flat is seen to exhibit excellent flatness at 10.6 µm that implies that all fringe shifts observed by the interferometer will be entirely a result of the reflectarray. Observation of the two fabricated reflectarrays successfully demonstrated that each row in the reflectarray does in fact demonstrate a unique phase shift (fringe shift) and that the phase shift is entirely dependent on the size of the reflectarray patches.

One unavoidable issue that arose from the testing was non-uniform illumination of the device under test. Although present in the testing of the first device, this phenomenon

was especially pronounced for the second device, which was tested a few weeks later. The non-uniform illumination was a result of misalignment of the interferometer and not a product of the devices under test. Additionally, this misalignment makes efficiency characterization with the interferometer impossible and increases the difficulty of post analysis. These issues motivated the construction of in-house interferometric capabilities for reflectarray characterization.



Figure 59: Interferogram of coated wafer. Non-uniform illumination is clearly present, with the bottom of the wafer "hotter" than the top of the wafer



Figure 60: Interferograms of fabricated reflectarrays.

With the measured data in hand, it was now possible to begin analysis of the results for relative phase extraction. For this purpose, a MATLAB function was written to carry out fringe following. Each interferogram, with limited masking to remove noise from the edge of the image, was read in and converted to an intensity matrix, normalized to account for the non-uniform illumination of the device, and cropped to contain only the center fringes passing over the reflectarray stripes. The MATLAB function then fit the cross-fringe profiles to a sinusoid. The primary purpose of the fitting was to allow for fringe smoothing and to reduce reflectarray edge noise and non-idealities in the testing setup. No numerical averaging was necessary, as the phase shift is optimized as a fitting parameter for the sinusoid. Results from both devices are presented in Figure 61 and Figure 62.





Figure 61: Smoothed phase response results for first reflectarray.



Figure 62: Smoothed phase response results for second reflectarray.

20 40

60 80

100 120 140

-20 L

To determine the actual phase shift introduced by the reflectarray stripes, a correction factor was introduced to account for the double pass offset due to the height difference between the reference groundplane and the reflectarray patches. The double pass phase shift for a 450 nm standoff layer with a permittivity of 2.0 was calculated to be 43.4 degrees. This problem did not affect modeling because the software assumed a fixed phase reference plane. By subtracting the phase offset from the values generated by the MATLAB function analysis, the phase shift introduced by the reflectarray was calculated and is summarized for the two proof of concept devices in Table 5. Table 5 also includes results from a third array fabricated using a slightly modified layout on a silicon wafer to accommodate a larger number of stripes with patch sizes of 3.3  $\mu$ m, 3.4  $\mu$ m, 3.7  $\mu$ m, 3.9  $\mu$ m, and 4.1  $\mu$ m.

Tuble 5. Wedsarde Telative phase shift vs. Tellecturity paten size.	
Patch Size	<b>Relative Phase Shift Upon Reflection</b>
2.82 μm	174.4 °
2.90 μm	172.4 °
2.98 μm	136.4 °
3.14 µm	124.1 °
3.24 µm	110.3 °
3.30 µm	79.7 °
3.40 µm	56.3 °
3.52 µm	-56.2 °
3.70 µm	-95.1 °
3.90 µm	-109.3 °
4.10 µm	-112.4 °

Table 5: Measured relative phase shift vs. reflectarray patch size.

In addition to verifying that reflectarray behavior was achievable at infrared frequencies, a secondary purpose of this portion of the infrared reflectarray research was to verify that phase behavior could be predicted using CEM. The results from Table 5 are plotted with results from an Ansoft Designer model in Figure 63. As demonstrated in the figure, excellent agreement is observed between measured and modeled results. Similar agreement was found using Ansoft HFSS.



Figure 63: Modeled (solid line) vs. measured (dark squares) results for initial element development.

#### **4.2.2** Low Loss Element Development

Following the positive results of the initial element's development, several modifications to the LWIR reflectarray's layout were introduced for production cost reduction and reflection efficiency gains. All of the gold metal layers in the design were replaced with aluminum for improved adhesion and lower deposition costs. The groundplane was also reduced to 80 nm thick and the elements making up the array were reduced to 75 nm to minimize phase errors caused by the physical height of the elements. In addition, the  $ZrO_2$  standoff layer was replaced with a layer of BCB, which has similar optical properties to  $ZrO_2$  without the need of sputter or evaporation deposition. The index of BCB, slightly higher than  $ZrO_2$  in LWIR, required a reduced unit cell periodicity of 5

μm. Fabrication also moved away from using an optical flat to a standard silicon wafer because of improvements in imaging capability from the in-house interferometer. The larger size of the silicon wafer allowed for more stripes to be tested per fabrication run and removed several tolerance issues inherent to fabrication with the thick substrates. Comparison of the new layout to the old is presented in Figure 64.



Figure 64: New layout prototype on a silicon wafer imaged below previous prototype on an optical flat.

For the reflectarray to be a viable technology for focusing applications, reflection loss of the patch elements must be minimized. Therefore, the thickness of the substrate layer was increased to 1.2  $\mu$ m to reduce loss introduced by the band gap of the resonant patches (Figure 65), with the added benefit of increasing the bandwidth of the reflectarray element. One detrimental result of the increased substrate thickness, as predicted by theory, was a decreased overall phase variation (Figure 66). To mitigate this issue, patches loaded with slots were determined through numerical modeling to provide the nearly complete phase transition required for up to a focusing reflectarray with phase steps of 45 degrees.



Figure 65: Modeled reflectarray reflectivity with varying standoff layer film thickness.



Figure 66: Modeled results for new layout (black line) vs. old layout (grey line). The black boxes represent slot loaded patch phase points (patch size =  $4.5 \mu m$ , slots = 1,  $1.5, 2, 2.5, 3, 3.5, 4 \mu m$ ).

A new prototype (bottom of Figure 64) was developed and tested to validate the numerical models of the proposed  $2^{nd}$  generation reflectarray element. These devices were tested for the first time at the IR Systems Lab using a newly developed 10.6 µm Twyman-Green interferometer. The new interferometer allowed for faster testing and better imaging with zoom when compared to the previous interferometer housed at Lockheed Martin (Figure 67). Figure 68 illustrates the improved contrast and lower loss of the new layout versus the older layout, while Figure 69 demonstrates the near 360-degree transition possible using slot loaded elements. Finally, two arrays of two different sized elements was fabricated demonstrating additional phase selectivity, with a phase response falling in between the two patch sizes (Figure 70).



OLD NEW Figure 67: Improved interferometer characterization capability demonstrating zoom, better edge transition resolution, and uniform illumination.



Figure 68: Improved efficiency of new layout with reflectarray regions denoted by red box.

The new layout exhibits better fringe contrast and has similar brightness to neighboring reference (no element) regions.



Figure 69: Square loop reflectarray element with reflectarray regions denoted by red box. Fringes across the reflectarray are nearly flush to the fringes in the neighboring reference regions due to a phase transition close to 360 degrees.



**1<sup>st</sup> Element 2<sup>nd</sup> Element Composite** Figure 70: Improved phase selection by mixing elements.

## 4.2.3 Continuous Reflectarray

Before a focusing reflectarray was attempted, a simple continuous phase variation reflectarray was fabricated. The reflectarray layout was rectangular and had a gradual decrease in patch dimensions from the top of the prototype to the bottom consisting of 10 sub-arrays (Figure 71). The purpose of this device was to verify that the individual elements making up the sub-array were not resolvable and that the transitions between differing elements were not significant. Measured results using the LWIR interferometer

verified that individual phase variations could not be isolated and that continuous phase transitions occurred between neighboring elements (Figure 72).

Element 1.5: 5250x525	
Element 2.0: 5250x525	
Element 2.3: 5250x525	
Element 2.4: 5250x525	
Element 2.5: 5250x525	
Element 2.6: 5250x525	
Element 2.7: 5250x525	
Element 2.8: 5250x525	
Element 3.0: 5250x525	
Element 4.0: 5250x525	

Figure 71: GDSII layout of continuous reflectarray.

Sizes were chosen to match the non-linear transition of the elements and the element dimensions are 1.5  $\mu$ m, 2  $\mu$ m, 2.3  $\mu$ m, 2.4  $\mu$ m, 2.5  $\mu$ m, 2.6  $\mu$ m, 2.7  $\mu$ m, 2.8  $\mu$ m, 3.0  $\mu$ m, and 4.0  $\mu$ m.



Figure 72: Continuous phase variation reflectarray device measured at 10.6 µm.

#### 4.2.4 Binary Focusing Reflectarray

For the first focusing prototype, only one discrete phase state of a two-element reflectarray was populated with patch elements of fixed dimension. The other phase state, or other element, was simply the combination of the groundplane and the standoff layer. Using the low-loss element layout dictated the use of aluminum patch elements of dimension 2.83 µm and thickness of 75 nm to achieve 180 degrees of phase shift upon reflection, relative to the region without patches. Unit cell spacing was 5 µm to prevent the appearance of grating lobes. A pattern layout was generated for an F/6, 25.4 mm diameter reflectarray with a total of 100 zones and greater than 20 million patch elements. This reflectarray layout is also notable since it is the largest known reflectarray ever tested in both number of elements and electrical diameter of the array [89]. One portion of the array is shown in Figure 73 and complete array resembled Figure 3. Modeled results predict a surface absorption loss due to the band gap of the reflectarray elements of 14%.



Figure 73: Visible micrograph of (a) reflectarray rings with layout schematic and (b) patch elements in the rings.

Measurement of the focusing reflectarray was done by imaging the beam profile of a collimated beam reflected off of the reflectarray. Specifically, the beam from a 10.6  $\mu$ m, 10 W CO<sub>2</sub> laser was initially expanded with a telescope to a collimated beam of diameter of 25.4 mm. The collimated beam was then directed by a beam splitter onto the surface of the reflectarray. The reflected focusing beam was projected onto a Spiricon pyroelectric detector array. The specific testing setup is discussed in greater detail in Appendix A.3. Determination of the optimal focal point was found by shifting the position of the device under test on a rail to minimize the focused spot size. Figure 74a is an image of the fabricated reflectarray element at optimal focus, 152.4 mm away from the camera. Introducing a defocus, as in Figure 74b, yields a nearly symmetric blur spot, as expected. Thus, a focusing reflectarray has been successfully demonstrated at infrared frequencies.



Figure 74: (a) Reflected beam profiles of reflectarray at optimal focus and (b) reflectarray outside of optimal focus.

### 4.2.5 Two-Element Focusing Reflectarray

When the binary reflectarray was originally tested, there was a question whether or not the focusing was arising entirely due to a diffractive effect. Based on modeling, the BCB layer thickness of 1.2  $\mu$ m would still provide a reduced volume compared to a groundplane-backed Soret FZP, which would require a standoff layer height of approximately  $\lambda/4$  or 1.8  $\mu$ m. To validate that the reflectarray was focusing due to the resonant elements, and not another phenomenon, the same two-element layout was fabricated again, but with elements in every zone. The two-elements chosen were 2  $\mu$ m (138.22 degrees) and 3.45  $\mu$ m (-43.15 degrees), for a relative phase difference of 180 degrees. Measurements of this surface demonstrated the same behavior as the binary reflectarray. Modeled results predict a surface absorption loss due to the band gap of the reflectarray elements of 13%.

#### 4.2.6 Eight-Element Focusing Reflectarray

Finally, to verify that a smaller focused spot (reduced power into the higher order foci) was possible with a graded reflectarray, an eight-element reflectarray was developed using the same F/# and diameter as before. The eight-element reflectarray has phase steps of 45 degrees, as opposed to the 180 degrees in the two previous designs, which should yield 95% of the reflected power into the primary focus [6]. The eight elements used included (from center zone outward) a blank region (180 degrees), 2.025  $\mu$ m patch (135.82 degrees), 2.35  $\mu$ m patch (91.74 degrees), 2.60  $\mu$ m patch (45.28 degrees), 2.825  $\mu$ m patch (7.74 degrees), 3.5  $\mu$ m patch (-45.12 degrees), 4.5  $\mu$ m patch loaded with a 2.25  $\mu$ m slot (-89. 14 degrees), and a 4.5  $\mu$ m patch loaded with a 4  $\mu$ m slot (-132.43 degrees).

Modeled results predict a surface absorption loss due to the band gap of the reflectarray elements of 16%. The reflectarray was fabricated and the measured results are shown Figure 75. As expected, the focused spot is significantly smaller, with a diameter less than 200  $\mu$ m, as opposed to the greater than 500  $\mu$ m spot size observed with the binary reflectarray. In spite of the smaller spot, significant stray light is present and lower power is observed in the focus spot than the two previous designs.



Figure 75: 8 element graded reflectarray. While a smaller focus (purple spot) is observed, significant stray light (colored gray) is present.

# 4.3 MWIR Reflectarray Element Development

Following the progress of the LWIR reflectarray, it was of high interest to expand IR reflectarray technology into the MWIR. To aid in measuring these new devices, a MWIR Tywman-Green interferometer was constructed. The MWIR interferometer was built on

the same optical train as the LWIR interferometer using two turn mirrors and a helium neon (HeNe) laser operating at  $3.39\mu$ m (Figure 76). Due to the low power of the HeNe, a platinum silicide (PtSi) camera was utilized in place of the 10.6  $\mu$ m Pyrocam.



Figure 76: 3.39 µm interferometer setup using portions of the 10.6 µm optical train. The wire grid polarizer allows for power control since the source laser is linearly polarized.

With testing capability in hand, development of the MWIR reflectarray followed an identical path as the earlier, multi-stripe LWIR reflectarray. BCB was again chosen as the standoff layer due to the material's low-loss in the MWIR and ease of deposition. Aluminum was similarly picked due to the metal's high reflectivity and limited damping at  $3.39 \mu m$ , although the patch thickness was reduced to 40 nm. New Designer models were developed and the optimal thickness of the BCB standoff layer was found to be 350 nm, which required adding additional solvent to thin the BCB. An optimal unit cell length of 1.5  $\mu m$  was chosen. A total of 9 patch sizes were fabricated on a silicon wafer and

imaged. Figure 77 contains several imaged stripes demonstrating unique fringe shifting. Measured results are presented against modeled results in Figure 78. Some surface curvature was observed, due to the substrate, but this was accounted for during analysis. Thus, the feasibility of reflectarray behavior in the MWIR has been demonstrated.





Figure 77: Four stripes of the 3.39µm reflectarray prototype demonstrating unique fringe shifting at each stripe.



Figure 78: Measured and modeled results for the 3.39 µm reflectarray prototype. Solid line is modeled data and data points correspond to measured results.

### 4.4 NIR Reflectarray Element Development

Similarly, it was of high interest to expand the reflectarray technology into the NIR. For testing, a custom interferometer system was built (Figure 79). The source for the NIR interferometer was a 1.55 µm diode laser and the detector was a SUI Goodrich camera.



Figure 79: 1.55 µm interferometer source and testing setup.

As with the MWIR design, BCB and aluminum (30 nm thickness for patches) were determined to be optimal materials for use in development of the NIR reflectarray elements. Another model was developed and the thickness of the BCB standoff layer was chosen to be 200 nm with a unit cell length of 0.8  $\mu$ m. Figure 80 contains images of two of the total 9 patch sizes fabricated demonstrating fringe shifting. Measured results are presented against modeled results in Figure 81. Several complications were identified in development of the NIR Reflectarray:

• Great care must be taken to ensure that a uniform thickness BCB layer is applied, due to the dilution need to achieve layer thicknesses of 200 nm. The discrepancies

between modeled and measured results are due to height variations in the standoff layer

- Patch elements of size 100, 150, and 200 nm were written, but destroyed during lift-off due to their small size
- Patch elements of size 250 and 300 nm exhibited array damage during lift-off
- Significant substrate curvature was observed, leading to potential error in measured phase values.

It is important to note that all four issues can be readily corrected. The MWIR reflectarray was fabricated after the NIR reflectarray using two layers of thinned BCB and did not contain significant height variation. Extremely small elements do not appear to be necessary and alternative approaches for the fabrication of sub-micron elements are available in the lab. Finally, surface curvature can be reduced by carrying out fabrication to use an optical flat. All of these issues will be addressed in future work; however, the results do demonstrate the feasibility of reflectarray behavior in the NIR.



Figure 80: Two stripes of 1.55µm reflectarray prototype demonstrating unique fringe shifting at each stripe.



Figure 81: Measured and modeled results for the 1.55µm reflectarray prototype. Solid line is modeled data and data points correspond to measured results.

# **4.5 Aberration Behavior**

The aberration behavior of the reflectarray arises from two separate sources. The first is due to the geometrical shape of the reflectarray and the second is due to the chromatic and angular behavior of the elements making up the reflectarray. The geometrical aberrations are well documented in the FZP community [6] [90]. Prior to this work, no known investigations have been carried out in the aberration performance of metamaterial surfaces as imaging devices. The next two sections will investigate methods for predicting the aberration performance of a reflectarray.

# 4.5.1 Angular

From section 4.1.7, the one-dimensional optical path difference (OPD) of a graded reflectarray can be written as (following the process outlined in [90]):

$$OPD = \rho_n - f \tag{65}$$

 $\rho_n$  can be rewritten on axis as:

$$\rho_n = \sqrt{f^2 + r_n^2} \tag{66}$$

which yields:

$$OPD_{on-axis} = \sqrt{f^2 + r_n^2} - f \tag{67}$$

Thus can be expanded with a power series:

$$OPD_{on-axis} = \frac{r_n^2}{2f} - \frac{r_n^4}{8f^3} + \dots$$
(68)

It should be noted that this term is equal to  $\rho_{n,n-I}$  (difference in path between adjacent zones, from before) if the higher order term is neglected. The higher order term represents the spherical aberration component of the graded FZP at normal incidence.

The OPD developed can be expanded to account for an incident marginal ray at an angle of  $\alpha$ . The off-axis OPD can be represented as:

$$OPD_{off-axis} = r_n \sin \alpha + f(\sqrt{1 + (\tan \alpha - \frac{r_n}{f})^2} - \sqrt{1 + \tan^2 \alpha})$$
(69)

which can be expanded to:

$$OPD_{off-axis} = \frac{r_n^2}{2f} - \frac{r_n^4}{8f^3} + \frac{r_n^3\alpha}{2f^2} - \frac{3r_n^3\alpha^2}{4f}$$
(70)

The first term of the off-axis OPD represents the diffraction-limited performance and the fourth term represents the spherical aberration, as before. The third term describes coma and the second term describes astigmatism and field curvature, with no distortion present. Several important conclusions can be made from this result. Remembering that:

$$r_n \approx \sqrt{\frac{2 fn\lambda}{i}} \tag{71}$$

and

$$N = \frac{iD}{8F / \# \lambda} \tag{72}$$

it can be shown that increasing the diameter or decreasing F/# of the graded reflectarray will result in a greater coma contribution. Expansion into two dimensions is not shown and would involve the inclusion of an appropriate  $\cos(\varphi)$  term (see [90]).

Inclusion of the performance of the reflectarray elements is not trivial. The phase delay of the elements are dependent on both the incident angle of the radiation, as shown in Figure 13, and, thus, the angle of re-radiation, from reciprocity. From that same figure, the additional phase delay introduced by off-angle radiation can be expressed as:

$$\theta_{addtional,n} = \theta_n(\alpha) + \theta_n(\beta_n) \tag{73}$$

Where  $\alpha$  is the incident angle,  $\beta_n$  is the angle formed by OZ<sub>n</sub>F from Figure 50, and  $\theta_n$  is the relative phase difference between the off-axis and on-axis responses of the reflectarray element. Phase delay can be related to path length (*l*) by:

$$l = \frac{\lambda}{2\pi} \theta \tag{74}$$

Thus, the additional path contribution from one of the reflectarray's sub-zones at nonnormal illumination/re-radiation would be:

$$l_{addtional,n} = \frac{\lambda}{2\pi} [\theta_n(\alpha) + \theta_n(\beta_n)]$$
(75)

This term can incorporated directly into the two OPD expressions developed previously:

$$OPD_{on-axis} = \frac{r_n^2}{2f} - \frac{r_n^4}{8f^3} + \frac{\lambda}{2\pi} \theta_n(\beta_n)$$
(76)

$$OPD_{off-axis} = \frac{r_n^2}{2f} - \frac{r_n^4}{8f^3} + \frac{r_n^3\alpha}{2f^2} - \frac{3r_n^3\alpha^2}{4f} + \frac{\lambda}{2\pi} [\theta_n(\alpha) + \theta_n(\beta_n)]$$
(77)

The next step for finding the aberration contribution of the reflectarray elements is to develop an analytical representation of the off-axis behavior of the reflectarray. Unfortunately, there is no way to predict this response prior to modeling, as shown in previous sections 2.4, 2.5, and 4.1.2. Nevertheless, from Figure 13 and assuming only TE polarized radiation, the phase delay introduced due to non-normal incident or re-radiated radiation will have a quasi-sigmoid form resembling [91]:
$$\theta_n(x) = \theta_0 + \frac{A}{1 + e^{-|\frac{x}{B}|}}$$
(78)

Where *x* is some arbitrary angle,  $\theta_0$  is an offset term and *A* and *B* are structure dependent terms. Using a Taylor expansion, converting to path length, and neglecting the piston terms, the equation can be expanded out to:

$$l_n(x) = \frac{A\lambda}{2\pi} \left[ \frac{1}{4} \frac{x}{B} - \frac{1}{48} \left( \frac{x}{B} \right)^3 + \dots \right]$$
(79)

Several exciting conclusions can be made from this result. First, the off-axis behavior of the reflectarray will primarily contribute to the tilt (first term) and to the coma (second term) of the ideal graded reflectarray developed previously. Secondly, even at normal incidence, the reflectarray can produce off-axis aberrations due to  $\beta_n$  not being equal to zero at the edges of the reflectarray:

$$Max(\beta_n) = \tan^{-1}(\frac{1}{2}F/\#^{-1})$$
(80)

Thus, smaller F/# reflectarrays will experience more significant aberration contributions arising from the antenna elements making up the array. This type of behavior is completely different from similar refractive or diffractive surfaces where only spherical aberration should be present at normal illumination. Figure 82 shows an astigmatism dominated imaged spot from a reflectarray device for nearly normally incident illumination.



Figure 82: Images of binary reflectarray aberration behavior: (left) image of blur spot with detector placed inside focus and (right) detector placed outside focus demonstrating astigmatism.

In general, it is desirable to get rid of the on-axis higher order aberrations at normal incidence. To do this, each sub-zone of the graded reflectarray will have to be corrected for their position within the array and their corresponding  $\beta_n$ . As mentioned previously, no simple expression exists for determining the off-axis behavior of a reflectarray element. Instead, the reflectarray layout generator can be hooked directly into a database or CEM to pick out reflectarray elements on the fly (Figure 83). Furthermore, it should be feasible to select elements with phase delays to correct for the on-axis spherical aberration allowing for near diffraction limit imaging without altering the array's layout. Similarly, the discrete phase control achievable by the reflectarray can be utilized for more complicated correction, including suppression of system aberrations.



Figure 83: Future design method for aberration corrected reflectarray.

# 4.5.2 Chromatic

The graded reflectarray will once again obey the aberration behavior of the FZP. Following the development from [90], chromatic aberration will not be appreciable until the OPD has been shifted by a quarter-wave. This can be used to find the bandwidth of the structure:

$$OPD' = \frac{n\lambda'}{2i} \pm \frac{\lambda'}{4}$$
(81)

Setting this equal to the  $\rho_{n,n-1}$  can yield the bandwidth of the structure:

$$\frac{n\lambda}{2i} = \frac{n\lambda'}{2i} \pm \frac{\lambda'}{4}$$
(82)

$$\Delta \lambda = \frac{\lambda i}{N} \tag{83}$$

Thus, increasing the diameter or decreasing the F/# will result in a decrease in the effective bandwidth of the reflectarray.

As with the angular aberration discussion, the chromatic behavior of the reflectarray element is quasi-sigmoid (Figure 10). This behavior is detrimental to the bandwidth of

the reflectarray. The phase delay introduced by changing the incident wavelength can be written as

$$\theta_n(\lambda) = \theta_0 - \frac{A}{1 + e^{-\frac{|\lambda|}{B}}}$$
(84)

where again A and B are dependent on the structure of the resonant element. From this expression, as the wavelength is increased, the phase delay introduced by the reflectarray will be decreased resulting in a further loss of path length. This phenomenon arises from the electrical current path length of the resonant element decreasing with larger wavelengths. Thus, assuming minimal dispersion in the materials making up the radiating structure, there appears to be no means of changing the direction of the phase curve to increase the phase delay at longer wavelengths, which would allow for achromatic type designs.

# **CHAPTER 5: THE INFRARED TRANSMITARRAY**

#### **5.1 Fundamental Theory**

Similar to the reflectarray in terms of layout and advantages, transmitarray devices work by varying the impedance of a 2-dimensional surface with resonant antenna elements to focus or diverge an impinging beam of known propagation direction upon transmission (Figure 84). Wavefront shaping still occurs due to the superposition of the fields reradiated by the resonant elements. The specific wavefront formed is dependent on the effective phase delay of each element in the array. Transmitarrays are desirable over graded index (GRIN) lenses in the infrared due to the lack of choices for transparent dielectrics in the band with significantly different indices. These surfaces are also substantially easier to fabricate then similarly behaved kinoform, transmissive FZPs.



Figure 84: Planar transmitarray focusing a planar wavefront.

Transmitarrays, in various forms, have existed since the mid-1980s [92], but have only experienced limited research interest. Most transmitarray designs are three-dimensional, have been proposed for use as power combiners [93], and consist of two patch arrays (one receiving and one transmitting) interconnected with two vias at either end of a delay line and an amplifier in either a tile [94] [95] or tray configuration [96]. Three-dimensional passive transmitarrays can be constructed by simply removing the amplifier [97]. Planar, passive transmitarrays have been realized through replacing the receiving patch array with a slot array and directly coupling the transmitting array [98] [99]. It should also be noted that the transmitarray defined here should not be confused with similarly named radar transmit arrays [100] or magnetic resonance transmit arrays [101].

In addition to the reasons already provided, the planar, passive transmitarray shares most of the advantages of its reflectarray cousin. The impedance, or progressive phase of the surface, is discrete, which allows for sub-wavelength control of aberrations. Transmitarrays are planar surfaces and can incorporate utility stacking or fabrication on a non-planar surface. The elements making up the array can be loaded with tunable elements for electronically controllable beamsteering. Furthermore, transmitarrays are lithographically generated and potentially cheaper than kinoform lenses or similar refractive lenses.

#### 5.2 Single Layer Transmitarray

The most logical way to investigate the feasibility of a single layer transmitarray is to follow the same design approach as the reflectarray, assuming that reciprocity is

126

maintained. To that end, a square patch transmitarray was developed. Due to a flaw in HFSS maintaining the polarity of the waveports used in the transmissive IAWM, it was not possible to model the patch with variable patch sizes. Instead, a fixed 1.8  $\mu$ m by 1.8  $\mu$ m patch was modeled over a portion of the LWIR that should provide a means for predicting the phase contrast available to the transmitarray at a fixed frequency. The patch was placed on an infinite half-space of silicon, was 75 nm thick, and had a periodicity of 2.5  $\mu$ m. Modeled results are presented in Figure 85. Unfortunately, only a small amount of phase variation is achieved (~120 degrees) while incurring significant reject band loss – far greater than the reflectarray.



Figure 85: Square patch transmitarray modeled performance. The gray line corresponds to the transmittivity (field) of the transmitarray and the black line corresponds phase delay introduced by the transmitarray.

To help mitigate the loss in the band gap of the transmitarray, a slot patch was also modeled. From Babinet's principle, the slot should exhibit the complimentary performance of the solid element in terms of spectral transmission. The results of this model, using the same periodicity and materials as before is presented in Figure 86. The slot exhibits a small range of possible phases. The same model was repeated using a cross slot with arm widths of 400 nm and lengths equal to  $1.8 \mu m$  to increase the strength of the resonance. The cross slot did improve the phase range, but not enough to be practical in a focusing device (Figure 87). Both slots will also require the usage of a lower index substrate material to reduce the reflection loss in the passband.



Figure 86: Patch slot transmitarray modeled performance. The gray line corresponds to the transmittivity (field) of the transmitarray and the black line corresponds phase delay introduced by the transmitarray.



Figure 87: Cross slot transmitarray modeled performance. The gray line corresponds to the transmittivity (field) of the transmitarray and the black line corresponds phase delay introduced by the transmitarray.

### 5.3 Feasibility and Alternatives To The Infrared Transmitarray

Modeling indicates that resonating elements make poor phased transmission devices. It could be easily argued that phase contrast could be increased through multiple layers or the use of aperture coupling to introduce a quasi-groundplane [98] [99]. Both of the approaches are undesirable due to the considerable effort necessary to align multiple layers with unit cell spacing in the single digit micron range. The lack of a groundplane, when compared to the reflectarray, also leads to a significant loss caused by the backward radiated lobe. Furthermore, the extreme loss of these structures rejection bands strongly limit the practicality of these surfaces when compared to far superior surfaces such as dielectric FZP or even GRIN lenses.

Some alternative approaches exist. The most obvious approach would be to utilize nonresonant antenna elements, analogous to the meanderline structure discussed in the first chapter. Exploiting the inductive or capacitive delay of these structures would accomplish the same goal as the transmitarray. Devices of this type include the metal grid double square slotted elements as shown in [6]. Another alternative would be to utilize all dielectric surfaces to approximate the indexes necessary to form a GRIN like surface. This type of surface would be analogous to a spatially varied moth-eye surface [102].

# **CHAPTER 6: THE INFRARED EMITARRAY**

#### **6.1 Fundamental Theory**

The formation of thermally excited, coherent emission using planar, ordered surfaces has been investigated over the past few years for potential applications in efficient energy harvesting [103] and related fields. Grating based emitters were one of the earliest surfaces shown to demonstrate coherent emission, specifically directional emission patterns [104], linear polarized emission [105], circular polarized emission, and short range coherence [106]. Prior publications have also noted the similarity of these surfaces to classical, resonant antennas [107], but have typically analyzed their behavior using band theory or plasmonics [108]. While promising, grating based emitters are limited in their practicality due to complexity of excitation and fundamental geometrical simplicity.

Emission frequency selective surfaces (eFSS) are promising alternative to diffractive phased surfaces. eFSS, just like illuminated absorber FSS, consist of an array of passive, resonant antenna elements fabricated above a groundplane forming a resonant cavity. The array will emit when in physical contact with a thermal source, such as a hot plate, and does not require localized excitation. Because eFSS require a groundplane to function, these structures do not rely on transparent substrates, like many emissive diffractive grating designs, and allow for direct spectral and emission magnitude control, independent of the surface of the thermal source. Furthermore, these surfaces enjoy the possibility of multi-band operation and robust design using well-established computational electromagnetic techniques. The eventual goal of this research is to utilize the phasing capabilities of an eFSS to focus emitted radiation, analogous to the reflectarray surface, thus creating an emitarray.

#### 6.1.1 Relating Thermal Performance to Electromagnetic Performance

The emissivity of a surface can be related to its electromagnetic properties through two relationships: the conservation of energy and Kirchhoff's law. The conservation of energy relates the reflectivity ( $\rho$ ), transmitivity ( $\tau$ ), and absorptivity ( $\alpha$ ):

$$\alpha(\lambda) + \tau(\lambda) + \rho(\lambda) = 1 \tag{85}$$

Kirchhoff's law for a surface at equilibrium relates the surface absorptivity to its emissivity ( $\epsilon$ ):

$$\alpha(\lambda) = \varepsilon(\lambda) \tag{86}$$

Kirchoff's law can then be substituted into the conservation of energy equation to yield:

$$\varepsilon(\lambda) = 1 - \tau(\lambda) - \rho(\lambda) \tag{87}$$

Thus, modeled or measured reflectivity and transmission data for a given surface can be used to directly calculate the emissivity of that surface.

Furthermore, for groundplane backed surfaces where transmitivity is equal to zero, such as the eFSS, reflectivity can be directly related to emissivity through:

$$\varepsilon(\lambda) = 1 - \rho(\lambda) \tag{88}$$

This conclusion has been validated in multiple publications including [109] and [110].

## 6.1.2 Phasing Thermal Emission

As mentioned previously, the performance of an illuminated, groundplane backed, planar resonant antenna element, assuming sub-wavelength unit cell spacing and minimal shorting to the groundplane, will experience some degree of coupling (mutual or otherwise) to any neighboring elements. This coupling will directly impact the resonance properties of the array, serving as a capacitive loading. Thus, given that the spectral reflection and emission performance of antenna element are directly related, then the thermally excited currents in an emitting eFSS must also be influenced by this coupling.

Consider two identical dipoles, with resonance in the infrared, in near proximity to one another. As the dipoles are heated, currents are excited on the surface of each element. The antennas begin to resonate and mutual coupling occurs due to the fields excited by the currents. In this way, short-range coherence is established. Now consider if one of the dipoles is slightly shorter than the other. Even though the elements are no longer identical, they will still couple and experience a capacitive load due to the coupling. From the analysis of the reflectarray, at a given frequency in the band gap of the two dipoles, the two elements should radiate out of phase. Assuming that the array is expanded out to an infinite extent, once the array reaches equilibrium, all of the elements in the array should be emitting coherently with one another because the current path on the surface of each dipole is significantly less than the coherence length. The exception to this rule would occur at the edges of the array where edge effect may become significant. However, if the eFSS element is symmetric about two or more planes and allows for orthogonal current modes to exist at resonance, no field mechanisms are present to ensure coherence between the two resonant modes. The lack of coherence between orthogonal current modes is significant and precludes the use of many reflective polarizer designs as polarized or directional emitters, since the two orthogonal polarization states will not radiate in phase. Thus, the only way to achieve circularly polarized emission or unpolarized coherent emission is to utilize a resonant element with a strong crosspolarization.

From prior publications, detecting or analytically proving coherent emission is extremely difficult. The rest of the chapter will initially look at some methods for demonstrating long-range coherent emission through polarization. Finally, an emitarray surface will be discussed and initial results will be presented.

### **6.2 Polarized Emission**

Polarized emitting materials have been demonstrated previously – most notably lasers and chiral materials [111]. The issue with all of these approaches is they require nonplanar structures or structures secondary to the emission surface (such as a linear polarizer) to achieve polarization. Planar eFSS are an excellent alternative to these surfaces and can be significantly easier to fabricate and tailor to meet specific emission needs. The next three sub-sections will investigate one linearly polarized eFSS and two planar circularly polarized emitters.

#### 6.2.1 Linearly Polarized Emission by Asymmetric Resonant Elements

Linearly polarized resonant elements can be realized through the usage of a dual band FSS element. The purpose of this experiment is to verify that emission from an eFSS can be polarized. The simplest dual-band emitter element is the dipole patch. By ensuring the width of the dipole is small enough that it will not resonate in the same band as the dipole itself, the element will emit strongly linearly polarized light, polarized along the direction of the length of the dipole. It should be noted that the dipole element is not the only geometry with dual-band behavior; thus, more complicated designs could be developed depending on the spectral properties desired.

Fabrication of the dipole-based linearly polarized emission surface was carried out using the standard electron beam lithography process outlined earlier. The groundplane and supporting structure of the device consisted of a 380  $\mu$ m thick silicon wafer with an 85nm-thick film of aluminum deposited directly on the wafer. A 1.2  $\mu$ m film of BCB was spun on to isolate the elements from the groundplane, and the asymmetric emission elements were made up of a lossy 100 nm film of titanium. The dipole antenna elements were 2.9  $\mu$ m by 0.5  $\mu$ m. The square unit cell spacing was 5  $\mu$ m. Measured polarization results using a polarizer and a power meter for the fabricated elements are presented in Table 6 and demonstrate linearly polarized emission. Even though an aperture was mounted to exclude the un-populated regions surrounding the emission surface, significant leakage was observed around the edges of the aperture diminishing the contrast of the co-polarized and cross-polarized power.

Table 6. Measured emitted power from a dipole array heated to 100°C.	
<b>Polarization State</b>	Power
Co-Polarized	0.125 mW
Cross-Polarized	0.095 mW

Table 6: Measured emitted power from a dipole array heated to 100 °C.

For visual verification and a better demonstration of the linear nature of the emission surface, a pattern using the University of Central Florida's emblem, the Pegasus, was fabricated using orthogonally orientated dipoles (Figure 88). The Pegasus was 27.5 mm in diameter. Images of this device (Figure 89) produced using an integrated 8-12  $\mu$ m camera and a linear polarizer demonstrate the high contrast response of the dipole over a large bandwidth. When the array is imaged without a polarizer present in the optical train of the Pegasus, as in Figure 90, both orthogonal dipoles emit equally and the shape cannot be resolved.



Figure 88: UCF Pegasus logo with dipole orientations.



Figure 89: Images of the linearly polarized Pegasus with the camera's linear polarizer (a) horizontally oriented and (b) vertically oriented.



Figure 90: Images of the linearly polarized Pegasus with the camera's linear polarizer not present. The dark spot in the circle is due to a defect in the array.

# 6.2.2 Circularly Polarized Emission by a Compact Multi-Layer Structure

Achieving circular polarization from thermal emission is significantly more challenging than linear polarization. Instead of shifting one linear polarization state to emit into a separate spectral band or suppressing emission into that state completely, circular polarization requires emission into both linear states with a 90° phase difference between the two states. Prior to this research, the best way to achieve circular polarized emission is to linearly polarize the emitted radiation using a wire-grid polarizer and align the now linearly polarized radiation at a 45° angle relative to a quarter-wave plate. The quarterwave plate delays one orthogonal state of the linearly polarized radiation by 90° allowing circular polarization to be transmitted. The process is shown in Figure 91 and described in detail in [25].



Figure 91: State-of-the-art circular polarizer for emitted radiation. In this schematic, the fast axis of the quarter-wave plate is rotated from the normal of the page 45° about an axis parallel to the incident radiation's propagation (black line). The wire grid is orientated in the plane of the paper, perpendicular to the incident radiation's propagation (black line).

While fundamentally simple, the process of using a wire-grid polarizer and a quarterwave plate is inefficient and inelegant. The most obvious limitation is that the wire-grid polarizer will suppress half of the emitted light, greatly reducing the power of an already inefficient emission process and causing unwanted scatter and thermal loading. The wire grid, the surface closest to the emitting surface, is not resonant, which means that another coating or structured surface must be placed on the emitting surface if spectral or magnitude control is desired. Furthermore, previous designs have typically mounted the polarizer and plate on a separate substrate that cannot be easily placed in contact with the emitting surface due to structural or thermal shock reasons [112]. This leads to volumetrically large systems and limits the possibility of conformal polarization control.

All of these limitations can be corrected for by replacing the wire-grid polarizer with a linearly polarized emission surface. The linearly polarized surface will allow for more emitted energy in the band of interest than the 50% allowed by the wire-grid, since the undesired polarization emission path is not present and all emission from the heated surface must occur in the desired polarization state only, if designed properly. Spectral and emission magnitude control are readily available by altering the geometry and metal of the periodic asymmetric elements. The system is also significantly more compact than the wire-grid design with the substrate replaced with an electrically thin, thermal insulation layer. A schematic for this polarizer is presented in Figure 92.



Figure 92: Compact, multi-layer circular polarizer for emitted radiation. In this schematic, the fast axis of the quarter-wave plate is rotated from the normal of the page 45° about an axis parallel to the incident radiation's propagation (black line). The linearly polarized emission surface is orientated for emission in the plane of the paper, perpendicular to the incident radiation's propagation (black line).

# 6.2.3 Circularly Polarized Emission by a Single-Layer Structure

As mentioned before, to achieve circularly polarized emission using a planar element, a structure with no symmetry and the presence of two interacting orthogonal modes must be utilized. To meet that end, a tripole [113] eFSS structure was adapted. The tripole was initially chosen due to its significant cross-polarization and the fact the two orthogonal polarization states share common current paths. By shortening one of the legs of tripole, it is possible to shift the resonance of the element resulting in a progressive phasing between the orthogonal linear polarization states (Figure 93). Choosing the length of the leg to result in a phase delay of 90 degrees and minimizing the difference in emissivity between the two orthogonal states, the resulting tripole will have circularly polarized

emission. The modeled results of an asymmetric tripole with two legs 1.65  $\mu$ m long, one leg 1.05  $\mu$ m, and an arm width of 0.5  $\mu$ m are presented. The elements had a unit cell spacing of 3.5 by 4.0  $\mu$ m. The modeled device exhibits a fairly large band of operation from about 8.5 to 10.8  $\mu$ m. Other than the geometry of the element, all design parameters are identical to the linearly polarized eFSS in the previous section.



Figure 93: Asymmetric tripole. The red line corresponds to the horizontal current mode and the blue line corresponds to the vertical current mode.



Figure 94: Modeled emission and phase properties of asymmetric tripole. The gray line corresponds to the difference in emissivity and the black line corresponds to the difference in phase upon emission.

The asymmetric tripole was fabricated into a four-square checkerboard pattern as shown in Figure 95. The piece was then mounted on a hot plate and imaged twice with a linear polarizer; aligned once in the horizontal direction and once in the vertical direction, as shown in Figure 96. The arrays are seen to exhibit a degree of linearly polarized emission, which is consistent with the model ( $\Delta\epsilon$  is not equal to zero over the entire 8 to 12 µm band). Next, a quarter-wave plate was introduced and the array was re-imaged in the two orthogonal circularly polarized states (Figure 97). The images suggest that the arrays exhibit circularly polarized emission, as only the quarter-wave plate polarizer was rotated when taking the pictures. There are several reasons to be suspicious of this result: the quarter-wave plate used was a meanderline polarizer, which, while more broadband than a conventional crystal quarter-wave plate, exhibits appreciable chromatism and also exhibits a preference towards one linear polarization state due to fabrication errors. Never the less, these initial results are promising and suggest the need for additional investigation.



Figure 95: Visible micrograph of four-quadrant checkerboard of asymmetric tripoles. Each quadrant has a different emission polarization than the other two adjacent regions.



Figure 96: Two thermal images (side by side) of checkerboard asymmetric tripole with linear polarizer filter present in front of the camera in two orthogonal directions.



Figure 97: Two thermal images (side by side) of checkerboard asymmetric tripole with circular polarizer in front of the camera and in two orthogonal directions.

## 6.3 Mutually Coupling Induced, Long-Range Coherent Emission: The Emitarray

All of these results lead up to the question of whether it is possible to exploit interelement coupling to achieve far-field, focused, spectral emission. From reciprocity, any reflectarray designed in the dissertation, assuming a coherent thermal excitation, should emit into a focused spot at the design frequency. In reality, the previously designed reflectarrays cannot focus emission due to the low cross-polarization of the patch elements preventing coherence between the two polarization states. This limitation can be overcome by using a high cross-polarization element, such as the tripole used in the previous section. On a more practical note, focused emission is extremely difficult to detect given the presence of significantly stronger out-of-band emission due to the broad nature of resonant elements in the infrared. Thin-film-based narrowband infrared filters typically have too large of bandwidth to be beneficial. Diffractive based directional emitters have overcome this issue through the use of multiple-pin-hole optical trains to suppress un-directed emission [114]. Regardless of the noted limitations, the two-element reflectarray developed in Chapter 4 was mounted on a hot plate, heated to approximately 120 °C, and imaged using the IR microscope described in A.4 and a 10 - 11 µm narrowband filter. With the object plane of the microscope placed at the hypothetical image plane of the reflectarray, the expected focused spot was not observed. Unexpectedly, slightly tilting the hot plate resulted in the appearance of a null at the focal point of the reflectarray that was not present when a non-structured wafer was mounted on the active hot plate and similarly tilted. At the moment, no method for validation if this null is due to far-field interference, nevertheless, it is a promising result.



Figure 98: Thermal image of reflectarray at focus. The rings of the reflectarray can be seen suggesting no interference is occurring.



Figure 99: Thermal image of reflectarray at focus with slight tilt added to the hot plate. A null can be clearly seen suggesting that interference may be occurring and that coherence is being maintained.

# **CHAPTER 7: CONCLUSIONS**

#### 7.1 Summary

The purpose of this dissertation was to investigate the use of surfaces consisting of subwavelength, resonant antenna elements for phase shifting, beam steering, and focusing of electromagnetic radiation in the infrared. All aspects of surface interaction has been considered for phasing – reflection, transmission, and emission. The overriding goal of the dissertation was to establish a new class of optical component as an alternative to conventional diffractive or polished elements. These surfaces are especially exciting due to their low footprint and design flexibility.

Investigation of these phased surfaces began with the development of the modeling and fabrication techniques necessary for creating passive, planar phase elements with resonance in the infrared. This development included descriptions of the chromatic and angular performance of a typical groundplane-backed element. Due to the complexity of these devices, it was necessary to explore multiple modeling techniques including FEM and MoM. Finally, a fabrication procedure for these devices was developed using electron beam lithography.

With the several modeling approaches developed, the unique properties of the infrared materials used in forming resonant antenna elements were investigated. Initial work using PMM and Designer demonstrated the need for accurate material properties from ellipsometry and inclusion in modeling software. The metamaterial damping

phenomenon was analytically developed and demonstrated in FSS arrays. Damping has a critical effect on the performance of phased devices due to its impact on both resonant frequency and bandwidth.

The first device investigated experimentally was the infrared reflectarray, a reflective phase shaping device. From a waveguide circuit model, the performance of a patch elemnt in the variable reflectarray was investigated. Initial testing of phased reflective resonant antenna elements was demonstrated in the LWIR and then expanded into the MWIR and NIR. With the performance of the antenna elements known, three focusing reflectarrays were developed in the LWIR, including an eight-element graded reflectarray was concluded with an investigation of the surface's aberration behavior.

The next two devices considered were the transmitarray and emitarray. The transmitarray, through modeling, was determined to be too lossy to be practical. After initial development demonstrating the relationship between reflection and emission properties, several polarized emission surfaces were fabricated and tested demonstrating the feasibility of coherent emission. Initial testing of an emitarray prototype demonstrated promising results include the presence of a possible far-field null.

In general, there are numerous aspects of this dissertation that can be seen as a notable contributions to the field of Electrical Engineering and Optics:

- Establishment of a new class of infrared focusing element that is both practical and offers new properties not found in existing elements at the infrared
- First documented demonstration of a reflectarray operating in the infrared frequency band
- Highest documented frequency for reflectarray element operation (193.5 THz) and focusing reflectarray operation (28.3 THz)
- First exploration of the feasibility of transmitarrays in the infrared
- Exploration of FSS absorber phasing and demonstration of FSS directional emission
- First documented demonstration of a polarized-emission FSS.

## 7.2 Future Work

In addition to the static reflectarray, it would be desirable to have a tunable focusing surface for beamsteering at infrared. This would allow for an electronically steerable field of regard reflector, without the need for bulky gimble systems. Tuning has already been demonstrated in the rf portion of the spectrum and an infrared tunable device could be realized either using high-frequency diodes or photo-conductive elements.

Another area of interest would be to integrate a reflectarray surface with an antennacoupled bolometer. Prior work [115] has suggested that similar surfaces will yield improved D\* and capture area. One of the primary advantages of the reflectarray is it can be placed behind the detector to ensure minimal loss by allowing the detector to be illuminated first before interacting with the wavefront shaping surface. A further area of investigation for these surfaces is commercial feasibility. Every device tested in this dissertation was fabricated using electron beam lithography. E-beam lithography, while highly accurate and excellent for prototyping, is an extremely expensive and slow process. Future devices will have to be fabricated using more cost-effective methods such as nano-imprint lithography or immersion lithography. Similarly, investigations will have to be carried out to devise methods for encapsulating the phased devices for use outside of the laboratory environment. For reflectarray surfaces, this will not be especially challenging; however, determining capping layers that will not interfere with emitarray surfaces will be a much larger challenge. Results of initial investigations on how to reduce material costs and have been presented.

As with any work, there is considerable need for future improvement and revision. With elements measured in the MWIR and NIR it should be possible to fabricate focusing devices in these bands. Additional research can be done to improve the understanding of aberrated performance of these types of surface, with an emphasis on integrating into a commercially available optical system design package such as CodeV or ZEMAX. With the aberration treatment provided, it should be possible to develop a new LWIR reflectarray with better image quality by tapering the phase of the elements on the edge of the aperture to correct for the change in re-radiation angle. Developing a method of properly testing the emitarray would open doors into demonstrating long-range, wide-band focused emission. Finally, numerous other reflectarray designs were proposed, such

as the circularly polarized reflectarray, the broadband reflectarray, and the dielectric reflectarray that have not yet been verified in the laboratory.

# **APPENDIX A: PHASE CHARACTERIZATION IN THE INFRARED**

## A.1 The Variable Angle Spectral Ellipsometer

### A.1.1 The Variable Angle Spectral Ellipsometer Fundamentals

The variable angle spectral ellipsometer (VASE) is a useful tool for the characterization of the optical properties of deposited thin-films and for wideband phase characterization of transmissive and off-axis reflective phased devices. The specific ellipsometer used in the course of this dissertation is a J.A. Woollam IR-VASE (Figure 100). The IR-VASE has an operating spectral range of approximately 2 to 45  $\mu$ m with the spectral range determined by the IR-VASE's source (a glowbar) and the cut-off of the optics in the system. In terms of thin-film characterization, the instrument is capable of measuring standard fabrication materials, film thickness, bulk materials, films consisting of multiple layers, and anisotropic materials.



Figure 100: J. A. Woollam IR-VASE.

The measurement process of the ellipsometer is described in detail in [116]. The key component to recognize is that the ellipsometer, when not in FTIR mode, measures the  $\Psi$ 

and  $\Delta$  of the surface under test. These terms are related to the measured change in polarization upon reflection from the test surface by:

$$\rho = \tan(\Psi)e^{j\Delta} = \frac{R_p}{R_s}$$
(89)

where  $R_p$  is the ratio of the reflected and incident *parallel* polarized light and  $R_s$  is the ratio of the reflected and incident *senkrecht* polarized light.  $\Psi$  and  $\Delta$  can then be fit using an oscillator model for determination of the optical properties of the surface. Furthermore, if the orientation of the incident light is known,  $\Delta$  can be used to determine the relative phase difference between two known states and  $tan(\Psi)$  can be used to determine the relative magnitude difference between the two states.

#### A.1.2 IR-VASE Characterization of the Optical Properties of Thin Films

Detailed instructions on how to use the IR-VASE and WVASE software for characterization of thin-films at IR frequencies falls outside the scope of this dissertation and is documented in [117]. The following sub-sections contain recommendations and techniques used in measuring the materials presented in this dissertation. As with all ellipsometry, these are general recommendations and exceptions are possible and expected in many situations.

#### A.1.2.1 Sample Preparation

In general, the easiest thin-film materials to characterize are metals. Metal films should be deposited with the desired fabrication thickness or, for a nominal measurement, at a thickness greater than three (maximum-wavelength) skin depths. In this case, the substrate the metal is deposited on is not an issue, so long as it is reasonably flat, since the metal should not be transmissive.

Dielectric or semiconducting materials; however, require additional care when characterizing. Because these films are typically optically transmissive over some portion of the IR band, the choice of substrate material can greatly influence the accuracy of the measurement. Ideally, a high reflectivity material should always be used as a substrate to ensure high index contrast with the thin-film and to limit unwanted transmission through the substrate, which will lead to a reduced signal-to-noise ratio. Thus, the best material to use as a substrate when characterizing transparent thin-films in the infrared is a low-resistivity metal such as gold or aluminum. If these materials pose a processing problem or if the index of the thin-film material is sufficiently low, a high index material, such as silicon or germanium, can also be used. Lossy materials such as silicon dioxide should be avoided because characterization is difficult in the loss bands of these substrates.

#### A.1.2.2 Minimizing Backside Reflections

Unlike visible ellipsometers, the IR-VASE is susceptible to un-wanted interference from light reflecting off the backside of the substrate. This is due to the coherence length of the incident radiation in the infrared being on the order of the thickness of a typical wafer. Several methods to avoid this issue include using metal substrates, roughening the backside of the substrate, using optically thick substrates, or treating the backside of the substrate with a scattering material.

### A.1.2.3 Ensuring Unique/Realistic Solutions When Modeling

When fitting oscillators to the measured  $\Psi$  and  $\Delta$ , it may be tempting to fit every single spectral feature with an oscillator or employ overly complicated oscillator models (such as the Gaussian-Lorentz or the PSEMI). This approach should be avoid at all costs and will lead to non-realistic and non-unique solutions. To increase the uniqueness of the fit, several simple steps should be followed:

- Start from an existing model, if possible. Most of the existing models should already be unique.
- When developing a new model, consult prior publications to determine where oscillations should be present or for approximate dc conductivities or carrier concentrations.
- If no information is available about the material, the simplest model is typically the best (Occam's Razor). Additional oscillators, non-ideal model effects, and angular offsets should only be added if they are realistic and have an appreciable impact on the mean square error (MSE) of the model.
- Use appropriate oscillator models depending on the portion spectrum that was measured. This includes using Lorentzian oscillators in the visible and Gaussian oscillators in the infrared.
- When in doubt, check the "Fit Statistics." No free variable should have a correlation greater than 0.75 and the 90% confidence limits should be realistic in a unique fit.
## A.1.2.4 Miscellaneous Tips

- As the system ages, the glowbar will gradually lose power. The system must be monitored to ensure that the proper voltage is being applied to the system. Over biasing the system will increase the signal-to-noise ratio of the machine at the risk of reducing the life of the glowbar.
- Additional signal-to-noise gains can be achieved by increasing the averaging of the system.
- In addition to ellipsometry, the machine can also act as a standard FTIR. In this configuration, the ellipsometer will supply  $R_p$  and  $R_s$ .
- For visibly transparent materials, best effort alignment may be necessary, as the alignment laser is a HeNe. In these situations, it may also be desirable to attempt transmissive emissivity if possible, which is only sensitive to the tilt of the surface and not its z-position. Best effort alignment is also necessary for rough surfaces.
- For testing of films smaller than the incident beam, the sample should be physically affixed to the surface of a larger wafer for mounting on the IR-VASE. Successful masking of materials smaller than the incident beam has been achieved using a "sticky-note" to cover the exposed mounting surface.
- Anisotropic thin-films are difficult to characterize and, typically, can be treated as isotropic. Thicker anisotropic films should always be characterized with transmission and reflection ellipsometry over several angles for determination of the orientation of the anisotropic molecules. It should also be noted that materials such as polymers might exhibit slight anisotropic behavior due to film stress occurring during deposition [118].

- As a general rule of thumb, non-ideal modeling is rarely necessary, although, adding an angle offset is typically beneficial.
- Fitting should always be done over the entire spectrum of the measurement because out-of-band oscillations still influence the material's performance in-band (i.e. ordinary dispersion). The most common exception to this rule is when the measurement has sufficiently large depolarization in the long and/or short wave that the measured results are invalid.

## A.1.3 IR-VASE Characterization of Phased Devices

As mentioned previously, the IR-VASE can also be used to characterize the phase properties of a surface through the value  $\Delta$ . For transmissive devices, the *senkrecht* and *parallel* polarizations can be aligned about the y- and x-axis of the element at normal incidence. By introducing non-symmetry in the element about the two axes, it is possible to use  $\Delta$  as the relative phase difference between the two orthogonal dimensions. For a patch element, one orthogonal state could be reduced in width to form a dipole and suppress resonance in that direction. Thus, in this example,  $\Delta$ , at the frequency of interest, would be a direct measurement of the phase delay introduced by a patch element of length equal to the length of the dipole. Similar steps could be taken for reflected phase measurement; however, this must be done at an angle of 26 degrees or greater.

#### A.2 The Twyman-Green Interferometer

The Twyman-Green interferometer is a modified version of the Michelson interferometer [119]. For a typical unequal-path Twyman-Green (Figure 101), a coherent,

monochromatic light source (typically a laser) is focused into a pinhole and collimated with a collimating lens (telescope) to form planar wavefronts. Half of the reference beam passes through a beam splitter and reflects off the test surface or test mirror back into the interferometer. The other half of the reference beam reflects off the beam splitter and impinges on a flat reference surface, typically a gold mirror on a piezoelectric substrate or mechanical stage for tilt adjustment and correction, and is reflected back into the interferometer. The beams are then redirected to an IR camera or detector by the beam splitter (neglecting the beam reflected back into the laser) for imaging of the generated interference pattern. This resulting image is also known as an interferogram.



**Twyman-Green Interferometer** 

Figure 101: Twyman-Green interferometer schematic [120].

The interference pattern imaged at the detector can be utilized to determine surface variations between the device under test and the reference mirror. If an ideal test mirror

was placed perfectly flat to the reference mirror, the two reflected beams will arrive at the detector with a spatially uniform phase difference and the detector will image a uniform illumination across the field of view (the image will be bright if the beams arrive in phase and dark if the beams arrive out of phase). If either the reference mirror or the test surface were at a slight tilt, a series of parallel bright and light fringes would be imaged, corresponding to the interference resulting from the two beams experiencing different path lengths. Thus, increasing the tilt of the either surface further will increase the number of fringes imaged corresponding to the increased change in path difference across the plane of tilt.

Assuming the test surface had some local physical deformity, such as a curvature, the fringes of the generated interference pattern of the tilted mirrors would be shifted to reflect the difference in path length introduced by the physical height difference arising from the deformity. By measuring the fringe shift relative to a known reference position, it is possible to determine relative height error difference at various points across the test device by relating phase difference to path length difference. As such, the Twyman-Green is referred to as a two-pass interferometer because a physical height change on the test device will result in a fringe shift (phase shift,  $\theta_{fringeshift}$ ) in the interference corresponding to twice the height of the deformity:

$$\theta_{fringeshift} = \frac{2*height}{\lambda_0} * 360^{\circ}$$
<sup>(90)</sup>

This is a result of the path length for both the incoming wave and reflected wave being altered equally by the height change. Therefore, the Twyman-Green can only resolve height differences equal to half the wavelength, otherwise wrapping will occur and the height difference cannot be uniquely determined using the techniques outlined in this appendix.

For testing of resonant elements in the infrared, a similar approach can be used to determine phase shifting. Although these devices are typically physically flat, the phase difference introduced upon reflection will still introduce interference, which can be characterized using the interferometer. For the reflectarray proof of concept devices, the interference fringes were placed orthogonal to the longest extent of the patch rows and, thus, phase shifting by the reflectarray elements can be measured relative from the regions outside the reflectarray stripes (standoff layer and groundplane only) by observing how much the fringes shifts. Unlike conventional polished optics characterization; however, the interferometer is no longer double pass, as phase shifting occurs upon re-radiation and the reflected or incident path length is unchanged, reducing Eq. (90) to (91):

$$\theta_{fringeshift} = \frac{height(effective)}{\lambda_0} * 360^{\circ}$$
(91)

This also allows for the measurement of phase variations greater than 180 degrees or onehalf wavelength. Additionally, it should be noted that the fringe shifting introduced by the patches is linearly shifted compared to reference regions due to the double pass height difference between the groundplane and the patch height.

#### A.2.1 Identifying Phase Shaping Elements in Interferograms

Identification of reflected phase modulation by resonant elements can be found directly from observation of the surface's interferogram. First, consider Figure 102, which is an interferogram of an optically flat surface. Dark regions of the interferograms are points where the wavefronts reflected off the reference and test surface destructively interfere and bright points correspond to regions where the two beams fully constructively interfere. In the figure, the light and dark fringes are parallel and equally spaced across the entire image indicating, from Eq. (90) and the recognition that either the reference or the test surface is physically tilted, there is no or little phase shifting caused by the surface. This behavior is expected from a flat surface, free of physical height deformity. It should be noted that the fringes only need to be parallel for the surface to be flat relative to the reference - rotation of the fringes about the center of the interferogram only indicates a rotation of the tilt plane.



Figure 102: Measured interferogram of a nearly flat surface.

Next, consider Figure 103, which contains an array of phased elements surrounded by regions of no elements (standoff layer and groundplane only) on the top and the bottom. In addition to a darker bright fringe due to element loss, the region containing the phased elements can be identified by a shifting of the dark and light fringes to the right (in this example). Assuming that the element region is uniform in its phasing, the amount each fringe has been shifted relative to the no element region should be equal for each fringe, moving from the left to the right and the top to bottom. For this specific device, the phase shift is not uniform and, thus, the fringe has a curved sag indicating a larger phase shift at the center of the element stripe compared to the edge of the array.



Figure 103: Measured interferogram of a region of phase shifting elements. This caption is at the bottom of the figure.

In addition to spatially varying phase across the array, further considerations must be made before calculation of element phase shaping. Figure 104 demonstrates what happens when one element has a phase shift greater than 180 degrees. In the top array, the fringes have been shifted to the right, just like in the case of Figure 103. In the bottom array, the fringe has been shifted in the opposite direction indicating that one of the arrays is generating a phase variation greater than 180 degrees. The reason for this will be discussed in the next section. Another issue that can arise when imaging phased element arrays is reference regions with curvature. Both Figure 103 and Figure 104 have parallel

and well-aligned reference regions that make analysis straightforward. In Figure 80, for example, the reference regions are no longer parallel or equally spaced indicating that the substrate has a physical deformity or curvature. Curved reference regions can be corrected for in post processing or by always using an optically flat substrate.



Figure 104: Measured interferogram of two arrays of phase shaping elements with one array having a phase variation greater than 180 degrees.

### A.2.2 Analyzing Interferograms for Phase Information

Once a phase shifting region has been identified, along with a point of reference, the phase of the elements in the region can be found with spatial measurement. Figure 105 shows an ideal interferogram consisting of three reference regions (R1, R2, and R3) and two-element stripes (S1 and S2). For the purpose of this analysis, only the dark fringes will be considered, since the brighter regions can be difficult to observe if the element arrays have loss (as in Figure 104). From interference theory, the distance between the center of two neighboring dark fringes (points of complete destructive interference) in a single pass system will be 180 degrees (from Eq. (5)). Element phase delays of less than

180 degrees will result in the fringe being shifted along the direction of the reference mirror that is tilted closest to the beam splitter (assuming only the reference mirror has been tilted). Element phase delays greater than 180 degrees, but less than 360 degrees, can be rewritten as  $180 - \theta$ . Negative phase delays will cause the fringe to shift along the direction of the reference mirror that is tilted furthest away from the beam splitter. Phase delays greater than 360 degrees will wrap and cannot be directly determined using this method. For the purpose of this example, Figure 105 is assumed to have been generated with the reference mirror tilted such that the right side of the Figure is closest to the beam splitter. Thus, S1 has a phase delay less than 180 degrees and S2 has a phase delay greater than 180 degrees.



Figure 105: Ideal interferogram of two phase-shaping arrays. Phased regions are denoted by S1 and S2. Reference regions are denoted by R1, R2, and R3.

Because the distance between the dark fringes in the reference regions is known to be 180 degrees, the phase shifting caused by each stripe region can be found by the ratio of the spatial distance between the reference fringe and the shifted region and the total distance between two reference fringes. These values can be found using a ruler or using an image manipulation program such as Photoshop or The GIMP to measure pixel distance. Thus, for S1, the phase delay would be (using The GIMP to measure):

$$\theta_{S1} = 50 \, px / 150 \, px * 180^\circ = 60^\circ \tag{92}$$

Similarly, S2's phase delay can be found with correction for the fact the phase delay is greater than 180 degrees:

$$\theta_{S1} = 50 \, px / 150 \, px \,^* - 180^\circ = -60^\circ = 300^\circ \tag{93}$$

From this analysis, it is clear that the phase resolution (smallest phase variation that can be uniquely measured) of the interferogram is equal to the inverse of the number of pixels between the two dark fringes divided by two (Nyquist Criterion). In this example, the smallest resolvable phase variation is equal to:

$$\theta_{resolve} = 1/(2*150\,px)*180^\circ = 0.6^\circ \tag{94}$$

#### A.2.3 Interferogram Analysis Post Processing

Post processing of the phase calculated from the interferogram takes two forms: correction for physical height differences between the reference and phasing elements and correction for non-idealities in the substrate. Assuming a uniform standoff layer thickness with known index of refraction, the linear phase difference ( $\theta_{offset}$ ) between the reference region can be calculated by:

$$\theta_{offset} = \frac{2*thickness*n}{\lambda_0} * 360^{\circ}$$
(95)

Thus the actual phase delay introduced by the resonant array is equal to:

$$\theta_{actual} = \theta_{fringeshift} - \theta_{offset} \tag{96}$$

It should be noted that this calculation is only valid if the phase reference plane is at the height of the elements, which is customarily the case in modeling.

Correction for substrate non-idealities is more complicated. The simplest error to correct for is fringe tilt. Figure 106 is the same design as Figure 105, but with a gradual (nonrealistic) tilt introduced. The tilt can be readily indentified by the fact the top and bottom reference fringes are no longer parallel about the y-axis only. Correction can either be carried out during testing by rotating the substrate slightly, after testing by rotating the image, or through numerical analysis (the next section). Non-linear surface or device variations (such as those seen in Figure 80, Figure 103, and Figure 104) cannot be as easily corrected for and require averaging, additional numerical analysis, and best guess efforts by the designer.



Figure 106: Ideal interferogram of two phase-shaping arrays with tilt added.

# A.2.4 Interferogram Analysis using MATLAB

Prior to this point, all analysis has been carried out by hand, which limits the ability of the designer to correct for issues such as surface tilt or curvature. This section will discuss one method to use MATLAB to automate the analysis of interferograms for phase extraction through a series of code snippets:

```
%Read image file
ifg = imread(file);
%Normalize Plot
ifg = cast(ifg, 'double'); %need to cast to renormalize
min_val = min(min(ifg));
ifg = ifg - min_val;
max_val = max(max(ifg));
ifg = ifg * 60 / max val;
```

The first step is to import the image file (stored in the variable *file*) and normalize the array to a known maximum value (60 is chosen for improved plotting). Note any linear magnitude offset is removed from the image, since only phase information is needed.

```
ifg = ifg(start_cord(2):end_cord(2),start_cord(1):end_cord(1));
[Y_len,X_len] = size(ifg);
y_points = 1:Y_len;
x_points = 0:X_len-1;
for idx = y_points
    data = ifg(idx,:) - min(ifg(idx,:));
    ifg(idx,:) = data * 60/max(data);
end
```

Next, the code clips the image to contain only reference regions and element stripes (defined by *start\_cord* and *end\_cord*). The design is then renormalized by taking x-directional slices with a fixed y-coordinate to remove any unwanted loss and normalize the fringes across both the reference and stripe regions in magnitude (the fringes are assumed to be in the y-direction with the long portion of the element arrays orientated orthogonally along the x-direction).

```
%Seed
vars = [44 0];
%Curve fit the interferogram to sinusoids
for idx = y_points
    vars = fminsearch(@merit_sine,[vars(1) vars(2)], [],
x_points,ifg(idx,:));
    f_values(idx) = vars(1);
end
f_value = mean(f_values);
function [merit] = merit_sine(vars,x,data)
sine=30*sin(2*pi*x/vars(1) + vars(2)) + 30;
merit=mean((data-sine).^2);
```

Using a minimum searching algorithm, with an assumed phase delay of 0 degrees and a frequency of 44, MATLAB fits the interferogram fringes to a sinusoid to determine the frequency (spacing) of the fringes. The frequency is found at each y-coordinate and then averaged over all y-coordinates to find the fixed fringe spacing across the surface (assuming that the test surface does not have significant surface curvature).

```
var = 0;
for idx = y_points
    var = fminsearch(@merit_sine2,[var], [],
f_value,x_points,ifg(idx,:));
    phi_values(idx) = var;
end
function [merit] = merit_sine2(var,F,x,data)
sine=30*sin(2*pi*x/F + var) + 30;
merit=mean((data-sine).^2);
```

Finally, the minimum searching algorithm is rerun to determine the phase delay necessary to fit the fringes to a sinusoid with the frequency determined previously. The variable *phi\_values* contains the phase delay across the surface indexed by the y-coordinate. Once these values are known, additional steps can be taken to account and correct for surface non-idealities by using matrix manipulation. The results of this algorithm can be seen in Figure 61 and Figure 62.

#### A.2.5 Interferometry of Transmissive Devices

Similar analysis of transmissive devices can be carried out using a Mach-Zehnder interferometer. This type of analysis falls outside the scope of this dissertation, but this interferometer is discussed in detail in [119].

## A.3 At-Focus Spot Imaging

Characterization of complete, focusing reflectarray devices was carried out using at-focus spot imaging (AFSI). AFSI requires placement of the imaging detector in the plane of the primary focus of the reflectarray to image the focused spot formed. This presents an issue because the image plane of the reflectarray falls within the path of the incident beam. To overcome this challenge, a modified Twyman-Green configuration (Figure 101) was employed. For the AFSI setup, the reference mirror was replaced with a scattering or absorbing surface and the test surface was replaced with the focusing element. With the second channel disabled, no interference will occur and the detector can directly image the reflected spot without blocking the incident beam (Figure 107).



Figure 107: AFSI test setup.

The image shows the end of the telescope of the Twyman-Green Interferometer. The collimating lens is shown in the foreground, the beam splitter is in the center, the reflectarray is on the adjustable mount on the right, the detector is shown on the left (SPIRICON camera), and the concrete block is being used to block the second channel.

Additional care must be used when carrying out AFSI. Placement of the reflectarray too far away from the beam splitter, such that the image plane falls close to the location of the beam splitter, may lead to localized heating and undesirable damage. Thus, it is highly recommended that the reflectarray be placed as close as possible to the beam splitter when testing. For aberration characterization, a flat, reflective surface should initially be used in place of reflectarray for characterization of baseline system aberrations (from the laser, lenses, or beam splitter).

### A.4 The Infrared Microscope

AFSI is an appropriate method for verifying surface focusing and approximate focal length, but is not necessarily the most effective method for characterizing surface aberrations. For characterizing aberrations, it was necessary to construct and infrared microscope. By placing the image plane of the reflective surface at the object plane of the microscope, it is possible to magnify the focused spot generated and characterize the shape of the spot for unwanted aberrations. The specific microscope constructed is shown in Figure 108 and consisted of a Raytheon LWIR camera front-end lens to form a telescope with the camera (focused at infinity). Using this setup, the field of view was approximately 5mm by 5mm.



Figure 108: Infrared microscope.

# **APPENDIX B: GDSII TOOLBOX**

#### **B.1 About the GDSII Toolbox**

To provide the progressive phasing across the surface of the reflectarray necessary to impose a spherical phase front, a layout generator was initially developed based on the theory presented in Section 4.1.7 to individually place reflectarray elements for optimal performance. The program was written using MATLAB and automatically generates a binary GDSII layout file given layout files for the reflectarray element's geometry, the reflectarray's diameter, the design wavelength, and the desired F/#. From this original program, the GDSII Toolbox was developed.

The GDSII Toolbox is a robust collection of MATLAB functions specifically for automating the process of array generation in nanofabrication. The toolbox was written in a way that someone, without understanding of GDSII file structure, could create complicated layouts and elements directly in the flexible MATLAB environment. Resulting GDSII file sizes from the toolbox are smaller than identical designs generated by the commercially available editor used at the Infrared Systems Laboratory and file generation times are typically greater than 5,000 elements written a second.

## **B.2 Using the GDSII Toolbox**

Installation of the GDSII Toolbox is possible by placing the toolbox's \*.m in the MATLAB path. To run the pre-developed modules (higher level functions), a user only needs to run "gdsii\_toolbox" in the MATLAB console and follow the resulting prompts. A total of 6 top-level modules are available in the original version:

• array\_gen – Generates a square or circular array.

- image2pattern Generates an array layout based on a grayscale image. Elements are placed based on the relative darkness of each pixel.
- image2pattern\_0 The same as image2pattern with bright regions left without elements.
- fzp the original reflectarray layout generator.
- $fzp_0$  the same as fzp, but the first repeated zone is left without elements.
- element\_gen Generates a single patch, loop, cross, or meander line element.
   This function is useful for pre-generating elements for use in the other modules.

## **B.3 Creating New GDSII Toolbox Modules**

The easiest way to describe how to create a new toolbox module is to consider an existing

module, array\_gen:

```
%Function for generating a circular or rectangular array.
%Form: array gen()
2
%Parameters:
%This function does not take any parameters from command line.
2
%Returns:
%GDS-II file of desired circular pattern.
%Copyright 2008
%IR Systems Lab, CREOL
%University of Central Florida
%Last updated 07/22/2008
%Changelog
%10/29/2007 - Initial Version
%10/31/2007 - Minor UI cleanup
%07/22/2008 - Updated to match code in write array
              Improved circle generating algorithm
0/2
function array gen()
```

The initial section contains help information about the function, as well as copyright information and a changelog. The function name should match the name of the \*.m file and should be something easy to remember. Once completed, the new function should be added to gdsii toolbox.

The variable section is the portion of the code where all of the global functions variables are defined. This includes the preference filename (pref\_fname), dialog strings (p\_string1, p\_string2, dlg\_title1, dlg\_title2, geom\_sname1, geom\_sname2, num\_lines), and other useful variables (flag). At the start of all functions, the write\_units('fetch') command may be run to determine the software's defaults units.

The function first prompts the user whether they would like to use an existing element geometry or generate a new geometry using element\_gen. To invoke element\_gen in automatic mode, the function must pass it a filename and element\_gen will return the passed filename. For this situation, the actual filename of the temporary element gdsii\_file is not important to the user. If the user wishes to use an existing GDSII file, locate\_gdsii can be used to prompt the user for which file to use and the function will return the desired filename. The value passed to locatate\_gdsii is the number of times the user should be prompted (for batch opening). It should be also noted that the flag is no longer zero if a new element is created.

```
%Get additional information from the user
%Put up prompt
[ptr, res] =
listdlg('PromptString',p string2,'Name',dlg title2,'SelectionMode','sin
gle', 'ListSize', [225 200], 'ListString', geom sname2);
%Did the user cancel?
if ~res
   return;
end
if ptr == 1
   geom = 'rect';
   prompt = {'Element Periodicity (in X, um):', 'Element Periodicity
(in Y, um):','Number of Elements in Array (in X):','Number of Elements
in Array (in Y):'};
   defAns = open pref(pref fname, 4);
```

```
answer = inputdlg(prompt,dlg title2,num lines,defAns,'on');
    %Check for abort
    if isempty(answer)
        return;
    end
    %Convert all of the user input
   ARRAYX = str2double(cell2mat(answer(3,:)));
   ARRAYY = str2double(cell2mat(answer(4,:)));
   ELEMENT PERIODICITYX = str2double(cell2mat(answer(1,:)));
   ELEMENT PERIODICITYY = str2double(cell2mat(answer(2,:)));
   %I need to cheat a little for perfomance/codesize reasons - force
    %square pixels.
    PERIODICITYX = ARRAYX * ELEMENT PERIODICITYX; %Calc size of pixel
in x
    PERIODICITYY = ARRAYY * ELEMENT PERIODICITYY; %Calc size of pixel
in y
    %Save Current Settings
    save pref(pref fname, answer);
    %Pregenerate some values for performance
   sPERIODICITYX = 1/UNIT * PERIODICITYX;
    sPERIODICITYY = 1/UNIT * PERIODICITYY;
   %Save Current Settings
    save pref(pref fname, answer);
else
   geom = 'circle';
   prompt = {'Element Periodicity (in X, um):', 'Element Periodicity
(in Y, um):', 'Number of Elements in a Pixel (in X):', 'Number of
Elements in a Pixel (in Y):', 'Diameter of array (in mm):'};
    defAns = open pref(pref fname, 5);
   while 1
        answer = inputdlq(prompt,dlq title2,num lines,defAns,'on');
        %Check for abort
        if isempty(answer)
            return;
        end
        %Convert all of the user input
        DIAMETER = str2double(cell2mat(answer(5,:)));
        ARRAYX = str2double(cell2mat(answer(3,:)));
        ARRAYY = str2double(cell2mat(answer(4,:)));
        ELEMENT PERIODICITYX = str2double(cell2mat(answer(1,:)));
        ELEMENT PERIODICITYY = str2double(cell2mat(answer(2,:)));
        %I need to cheat a little for perfomance/codesize reasons -
force
        %square pixels.
```

```
PERIODICITYX = ARRAYX * ELEMENT PERIODICITYX; %Calc size of
pixel in x
        PERIODICITYY = ARRAYY * ELEMENT PERIODICITYY; %Calc size of
pixel in y
        if PERIODICITYX ~= PERIODICITYY
            answer{3} = [num2str(ARRAYX), ' - Pixel is not a square!'];
        else
            PERIODICITY = PERIODICITYX;
           break;
        end
        defAns = answer;
    end
    %Save Current Settings
    save pref(pref fname, answer);
    %Pregenerate some values for performance
    DIAMETER = DIAMETER*1000;
    sPERIODICITYX = 1/UNIT * PERIODICITY;
    sPERIODICITYY = 1/UNIT * PERIODICITY;
end
```

This portion of the code asks the user which array shape they wish to create (square or circle) and then prompts for dimensions. It is important to note that the open\_pref/save\_pref function can be used to save the user's response, so that when the function is run a second time, the previous response will automatically be inputted. Also, it is important that the inputted sizes are scaled properly before proceeding (using UNIT, from before).

The function create\_gdsii\_file prompts the user where to save the final GDSII file. CELL\_NAME is the name of the of the final array cell and the same as the filename, without extension.

write\_header writes the header portion of the GDSII file. This also includes the time and the library (IR LAB CREOL MATLAB CUSTOM). The header is the first thing to write.

write\_units writes the units information to the GDSII file and must follow write\_header. write\_elements will write the baseline geometry to the GDSII file (specified with FILE\_ARRAY) and create an array based on the array size (ARRAYX, ARRAYY), element spacing (ELEMENT\_PERIODICITYX, ELEMENT\_PERIODICITYY), the starting points of the array (-sPERIODICITYX/2, -sPERIODICITYY/2), and the number of elements (1). If flag has been set to 451, as is the case when element\_gen has been called, the element geometry gdsii file is deleted. NAME\_ARRAY is the name(s) of the array (cells) created. This completes the element placement portion of the code for the square array.

```
%Write Structure
   outer radius = (DIAMETER/2)/PERIODICITY;
   %Search and place elements
   for y = 1:outer radius
        %Find edge positions and array sizes
       num x = floor(2*sqrt(outer radius^2-(y-0.5)^2));
       num x = num x + rem(num x, 2);
       px = -num x/2*sPERIODICITYX;
       py = y*sPERIODICITYY;
       %Write a single row
       write array(fid, NAME ARRAY{1}, num x, 1, sPERIODICITYX,
sPERIODICITYY, px, py-sPERIODICITYY);
       %Write the mirrored row
       write array(fid, NAME ARRAY{1}, num x, 1, sPERIODICITYX,
sPERIODICITYY, px, -py);
   end
   close gdsii structure(fid);
end
```

This portion of the code illustrates how more complicated layouts can be generated (in this case, a circle). To create a custom array layout, it is necessary to open the GDSII array (open\_gdsii\_structure), write the array (write\_array, with similar parameters as write\_elements), and close the array (close\_gdsii\_structure).

The last step is to close the file with the command close\_gdsii\_file. Further information, including the input and return values, of each function in the toolbox can be found by typing "help" followed by the function name in the console.

#### **B.4 Limitations of the GDSII Toolbox**

Several limitations are present in the software. Many of the mid-level functions, such as write\_array, write\_elements, etc., were written specifically for the original six modules. Lower level functions are either hard-coded or difficult to directly interface without using these middle functions. All of the functions are not suited for handling a GDSII stream spanning multiple files. While not impossible to alter, the code places all elements on layer 1 or on the layer of the original imported file. All of the modules lack command line modes and require a GUI. image\_pattern requires a grayscale image and will fail with any other type of image (RGB, indexed, etc.). Practically no error handling is present in the lower level code, except to prevent accidental data loss.

## LIST OF REFERENCES

- [1] C. Balanis, "Advanced Engineering Electromagnetics," Wiley, (1989).
- J. Maxwell, "A Dynamical Theory of the Electromagnetic Field," Phil Trans of the Royal Society of London 155, 459 – 512 (1865).
- [3] J. Goodman, "Statistical Optics," Wiley, (2000).
- [4] G. Fowles, "Introduction to Modern Optics," Wiley, (1975).
- [5] E. Hecht, "Optics," 4th ed., Wiley, (2002).
- [6] H. Hristov, "Fresnel Zones in Wireless Links, Zone Plate Lenses and Antennas," Artech, (2000).
- [7] O. Nairz, M. Arndt, and A. Zeilinger, "Quantum interference experiments with large molecules," Am. J Phys 71, 319 – 325 (2003).
- [8] C. Balanis, "Antenna Theory: Analysis and Design," 3rd ed., Wiley, (2005).
- B. Van Veen and K. Buckley, "Beamforming: A versatile approach to spatial filtering," IEEE ASSP Mag 5, 4 24 (1988).
- [10] E Bedrosian, "The curved passive reflector," IEEE Trans on AP 3, 168 173 (1955).
- [11] T. Pratt, C. Bostian, and J. Allnut, "Satellite Communications," Wiley, (2003).
- [12] J. Ojeda-Casaneda, C. Gomez-Reino, and B. J. Thomson (ed.), "Selected Papers on Zone Plates," SPIE, (1996).
- [13] J. Wiltse and J. Garret, "The Fresnel Zone Plate Antenna," Microwave J, 101 104 (1991).

- [14] J. Alda, J. Rico-García, J. López-Alonso, B. Lail, and G. Boreman, "Design of Fresnel lenses and binary-staircase kinoforms of low value of the aperture number," Opt Commun 260, 454 (2006).
- [15] B. Munk, "Frequency Selective Surfaces: Theory and Design," Wiley, (2000).
- [16] Y. Lee, J. Yeo, R. Mittra and W. Park, "Design of a High-Directivity Electromagnetic Band Gap (EBG) Resonator Antenna Using a Frequency Selective Surface (FSS) Superstrate", Microwave and Opt Tech Lett 43, 462 – 467 (2004).
- [17] J. B. Pendry, "Negative Refraction Makes a Perfect Lens," Phys Rev Lett 85, 3966 (2000).
- [18] N. Engheta, A. Alu, M. Silveirinha, A. Salandrino, and L. Jingjing, "DNG, SNG, ENZ and MNZ Metamaterials and Their Potential Applications," IEEE Mediterranean Electrotechnical Con, 258 – 261 (2006).
- [19] G. Goussetis, A. Feresidis, J. Vardaxoglou, "Tailoring the AMC and EBG characteristics of periodic metallic arrays printed on grounded dielectric substrate," IEE Trans on AP 54, 82 – 89 (2006).
- [20] B. Munk, "Finite antenna arrays and FSS," Wiley, (2003).
- [21] R. Chu and K. Lee, "Analytical Model of a Multilayered Meander-Line Polarizer Plate with Normal and Oblique Plane-Wave Incidence," IEEE Trans on AP 35, 652 – 661 (1987).
- [22] S. Seshadri, "Polarization control by a twist reflector," J Appl Phys 70, 1952 (1991).

- [23] K. Leong and W. Shiroma "Waffle-grid polarizer," Elec Lett 38, 1360 1361 (2002).
- [24] L. Young, L. Robinson, and C. Hacking, "Meander-Line Polarizer," IEEE Trans on AP 21, 376 – 378 (1973).
- [25] J. Tharp, "Design and demonstration of meanderline retarders at infrared frequencies," PhD dissertation, University of Central Florida, (2007).
- [26] J. Tharp, B. Lail, B. Munk, and G. Boreman, "Design and Demonstration of an Infrared Meanderline Phase Retarder," IEEE Trans on AP 55, 2983 – 2989 (2007).
- [27] J. Tharp, J. López-Alonso, J. Ginn, C. Middleton, B. Lail, B. Munk, and G. Boreman, "Demonstration of a single-layer meanderline phase retarder at infrared," Opt Lett 31, 2687 2689, (2006).
- [28] J. Tharp, J. Alda, and G. Boreman, "Off-axis behavior of an infrared meanderline waveplate," Opt Lett 32, 2852 – 2854 (2007).
- [29] C. Pell, "Phased-array radars," IEE Review **34**, 363 367 (1988).
- [30] H. Schippers, G. Spalluto, and G. Vos, "Radiation analysis of conformal phased array antennas on distorted structures," IEEE AP Soc Int Symp, 160 163 (2003).
- [31] C. Middlebrook, P. Krenz, B. Lail, and G. Boreman, "Infrared Phased-Array Antenna," Microwave and Opt Tech Lett 50, 719 – 723 (2008).
- [32] C. Middlebrook, "Infrared Antenna-Coupled Phased-Array," PhD dissertation, University of Central Florida (2007).
- [33] D. B. Rutledge, S. Schwarz, and A. Adams, "Infrared and Submillimeter Antennas," IR Phys 18, 713 – 729 (1978).

- [34] J. Bean, B. Tiwari, G. H. Bernstein, P. Fay, and W. Porod, "Thermal Infrared Detection Using Dipole Antenna-Coupled Metal-Oxide-Metal Diodes," JVSTB 27, 11 – 14 (2009).
- [35] H. Kazemi, K. Shinohara, G. Nagy, W. Ha, B. Lail, E. Grossman, G. Zummo, W. Folks, J. Alda, and G. Boreman, "First THz and IR characterization of nanometer-scaled antenna-coupled InGaAs/InP Schottky-diode detectors for room temperature infrared imaging," Proc. SPIE 6542, (2007).
- [36] W. Yang, M. Roggemann, K. Cooper, W. Buller, N. Subotic, C. Middlebrook and G. Boreman, "The IR antenna pair coupled sensor element and its potential application in wavefront sensing," IR Phys and Tech 51, 495 504 (2008).
- [37] R. Olmon, P. Krenz, A. Jones, G. Boreman, and M. Raschke, "Near-field imaging of optical antenna modes in the mid-infrared," Opt Exp 16, 20295 20305 (2008).
- [38] K. Shum, Q. Xue, C. Chan, K. Luk, "Investigation of microstrip reflectarray using a photonic bandgap structure," Microwave and Opt Tech Lett 18, 114 116 (2001).
- [39] J. Huang, "Capabilities of printed reflectarray antennas," IEEE Int Sym on Phased Array Systems and Tech, 131 – 134 (1996).
- [40] D. Berry, R. Malech, and W. Kennedy, "The reflectarray antenna," IEEE Trans. on AP 11, 645 – 651 (1963).
- [41] J. Montgomery, "Scattering by an infinite periodic array of microstrip elements," IEEE Trans. on AP 26, 850 – 854 (1978).

- [42] N. Misran, R. Cahill, and V. Fusco, "RCS reduction technique for reflectarray antennas," Elec Lett 39, 1630 – 1631 (2003).
- [43] D. Pozar, and T. Metzler, "Analysis of a reflectarray antenna using microstrip patches of variable size," Elec Lett 29, 657 – 658 (1993).
- [44] A. Zaman, and R. Lee, "A spherical to plane wave transformation using a reflectarray," IEEE Ant. and Prop Soc Int Symp, 1284 – 1287 (1997).
- [45] A. Kelkar, "FLAPS conformal phased reflecting surfaces," Proc of the 1991 IEEE
   Nat Radar Con, 58 62 (1991).
- [46] D. Gonzalez, G. Pollon, and J. Walker, "Microwave phasing structures for electromagnetically emulating reflective surfaces and focusing elements of selected geometry," US Patent 4905014, Malibu Research Associates, (1990).
- [47] From ILC Dover corporate website: http://www.ilcdover.com/products/aerospace\_defense/inflatableantenna.htm
- [48] From TRLabs corporate website:

http://www.trlabs.ca/trlabs/technology/technologybulletins/reconfigurableantenna .html

- [49] J. Ginn, "Demonstrating Reflectarray Behavior at Infrared," Masters Thesis, University of Central Florida, (2006).
- [50] J. Ginn, D. Shelton, P. Krenz, B. Lail, and G. Boreman, "Altering Infrared Metamaterial Performance Through Metal Resonance Damping," J of App Phys 105, 074304 (2009).

- [51] J. Ginn, B. Lail, D. Shelton, J. Tharp, W. Folks, and G. Boreman, "Characterizing Infrared Frequency Selective Surfaces on Dispersive Media," ACES J 22, 184 – 188, (2007).
- [52] J. Ginn, B. Lail, and G. Boreman, "Infrared Patch Reflectarray," IEEE AP Soc Int Symp, 4315 – 4318 (2006).
- [53] J. Ginn, B. Lail, and G. Boreman, "Phase Characterization of Reflectarray Elements at Infrared," IEEE Trans. on AP 55, 2989 – 2993 (2007).
- [54] J. Ginn, B. Lail, J. Alda, and G. Boreman, "Planar infrared binary phase reflectarray," Opt Lett 33, 779 – 781 (2008).
- [55] J. Ginn, D. Shelton, J. Alda, B. Lail, and G. Boreman, "Microstrip Reflectarray Design at MWIR and NIR," IEEE AP Soc Int. Symp, 1 – 4 (2008).
- [56] J. Ginn, B. Lail, and G. Boreman, "Sub-Millimeter and Infrared Reflectarray," US Patent 20070132645, under review, (2006).
- [57] F. Wooten, "Optical Properties of Solids," Academic Press, (1972).
- [58] D. Pozar, S. Targonski, and H. Syrigos, "Design of Millimeter Wave Microstrip Reflectarrays," IEEE Trans. on AP 45, 287 – 296 (1997).
- [59] Ansoft HFSS reference manual, Rel. 10.1, Ansoft Corporation, (2006).
- [60] Ansoft Designer reference manual, Rel. 4.0, Ansoft Corporation, (2008).
- [61] N. Lenin, and P. Rao, "Evaluation of the reflected phase of a patch using waveguide simulator for reflectarray design," Microwave and Opt Tech Lett 45, 528 531 (2005).

- [62] I. Bardi, M. Vogel, and Z. Cendes, "Modeling Large Screens via Homogenization with the Finite Element Method," IEE MTT-S Int Micowave Symp, 1315 1318 (2006).
- [63] "Cyclotene 3000 Series Advanced Electronic Resins: Processing Procedures for CYCLOTENE 3000 Series Dry Etch Resin," Dow Corporation, (2005).
- [64] P. Rai-Choudhury, "Handbook of Microlithography, Micromachining, and Microfabrication," vol. 1, SPIE (1997).
- [65] M. Fox, "Optical Properties of Solids," Oxford, (2003).
- [66] J. Raynolds, B. Munk, J. Pryor, and R. Marhefka, "Ohmic loss in frequencyselective surfaces," J of App Phys 95, 5346 – 5358 (2003).
- [67] E. Topsakal, J. Volakis, "On the properties of materials for designing filters at optical frequencies" IEEE AP Soc Int Symp, 635 638 (2003).
- [68] L. Novotny, "Effective Wavelength Scaling for Optical Antennas," PRL 98, 266802 (2007).
- [69] I. Pusçasu, W. Schaich, and G. Boreman, "Modeling parameters for the spectral behavior of infrared frequency-selective surfaces," App Opt 40, 118 124 (2001).
- [70] C. Fumeaux, M. Gritz, I. Codreanu, W. Schaich, F. González, and G. Boreman,
   "Measurement of the resonant lengths of infrared dipole antennas," Inf Phys and Tech 41, 271-281 (2000).
- [71] A. Vengurlekar and T. Ishihara, "Surface Plasmon enhanced photon drag in metal films," Appl Phys Lett 87, 091118 (2005).

- [72] D. Shelton, T. Sun, J. Ginn, K. Coffey, and G. Boreman, "Relaxation time effects on dynamic conductivity of alloyed metallic thin films in the infrared band," J of App Phys 104, 103514 (2008).
- [73] F. Tsai, and M. Bialkowski, "Designing a 161-element Ku-band microstrip reflectarray of variable size patches using an equivalent unit cell waveguide approach," IEEE Trans on AP 51, 2953 – 2962 (2003).
- [74] D. Chang, and M. Huang, "Multiple-polarization microstrip reflectarray antenna with high efficiency and low cross-polarization" IEEE Trans on AP 43, 829 834 (1995).
- [75] F. Venneri, G. Angiulli, and G. Di Massa, "Design of microstrip reflect array using data from isolated patch analysis," Microwave and Opt Tech Lett 32, 411-414 (2002).
- [76] D. Pozar, and S. Targonski, "A microstrip reflectarray using crossed dipoles," IEEE AP Soc Int Symp, 1008 – 1011 (1998).
- [77] Z. Wu,W. Zhang, Z. Liu, and W. Shen, "Circularly polarised reflectarray with linearly polarised feed" Elect. Lett **41**, 387 388 (2005).
- [78] D. Pozar, "Bandwidth of reflectarrays," Elect Lett **39**, 1490 1491 (2003).
- [79] D. Pozar, "Microwave Engineering," Wiley, 3 ed., (2005).
- [80] J. Huang and J. Encinar "Reflectarray Antennas," Wiley, (2007).
- [81] M. Mussetta, N. Bliznyuk, P. Pirinoli, N. Engheta, and R. Zich, "Application of genetic algorithms for optimization of a FSS reflectarray antenna," Int Conference on Math Methods in EM Theory, 522 – 524 (2004).

- [82] M. Chaharmir, J. Shaker, M. Cuhaci, and A. Sebak, "Novel photonicallycontrolled reflectarray antenna," IEEE Trans on AP 54, 1134 – 1141 (2006).
- [83] D. Shelton, J. Ginn, K. Coffey, and G. Boreman, "Infrared photoconductivity in heavily nitrogen doped a-Si:H," MRS Symp Proc 1153, 11153-A02-02 (2009).
- [84] J. Encinar, "Design of two-layer printed reflectarrays using patches of variable size," IEEE Trans on AP 49, 1403 – 1410 (2001).
- [85] D. Wu, R. Hall, J. Huang, "Dual-frequency microstrip reflectarray," IEEE AP Soc Int Symp, 2128 – 2131 (1995).
- [86] J. Zornoza, and M. Bialkowski, "Design of a shaped beam multi-layer microstrip reflectarray with Australia and New Zealand coverage pattern," IEEE Int Symp on Phased Array Sys and Tech, 488 – 493 (2003).
- [87] K. Dey and P. Khastigir, "Comparative Focusing Properties of Spherical and Plane Microwave Zone Plate Antennas," Int J Electronics 35, 497 - 506 (1973).
- [88] M. Milon, R. Gillard, D. Cadoret, and H. Legay, "Analysis of mutual coupling for the simulation of reflectarrays radiating cells," EuCAP, (2006).
- [89] J. Volakis, ed., "Antenna Engineering Handbook," McGraw Hill, (2007).
- [90] M. Young, "Zone Plates and Their Aberrations," J Opt Soc Am 62, 972 976 (1972).
- [91] J. Alda, "Chromatic Aberrations in Reflectarrays," personal correspondence to the author, (2008).
- [92] D. McGrath, "Planar Three-Dimensional Constrained Lenses," IEEE Trans on AP
   34, 46 50 (1986).
- [93] F. Tsai and Marek Bialkowski, "Bandwidth enhancement of a transmitarray-type spatial power combiner with the use of stacked patches," Intl Con On Microwave, Radar, and Wireless Commun, 310 – 313 (2004).
- [94] Z. Popović and A. Mortazawi, "Quasi-Optical Transmit/Receive Front Ends," IEEE Trans on MTT 46, 1964 – 1975.
- [95] M. Bialkowski and H. Song, "A Ku-Band Active Transmit-Array Module with a Horn or Patch Array as a Signal Launching/Receiving Device," IEEE Trans. on AP 49, 535 – 541 (2001).
- [96] M. Bialkowski and H. Song, "Investigations into a Power-Combining Structure Formed by Trays of Uniplanar Quasi-Yagi Antennas," Microwave and Opt Tech Lett 27, 50 – 53 (2000)
- [97] P. Padilla de la Torre and M. Sierra-Castañer, "Design and Prototype of a 12-GHz Transmit-array," Microwave and Opt Tech Lett 49, 3020 – 3026 (2007).
- [98] M. Chaharmir, S. Raut, A. Ittipiboon, and A. Petosa "Cylindrical Multilayer Transmitarray Antennas," Int URSI Commun B EM Theory Symp, (2007).
- [99] K. Lam, S. Kwok, Y. Hwang, and T. Lo, "Implementation of Transmitarray Antenna Concept by Using Aperture-Coupled Microstrip Patches," Asia Pacific Microwave Con, 433 – 436 (1997).
- [100] W. Mayer, Stefan Buntz, H. Leier, and W. Menzel, "Imaging Radar Sensor Front-End with a Large Transmit Array," Euro Radar Con, 153 – 156 (2004).
- [101] L. Angelone, N. Makris, C. Vasios, L. Wald, and G. Bonmassar, "Effect of transmit array phase relationship on local Specific Absorption Rate (SAR)," Proc of Intl Soc Mag Reson Med 14, 2038 (2006).

- [102] P. Clapham and M. Hutley, "Reduction of Lens Reflexion by the Moth Eye Principle," Nature 233, 281 – 282 (1973).
- [103] B. Lee and Z. Zhang, "Coherent Thermal Emission From Modified Periodic Multilayer Structures," J Heat Trans 129, 17-26 (2007).
- [104] J. Greffet, R. Carminati, K. Joulain, J. Moulet, S. Mainguy and Y. Chen,"Coherent emission of thermal sources", Nature 416, 61-64 (2002).
- [105] E. Hasman, N. Dahan, A. Niv, G. Biener, and V. Kleiner, "Space-Variant Polarization Manipulation of a Thermal Emission by a SiO2 Subwavelength Grating Supporting Surface Phonon-Polariton," Con on Lasers and Electro-Optics, CTuL6 (2005).
- [106] L. Klein, H. Hamann, Y. Au, and S. Ingvarsson, "Coherence properties of infrared thermal emission from heated metallic nanowires", App Phy Lett 92, 213102 (2008).
- [107] S. Ingvarsson, L. Klein, Y. Au1, J. Lacey, and H. Hamann, "Enhanced thermal emission from individual antenna-like nanoheaters," Opt Exp 15, 11249-11254 (2007).
- [108] C. Wang, Y. Chang, M. Tsai, Y. Ye, C. Chen, Y. Jiang, Y. Chang, S. Lee, and D. Tsai, "Reflection and emission properties of an infrared emitter," Opt. Exp 15, 14673-14678 (2007).
- [109] B. Monacelli, J. Pryor, B. A. Munk, D. Kotter, and G. Boreman, "Infrared frequency selective surface based on circuit-analog square loop design," IEEE Trans on AP 53, 745 – 752 (2005).

- [110] I. Pusçasu, W. Schaich, and G. Boreman, "Resonant enhancement of emission and absorption using frequency selective surfaces in the infrared," IR Phys and Tech 43, 101 – 107 (2002).
- [111] H. Ruthanne, E. Swaim, N. Hammer, E. Richards, D. Venkataraman, and M.
  Barnes, "Robust Circular Polarized Emission from Nanoscopic Single-Molecule Sources: Application to Solid State Devices," MRS Proceedings, 192 – 197 (2007).
- [112] J. Tharp, D. Shelton, S. Wadsworth, and G. Boreman, "Electron-beam lithography of multiple-layer submicrometer periodic arrays on a barium fluoride substrate,"
  JVSTB 26, 1821 1823 (2008).
- [113] J. Vardaxoglou and E. Parker, "Performance of two tripole arrays as frequency selective surfaces," Elect Lett 19, 709 – 710 (1983).
- [114] Personal discussion with Mr. Prashant Nagpal at MRS 2009 Spring conference.
- [115] F.J. González, J. Alda, B. Ilic, and G.D. Boreman, "Infrared antennas coupled to lithographic Fresnel zone plate lenses," Apl Opt 43, 6067 – 6073 (2004).
- [116] J. Hilfiker, "Introduction to Ellipsometry," Presentation by J.A. Woollam Co. Inc., (2005).
- [117] "Guide to Using WVASE32," J.A. Woollam Co. Inc., (2007).
- [118] T. Sridhar, D. Nguyen, and G. Fuller, "Birefringence and stress growth in uniaxial extension of polymer solutions," J Non-Newtonian Fluid Mech 90, 299 315 (2000).
- [119] D. Malacara, "Optical Shop Testing," Wiley, 3 ed., (1992).

## [120] Image taken from:

http://microgravity.grc.nasa.gov/MSD/GRC\_MSD\_Graphics/FluidPhysics/78457

TwymanGreenInterferome.pdf