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Design of a Trapezoidal Patch Antenna for Wearable Applications

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Abstract

Wearable body centric wireless communication technologies require antennas that can operate when being subjected to bending due to conformal placement and movement of the human body.

A trapezoidal patch antenna (TPA) has been found to be less susceptible to bending effects than a traditional rectangular patch antenna (RPA). The antenna's suitability for wearable off-body communication links operating at the 2.4 GHz ISM band has been investigated. The TPA is modelled for the first time using the Aperture Array Model and the Cylindrical Aperture Array Model which incorporates antenna E-plane bending. The analysis is taken further with full-wave simulations and shows that the TPA is less susceptible to bending effects, particularly E-plane bending compared to a RPA. However, there is a radiation performance degradation when compared to the RPA. To overcome this, the TPA is further improved by implementing an artificial magnetic conducting (AMC) metafilm array configuration for an edge fed antenna. The trapezoidal patch antenna loaded with the AMC metafilm (TPA-MF) leads to enhanced radiation performance with a realized gain of up to 6.56 dBi and a radiating element surface area reduction of 48.9% when compared to a RPA.

Contents

Abstract	i
Contents.....	ii
Abbreviations	v
List of Figures.....	vi
List of Tables	xv
1 Introduction.....	1
1.1 Project Motivation	1
1.2 Project Challenges and Aims.....	3
1.2.1 Key Contributions	5
1.3 Project Approach.....	5
1.4 Thesis Overview	7
1.5 References	9
2 Fundamentals.....	12
2.1 Fundamental Antenna Parameters	12
2.1.1 Reflection Coefficient and VSWR.....	12
2.1.2 Bandwidth	14
2.1.3 Efficiency.....	14
2.1.4 Radiation Pattern Parameters	15
2.2 Full-Wave Simulations.....	21
2.2.1 Excitation	21
2.2.2 Solver Options	21
2.2.3 Time Domain Convergence Parameters and Mesh Refinement.....	22
2.2.4 Frequency Domain Mesh Refinement	25
2.2.5 Solver and Mesh Structure Selection.....	26
2.2.6 Frequency Optimisation.....	28
2.3 References	33
3 Historical Overview	34
3.1 Microstrip Patch Antennas	34
3.2 Wearable Microstrip Patch Antennas	38
3.3 References	42
4 Patch Antenna Miniaturization.....	48
4.1 Developments in Antenna Miniaturization.....	49
4.2 Proposed Inset-Fed Trapezoidal Patch Antenna	56
4.3 Aperture Array Model.....	57

4.3.1	Theory and Analysis of an Aperture Antenna	58
4.3.2	Aperture Array Model for the Trapezoidal Patch Antenna	64
4.4	Full Wave Simulation Analysis	81
4.5	Conclusion	90
4.6	References	91
5	Investigating Patch Antenna Bending Effects	94
5.1	Literature	94
5.2	Aperture Array Model for Patch Antenna Bending	108
5.3	Modelling Resonant Frequency Detuning	115
5.4	Full-Wave Simulation Analysis of Patch Antenna Bending	118
5.5	References	139
6	Optimisation Techniques using an Artificial Magnetic Conductor (AMC) Metafilm	140
6.1	Background of an Artificial Magnetic Conductor Metafilm	140
6.2	Design of Proposed AMC Metafilm	151
6.3	Effective Parameters of the Proposed AMC Metafilm	156
6.4	Integrating the AMC Metafilm in The Trapezoidal Patch Antenna	164
6.5	Antenna Bending Analysis	169
6.6	Conclusion	183
6.7	References	185
7	Simulation of the Wearable Trapezoidal Patch Antenna on the Human Body	188
7.1	Dielectric Properties of Human Body Tissue	189
7.2	Specific Absorption Rate (SAR) and Recommended Human Exposure Limits	190
7.3	Effect of Body Proximity on Antenna Performance and SAR	194
7.4	Conclusion	203
7.5	References	205
8	Conclusions	207
8.1	Summary	207
8.1.1	Chapter 4	207
8.1.2	Chapter 5	207
8.1.3	Chapter 6	208
8.1.4	Chapter 7	209
8.2	Future Research Directions	210
8.2.1	Fabrication and Measurements	210
8.2.2	Analytical Analysis	210
8.2.3	Further Miniaturization	211
8.2.4	Applying Further Bandwidth Enhancement Techniques	211

8.2.5	Further Simulation of the Proposed Antennas on the Human Body.....	211
8.2.6	Method for Implementing Circular Polarization	211
8.3	Reflection.....	211
8.4	References	214
Appendix.....		215
A.	MATLAB Codes	215
A.1	Dual-AAM to compute far-field patterns for the TPA.....	215
A.2	Dual and Quad-AAM to compute directivity for the TPA	218
A.3	Quad-AAM to compute far-field pattern for the TPA	222
A.4	Cylindrical-AAM to compute directivity for the TPA.....	228
A.5	Cylindrical-AAM to compute far-field patterns for the TPA	232
A.6	Cylindrical-Rectangular Cavity Model to Compute Resonant Frequencies.....	235
B.	Measurements.....	238
B.1	Measured return loss for the fabricated RPA and TPA	238
B.2	Implementing Antenna E-plane Bending.....	239
C.	Simulation Results	242
D.	Human Tissue Material Definitions	244
D.1	Cubic and Cylindrical Tissue Model	244
D.2	Homogenous Phantom Model	246
D.3	FEM Human Model.....	247

Abbreviations

AAM	Aperture Array Model
AMC	Artificial Magnetic Conductor
CAAM	Cylindrical Aperture Array Model
DRL	Dosimetric Reference Level
EBG	Electromagnetic Band Gap
ERL	Exposure Reference Level
FDTD	Finite Difference Time Domain
FEM	Finite-Element Method
FSS	Frequency Selective Surface
GSTC	Generalized Sheet Transition Conditions
GTD	Geometrical Theory of Diffraction
HIS	High Impedance Surface
ICNIRP	The International Commission on Non-Ionizing Radiation Protection
IEEE ICES	Institute of Electrical and Electronic Engineers International Committee on Electromagnetic Safety
ISM	Industrial Scientific Medical
NanoVNA	Nano Vector Network Analyser
PEC	Perfect Electric Conductor
PMC	Perfect Magnetic Conductor
RAM	Random Access Memory
RPA	Rectangular Patch Antenna
TE	Transverse Electric
TM	Transverse Magnetic
TPA	Trapezoidal Patch Antenna
TPA-MF	Trapezoidal Patch Antenna with an AMC Metafilm
WHO	World Health Organization

List of Figures

Figure 1.1: A Jacket integrated with a circularly polarized patch antenna on the arm as part of a GPS receiver [10].	2
Figure 1.2: lifejacket equipped with patch antennas as part of a COSPAS-SARSAT search and rescue transmitter [11].	2
Figure 1.3: Six-element wearable microstrip patch antenna array for RF power harvesting [8].	2
Figure 2.1: Scattering parameters for a two-port network.	12
Figure 2.2: A comparison between (a) reflection coefficient and (b) VSWR analysis for antennas [2].	13
Figure 2.3: Field regions of an antenna [3].	15
Figure 2.4: Changes of an antennas amplitude pattern shape from the reactive near-field towards the far-field [6].	16
Figure 2.5: Spherical coordinate system geometry for antenna far-field analysis [8].	18
Figure 2.6: 2D normalized radiation power plots for a microstrip patch antenna for the (a) E-plane and (b) H-plane. The antennas geometry, radiating edges modelled as slots and corresponding planes are given in (c) [9].	19
Figure 2.7: Polar radiation power pattern plot showing the various radiation lobes and beamwidth [3].	20
Figure 2.8: Linear radiation power pattern plot showing the various radiation lobes and beamwidth [3].	20
Figure 2.9: Mesh refinement convergence criteria delta using the time-domain solver with a hexahedral mesh.	24
Figure 2.10: Mesh refinement variation in the S-parameters after each subsequent mesh pass.	24
Figure 2.11: Increase in the overall number of mesh cells after each subsequent mesh pass.	25
Figure 2.12: Rectangular patch antenna designed in CST Microwave Studio.	26
Figure 2.13: Tetrahedral mesh cross-section of a rectangular patch antenna along the (a) ZY plane (b) ZX plane and (c) YX plane.	27
Figure 2.14: Hexahedral mesh cross-section of a rectangular patch antenna along the (a) ZY plane (b) ZX plane and (c) YX plane.	27
Figure 2.15: Front view geometry of the rectangular patch antenna.	29
Figure 2.16: Simulated S-parameter result of the rectangular patch antenna using Trust Region Framework optimizer.	30
Figure 2.17: Simulation iterations of the rectangular patch antenna using the Trust Region Framework optimizer which is converging closer to the defined goal.	30

Figure 2.18: Optimization of the antenna's (a) vertical length and (b) inset feed gap width.	31
Figure 3.1: Cross sectional schematic of a satin weave fabric microstrip resonator for when (a) the conductive threads are placed underneath and (b) on top of the nonconductive threads [62].....	39
Figure 4.1: Geometry of the Planar Inverted-F Antenna (PIFA) [8].	50
Figure 4.2: Schematic of an edge-fed shorted patch antenna where the dimensions for a resonance at 530 MHz given are $L = 44.5$ mm, $W = 48.75$ mm, $L_f = 16$ mm, $W_f = 1$ mm, $L_{fi} = 11.5$ mm, $W_{fi} = 7$ mm, $L_1 = 1$ mm, $W_1 = 3$ mm, $L_2 = 2$ mm, $W_2 = 3$ mm, $x_{ps} = 0.5$ mm, $r_{0s} = 1$ mm [9].	51
Figure 4.3: Schematic of the radiating element of two types of printed rectangular spiral patch antennas where (a) has a resonance of 1.9 GHz (b) has a resonance of 715 MHz and (c) 440 MHz all with dimensions $\epsilon_r = 1.07$, $d = 10$ mm, $L = 17$ mm, $W = 25$ mm, $wt = 1.2$ mm, $wg = 1$ mm, feed probe radius = 0.6 mm, shorting pin radius = 0.325 mm [9].	52
Figure 4.4: Miniaturized patch antenna with CSRR loaded into substrate [6].	53
Figure 4.5: Trapezoidal Patch Antenna Geometry.	56
Figure 4.6: Charge distribution and current density formed on the microstrip patch antenna [28]. ..	57
Figure 4.7: The fields, current densities, and material properties for the (a) actual and (b) equivalent aperture model [28].	58
Figure 1.8: Love's equivalent model [28].	59
Figure 4.9: Far-field equivalent diagram for an aperture.	62
Figure 4.10: Equivalent models for a magnetic source near a ground plane or a perfect electric conductor [31].	63
Figure 4.11: Dual aperture array geometry and far-field approximations.	66
Figure 4.12: E-plane normalized linear far-field plot in terms of the radiation intensity U of a RPA where $W = W_1$ and a TPA where $W_2 = 5$ mm.	69
Figure 4.13: H-plane normalized linear far-field plot in terms of the radiation intensity U of a RPA where $W = W_1$ and a TPA where $W_2 = 5$ mm.	70
Figure 4.14: Parametric analysis modelling the patch antenna as a dual-aperture array by investigating the change in maximum directivity at $\theta = 90^\circ$ by varying the width W_2 of aperture #2.	71
Figure 4.15: Rectangular patch antenna geometry showing the electric fringe fields at each aperture.	72
Figure 4.16: Non-radiating edges along the length of the rectangular patch antenna.	72
Figure 4.17: Trapezoidal patch antenna geometry showing the electric fringe fields at each aperture.	73

Figure 4.18: Aperture Array Model analysis and far-field approximation for Aperture #3 and #4 with lengths of L_d .	74
Figure 4.19: Front view trapezoidal patch geometry modelled as a quad-aperture array.	76
Figure 4.20: E-plane far-field plot in terms of the normalized radiation intensity U comparing the Dual and Quad-AAM for a RPA and TPA.	78
Figure 4.21: H-plane far-field plot in terms of the normalized radiation intensity U comparing the Dual and Quad-AAM for a RPA and TPA.	78
Figure 4.22: Comparison between the Dual and Quad-AAM analysis calculating the directivity of the patch antenna varying the aperture width W_2 .	79
Figure 4.23: Front view geometry of the Trapezoidal Patch Antenna.	82
Figure 4.24: Comparison of simulated reflection coefficients for the TPA and RPA using the (a) frequency domain solver and (b) time domain solver.	84
Figure 4.25: Comparison between the time domain and frequency domain solver in CST for (a) bandwidth and (b) realized gain while varying the aperture width W_2 .	85
Figure 4.26: Change in the antenna's radiating element surface area while varying the aperture width W_2 and maintaining a resonant frequency of 2.4 GHz.	86
Figure 4.27: E-plane radiation pattern cut comparison for the (a) RPA and (b) TPA between the analytical Quad-AAM and full-wave simulations using the time and frequency domain solver.	87
Figure 4.28: H-plane radiation pattern cut comparison for the (a) RPA and (b) TPA between the analytical Quad-AAM and full-wave simulations using the time and frequency domain solver.	88
Figure 4.29: Comparison between the proposed analytical models and the simulated analysis calculating the directivity of the antenna varying the aperture width W_2 .	89
Figure 5.1: Perspective view of a rectangular patch antenna modelled as a cylindrical-rectangular cavity [2].	95
Figure 5.2: Cross-sectional view of (a) the cylindrical-rectangular cavity model and (b) the isolated sector that forms the patch antenna showing the associated boundary conditions [2].	95
Figure 5.3: Geometry of the cylindrical-rectangular patch antenna [3].	96
Figure 5.4: Normalized radiation pattern comparison between the theoretical analysis (solid curves) and measured results (broken curves) for TM ₀₁ mode [3].	98
Figure 5.5: Normalized power pattern plot showing the E and H plane for the TM ₁₀ mode with $\epsilon = 1.06$ [4].	99
Figure 5.6: Normalized power pattern plot showing the E and H plane for the TM ₀₁ mode with $\epsilon = 1.06$ [4].	99

Figure 5.7: Percent change of the resonant frequency while varying the bending radius along the textile patch antenna's length [7].	101
Figure 5.8: Cylindrical-rectangular cavity model showing the (a) geometry of the patch antenna bent on a cylindrical surface and the bending configuration for (b) E-plane bending and (c) H-plane bending [5].	102
Figure 5.9: Normalized frequency shift plotted against antenna bending angle using the cylindrical-rectangular cavity model, HFSS port driven full wave simulation and HFSS eigenmode simulation for (a) E-plane bending and (b) H-plane bending [5].	103
Figure 5.10: Measured normalized frequency shift of the patch antenna prototype compared with the simulated results for E and H-plane bending [5].	103
Figure 5.11: Microstrip patch antenna mounted on a cylinder to model antenna bending [8].	104
Figure 5.12: Radiation pattern displaying the E-plane and H-plane for a microstrip patch antenna on a cylinder.	106
Figure 5.13: Radiation pattern displaying the E-plane and H-plane for a planar microstrip patch antenna.	106
Figure 5.14: Model representing antenna bending as a sector of a cylinder where (a) represents a sector with a bending angle $\Phi < 180^\circ$, (b) $\Phi = 180^\circ$ and (c) $\Phi > 180^\circ$.	109
Figure 5.15: Aperture array model for the TPA on a cylinder.	110
Figure 5.16: E-plane far-field radiation intensity pattern for the TPA bent over a conductive cylinder for $0^\circ \leq \phi \leq 360^\circ$.	113
Figure 5.17: H-plane far-field radiation intensity pattern for the TPA bent over a conductive cylinder for $0^\circ \leq \theta \leq 360^\circ$.	113
Figure 5.18: Computed directivity using the cylindrical aperture array model for various bending angles.	114
Figure 5.19: Resonant frequency shift for increasing bending angles using the cylindrical-rectangular cavity model for H-plane and E-plane bending.	116
Figure 5.20: RPA cylindrically bent along its (a) width, YZ plane or E-plane and (b) length, XZ plane or H-plane.	119
Figure 5.21: Simulated input reflection coefficient for various E-plane bending angles Φ of the RPA.	119
Figure 5.22: Change in the RPA's resonance frequency while bending along the YZ plane or E-plane.	120
Figure 5.23: Change in the RPA's bandwidth while bending along the YZ-plane or E-plane.	120

Figure 5.24: Simulated radiation pattern for (a) X-Z or H-plane and (b) Y-Z or E-plane while bending the antenna across its length or E-Plane.	121
Figure 5.25: Change in realized gain while bending along the YZ plane or E-plane.	122
Figure 5.26: Change in total radiation efficiency while bending along the YZ plane or E-plane.	123
Figure 5.27: Simulated input reflection coefficient for various XZ Plane Bending angles Φ of the RPA.	124
Figure 5.28: Change in resonance frequency while bending the RPA along the XZ plane or H-plane.	124
Figure 5.29: Change in bandwidth while bending the RPA along the XZ plane or H-plane.	125
Figure 5.30: Simulated radiation pattern in terms of realized gain for the (a) X-Z plane and (b) Y-Z plane while bending the antenna across its width or H-Plane.	126
Figure 5.31: Change in realized gain while bending the RPA along the XZ plane or H-plane.	127
Figure 5.32: Change in total radiation efficiency while bending the RPA along the XZ plane or H-plane.	127
Figure 5.33: Trapezoidal Patch Antenna Cylindrically Bent Along Its (a) Width, YZ plane or E-plane and (b) Length, XZ plane or H-plane.	128
Figure 5.34: Simulated input reflection coefficient for E-plane bending of the TPA.	129
Figure 5.35: Change in the resonance frequency while bending the TPA along the YZ-plane or E-plane.	129
Figure 5.36: Change in the bandwidth while bending the TPA along the YZ-plane or E-plane.	130
Figure 5.37: Simulated radiation pattern for (a) X-Z plane, (b) Y-Z plane and (c) X-Y plane while bending the antenna across its length or E-Plane.	131
Figure 5.38: Change in realized gain while bending the TPA along the YZ plane or E-plane.	132
Figure 5.39: Change in radiation efficiency while bending the TPA along the YZ plane or E-plane.	132
Figure 5.40: Simulated input reflection coefficient for H-plane bending of the Trapezoidal Patch Antenna.	133
Figure 5.41: Change in resonance frequency of the TPA while bending along the XZ plane or H-plane.	134
Figure 5.42: Change in the bandwidth of the TPA while bending along the XZ plane or H-plane.	134
Figure 5.43: Simulated radiation pattern for (a) X-Z plane and (b) Y-Z plane while bending the antenna across its width or H-Plane.	135
Figure 5.44: Change in realized gain while bending the TPA along the XZ-plane or H-plane.	136
Figure 5.45: Change in radiation efficiency while bending the TPA along the XZ-plane or H-plane.	137

Figure 5.46: Comparison between the analytical and simulated directivity of the TPA during bending.	137
Figure 6.1: Radiating element close placed at a distance (a) less than $\lambda/4$, (b) approximate $\lambda/4$ from ground plane [1].	140
Figure 6.2: Standing wave fields at a frequency (a) far below resonance, (b) Near or below the resonance, (c) at the resonance and (d) above the resonance [1].	141
Figure 6.3: Interference of the space waves and surface waves at the edge of the ground plane. ..	142
Figure 6.4: Hexagonal high impedance surface (a) perspective view and (b) cross sectional side view [3].	142
Figure 6.5: Unit cell of the hexagonal high impedance surface (a) cross-sectional view and equivalent circuit [3].	143
Figure 6.6: E-plane radiation pattern of a patch antenna consisting of an ordinary ground plane and a high-impedance surface as a ground plane [1].	144
Figure 6.7: Dipole antenna mounted on a high impedance ground plane [1].	144
Figure 6.8: The (a) geometry of a high-impedance or electromagnetic band gap surface. The simulated (b) dispersion properties showing the band gap for surface waves and (c) reflection phase properties [4].	145
Figure 6.9: Illustration of a metafilm with a periodic arrangement of an arbitrarily shaped scatterer in the xy-plane [4].	146
Figure 6.10: Metafilm with (a) a periodic arrangement of an arbitrarily shaped scatterer and the (b) equivalent model for a metafilm where the GSTCs are applied [4].	147
Figure 6.11: Metafilm placed between two parallel metal plates [4].	149
Figure 6.12: Surface array of metallic square patches with a period 'p' and a patch width 'l' [12].	150
Figure 6.13: Resonant frequency of an air-filled parallel-plate resonator loaded with a square-patch metafilm [10].	150
Figure 6.14: Overall unit cell geometry of (a) square AMC and (b) AMC square ring centred between substrates. Front view dimensions of (c) square AMC (d) square ring AMC.	152
Figure 6.15: Floquet port boundary conditions for a square and square ring AMC. (a) and (b) shows the defined open boundary conditions, (c) is the defined periodic boundary condition around all four edges and (d) is the perfect electrical conductor boundary condition at the base of the structure in place of a ground plane.	153
Figure 6.16: Reflection phase comparison of a solid square and square ring AMC metafilm.	154
Figure 6.17: Adjacent AMC unit cells with equivalent circuit elements (a) front view and (b) cross-sectional view.	155

Figure 6.18: Equivalent circuit model for two-unit cells of the proposed AMC.....	155
Figure 6.19: Reflectance and Transmittance across the unit cell metafilm.	157
Figure 6.20: Phase of the transmittance of the unit cell metafilm.....	158
Figure 6.21: Complex Refractive Index of the unit cell metafilm.	158
Figure 6.22: Relative permittivity of the unit cell metafilm.	159
Figure 6.23: Electric susceptibility $H_d(\epsilon_r - 1)$ extracted using the effective medium approach of the AMC metafilm unit cell.	160
Figure 6.24: Electric susceptibility χ_{ESxx} using the equations derived from the generalized sheet transition conditions (GSTCs).....	161
Figure 6.25: Comparison of the real electric susceptibilities from computed using the effective medium approach and the generalized sheet transition conditions (GSTCs) approach.	162
Figure 6.26: Comparison of the imaginary electric susceptibilities from computed using the effective medium approach and the generalized sheet transition conditions (GSTCs) approach.	162
Figure 6.27: The equivalent effective medium model for a metafilm loaded into a dielectric substrate [1].....	164
Figure 6.28: Front view of the proposed AMC metafilm with the trapezoidal patch superimposed.	165
Figure 6.29: Front View Geometry of Trapezoidal Patch with a loaded AMC metafilm.	165
Figure 6.30: Simulated input reflection coefficient comparison between a rectangular patch antenna, trapezoidal patch antenna and trapezoidal patch antenna with an AMC metafilm loaded into the dielectric substrate.	167
Figure 6.31: Simulated radiation patterns in terms of realized gain for the antennas (a) XZ plane and (b) YZ plane.	168
Figure 6.32: Cross sectional view for antenna bending modelled as an annulus bent over an invisible cylinder with a sectoral bending angle Φ and radius R.	169
Figure 6.33: TPA-MF cylindrically bent along its (a) H-plane and (b) E-plane.	169
Figure 6.34: Simulated input reflection coefficient for H-plane bending of the TPA-MF.	170
Figure 6.35: Simulated radiation pattern in terms of realized gain for the antenna's (a) H-plane and (b) E-plane while bending the antenna across its width or H-Plane.....	171
Figure 6.36: Resonant frequency shift comparison between the antennas with respect to increasing bending angles across the H-plane.	172
Figure 6.37: Bandwidth comparison between the RPA, TPA and TPA-MF while bending along the H-plane.	172
Figure 6.38: Directivity comparison between the RPA, TPA and TPA-MF while bending along the H-plane.	173

Figure 6.39: Realized Gain comparison between the RPA, TPA and TPA-MF while bending along the H-plane.....	174
Figure 6.40: Total Radiation Efficiency comparison between the RPA, TPA and TPA-MF while bending along the H-plane.....	174
Figure 6.41: Dielectric losses for the various TPA-MF H-plane bending angles Φ	175
Figure 6.42: Conductor losses for the various TPA-MF bending angles Φ	175
Figure 6.43: Simulated input reflection coefficient for E-plane bending of the TPA-MF.	177
Figure 6.44: Resonant frequency shift comparison between the RPA, TPA and TPA-MF with respect to increasing bending angles across the E-plane.	178
Figure 6.45: Simulated radiation pattern in terms of realized gain for (a) XZ plane and (b) YZ plane while bending the TPA-MF across its length or E-Plane.	179
Figure 6.46: Bandwidth comparison between the RPA, TPA and TPA-MF while bending along the E-plane.	180
Figure 6.47: Simulated maximum realized gain comparison between the RPA, TPA and TPA-MF for various E-plane bending angles Φ	181
Figure 6.48: Simulated maximum directivity comparison between the RPA, TPA and TPA-MF for various E-plane bending angles Φ	181
Figure 6.49: Simulated Total Radiation Efficiency comparison between the RPA, TPA and TPA-MF for various E-plane bending angles Φ	182
Figure 6.50: dielectric losses for the various bending angles Φ along the YZ-plane or E-plane.	183
Figure 6.51: Conductor losses for the various bending angles Φ along the YZ-plane or E-plane.	183
Figure 7.1: Body model configurations to test the various antennas performance in the presence of human tissue.....	197
Figure 7.2: E-plane far-field cut in terms of realized gain for TPA while bending along its H-plane..	199
Figure 7.3: H-plane far-field cut in terms of realized gain for TPA while bending along its H-plane.	200
Figure 7.4: E-plane far-field cut in terms of realized gain for TPA-MF while bending along its E-plane.	201
Figure 7.5: H-plane far-field cut in terms of realized gain for TPA-MF while bending along its E-plane.	202
Figure B.1: Photographs of the fabricated prototype for the (a) RPA and (b) TPA.	238
Figure B.2: Measured S11 comparing RPA with the TPA.	238
Figure B.3: Incorporation of E-plane bending using the polystyrene fixtures.....	239
Figure B.4: Measured S11 for E-plane bending the RPA for various bending angles Φ	240
Figure B.5: Measured S11 for E-plane bending the TPA for various bending angles Φ	241

Figure D.1: Comparison of ϵ'_r for the layers of tissue used in the cubic and cylindrical tissue models.	244
Figure D.2: Comparison of ϵ''_r for the layers of tissue used in the cubic and cylindrical tissue models.	245
Figure D.3: ϵ'_r for the muscle-fat averaged tissue in the Homogenous Phantom Model.	246
Figure D.4: ϵ''_r for the muscle-fat averaged tissue in the Homogenous Phantom Model.	247
Figure D.5: Comparison of ϵ'_r for the layers of tissue used in the FEM Human model.	248
Figure D.6: Comparison of ϵ''_r for the layers of tissue used in the FEM Human model.	248

List of Tables

Table 2.1: Dimensions of the rectangular microstrip patch antenna approximated using the Transmission Line Model.	29
Table 3.1: Reported textile patch antennas in literature.	41
Table 4.1: Comparison of reported miniaturized planar antennas in literature.	55
Table 4.2: Dimensions obtained using the Transmission Line Analysis for a RPA for operation at 2.4 GHz.	81
Table 4.3: A comparison of the simulated dimensions and radiation performance parameters of the patch antenna by varying the aperture width W_2 using the frequency domain solver with a tetrahedral mesh.	83
Table 4.4: A comparison of the simulated dimensions and radiation performance parameters of the patch antenna by varying the aperture width W_2 using the time domain solver with a hexahedral mesh.	83
Table 5.1: Measured -10 dB bandwidth for the patch antenna bent over a cylinder of various radii [5].	104
Table 5.2: Simulated dimensions for a rectangular and trapezoidal patch antenna for operation at 2.4 GHz.	118
Table 5.3: Performance Parameters Comparison for YZ-plane or E-Plane Bending Angles.	122
Table 5.4: Performance parameter comparison for H-Plane bending angles.	125
Table 5.5: Performance parameters comparison for E-plane bending angles.	131
Table 5.6: Performance parameters comparison of the TPA for H-plane bending angles.	136
Table 6.1: Dimensions of Solid Square AMC unit cell.	153
Table 6.2: Dimensions of the Square Ring AMC unit cell.	153
Table 6.3: AMCs bandwidth performance in comparison to its occupied surface area at a zero-degree reflection phase at 2.4 GHz.	154
Table 6.4: Dimensions of Rectangular patch, Trapezoidal patch, and Trapezoidal patch with an AMC metafilm.	166
Table 6.5: Performance parameter comparison of a RPA, TPA and TPA-MF.	166
Table 6.6: Performance parameters comparison for XZ plane bending angles.	170
Table 6.7: Performance Parameters Comparison for YZ Plane Bending Angles.	177
Table 7.1: Summary of SAR exposure limits from various standards.	193
Table 7.2: Comparison between the TPA and TPA-MF simulated in free space and placed on the cubic tissue model.	198

Table 7.3: Comparison of the computational requirements using the frequency domain solver for the TPA and TPA-MF simulated in free space and placed on the cubic tissue model.	198
Table 7.4: E-plane bending comparison between the TPA and TPA-MF simulated in free space and placed on the FEM Human Model.	199
Table 7.5: H-plane bending comparison between the TPA and TPA-MF simulated in free space and placed on the FEM Human Model.	200
Table 7.6: Comparison of the computational requirements using the frequency domain solver for bending the TPA and TPA-MF in free space and placed on the FEM Human Model.	202
Table 8.1: Proposed antennas compared with some textile patch antennas reported in Chapter 2.	213
Table B.1: Resonant behaviour for E-plane bending the RPA over the cylindrical fixtures.	240
Table B.2: Resonant behaviour for E-plane bending the TPA over the cylindrical fixtures.....	241
Table C.1: Performance parameters of the TPA in the presence of various tissue models using the frequency domain solver in CST.....	242
Table C.2: Performance parameters of the TPA-MF in the presence of various tissue models using the frequency domain solver in CST.	242
Table C.3: Computational costs using the frequency domain solver.	243
Table D.1: Material property definitions in CST Microwave Studio for human tissue in the cubic and cylindrical tissue model.....	244
Table D.2: Material property definitions in CST Microwave Studio for Homogenous Phantom Model.	246
Table D.3: Material property definitions in CST Microwave Studio for the FEM Human tissue model.	247

1 Introduction

1.1 Project Motivation

With the growth of wearable technologies and device communication, there has been an increasing interest in research of antennas that are used in wearable systems. These antennas carry unique features such as physical flexibility, allowing for seamless system integration. The wearable antenna has application in military, medical, space and emergency service applications [1]. Designing wearable antennas can be challenging since they need to be non-invasive, inconspicuous, robust, lightweight, inexpensive and low maintenance [2][3].

Wearable antennas can be categorized into textile and non-textile, with textile producing the most seamless antenna structures into fabrics. Non-textile solutions are more suited to being embedded into wearable accessories such as glasses, buttons, belts and so on [2].

The microstrip patch antenna with its ability to be integrated into wearable items makes it an excellent candidate for body centric wireless communication systems. It can be placed conformally on the body and consists of a ground plane that prevents backpropagating radiation into the body and eliminates the frequency detune that occurs due to coupling between the human body and the antenna [4]. This makes the wearable antenna ideal for off-body communication links since the radiation pattern is broadside, however the microstrip patch antenna has also been used for on and off-body applications simultaneously by exciting another mode which has an omnidirectional radiation pattern [5][6].

Application examples for wearable patch antennas are:

- On body power harvesting [7][8].
- Soldiers uniform acting as a part in a communication and surveillance network between soldiers, other units, armoured vehicles, and UAVs [9].
- Survival suit embedded with a GPS antenna [10].
- Lifejacket with an embedded search and rescue transmitter [11].



Figure 1.1: A Jacket integrated with a circularly polarized patch antenna on the arm as part of a GPS receiver [10].



Figure 1.2: lifejacket equipped with patch antennas as part of a COSPAS-SARSAT search and rescue transmitter [11].

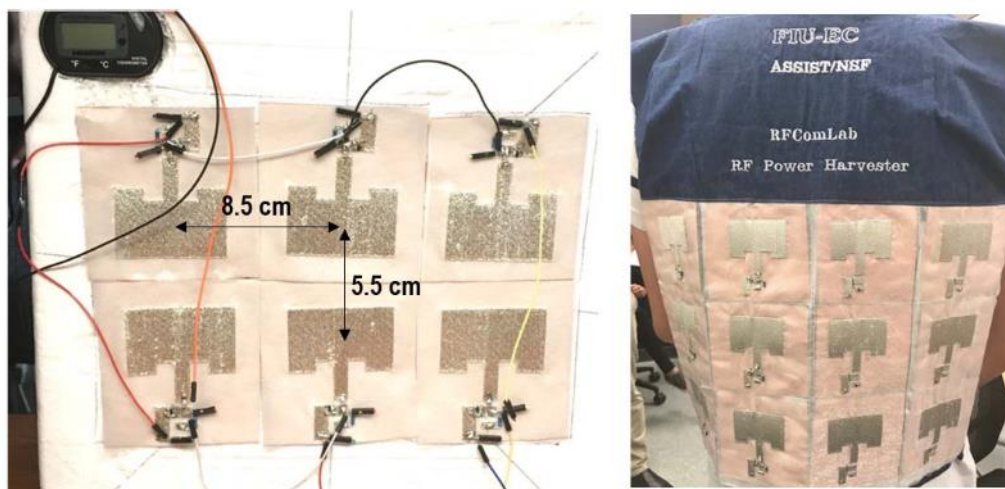


Figure 1.3: Six-element wearable microstrip patch antenna array for RF power harvesting [8].

Important design features and considerations for wearable antennas include:

- Selection of textile material
- Characterizing antenna deformations such as bending and crumpling
- Influence of ground plane and its size
- Effects of material conductivity
- Performance enhancement methods
- Effect of the human body on antenna radiation performance
- Positioning location on human body
- Specific absorption rate (SAR)

Miniaturization is an important requirement of the microstrip patch antenna to allow for seamless integration.

Understanding the interaction between human tissue/body and its effect on antenna radiation performance allows us to characterize the antennas suitability for wearable applications. With the increasing capabilities of CPUs, GPUs, RAM and efficient electromagnetic solvers, large electromagnetic body simulations can be performed accurately while being memory efficient and without the use of high-end workstations.

In this thesis an optimized trapezoidal patch antenna is proposed as a miniaturized variant of the rectangular patch antenna. This antenna aims to be appropriate as a textile wearable antenna for off-body communication systems.

1.2 Project Challenges and Aims

Planar wearable antennas must be designed to deal with the unique challenges of being on the human body. The key challenges are outlined below.

1. Antenna bending which results in resonant frequency detuning.

Conformal placement of the patch antenna on the human body will result in antenna bending to a certain degree. Stretching and crumpling in antennas with textile substrates can also alter resonant behaviour. This has been studied extensively in literature and several articles focusing on antenna deformation have been referenced [9], [12]–[19]. Some key findings are:

- Bending a patch antenna along its E-plane detunes reduces the resonant length of the antenna thus increasing the resonant frequency. Thus, modifying the antenna's E-plane has the dominant effect on antenna's performance.

- Radiation patterns are affected during antenna bending.
- Bending broadens the major lobe of the radiation pattern.
- The Front to Back Ratio of the radiation pattern is decreased for increased bending angles.
- A textile substrate if stretched can alter the permittivity, thickness, and thus resonant frequency of the antenna.

2. Ensuring a compact and low-profile antenna.

A low-profile antenna results in resilience to bending deformations when placed conformally on a curved inhomogeneous structure such as the human body. This also provide seamless and inconspicuous integration to on body applications. The aim of antenna miniaturization is to reduce the size by altering the length, area or volume, while maintaining performance of the original antenna [20].

Various miniaturized antennas are highlighted in Table 4.1 which shows that through miniaturization, antenna performance parameters such as bandwidth, radiation efficiency and realized gain can decrease. Techniques are required to maintain or enhance these performance parameters through the miniaturization process.

3. Antenna's proximity to lossy human tissue.

A further challenge is the proximity of lossy human tissue which reduces the antennas radiation efficiency, realized gain and detunes the resonant frequency due to introduced high dielectric constant of the various layers of human tissue [21]–[24].

4. Seamless integration of the antenna in clothing.

For practical use in wearable applications, seamless integration of the antenna is required. This can be achieved using electro-textiles woven into fabrics. This creates a robust structure without the need of any glues to bind the conductive elements to the substrate. When electro-textiles are integrated into fabrics as the conductive elements for wearable antennas, maintaining a low sheet resistance is crucial to minimizing conductor losses. The anisotropic nature of different weave patterns and roughness profile of electro-textiles can also effect the sheet resistance [25][26]. Higher antenna efficiencies can be obtained by ensuring the principle current flow is parallel to the stitch direction [1]. Moisture absorption into fabrics can change dimensional properties and electromagnetic properties by increasing the dielectric constant and loss tangent [27]. Adding a moisture barrier to the textile antenna can prevent the alterations [28]. Issues associated with seamless integration using electro-textiles

can be summarised as maintaining low sheet resistance of conductive fibres and moisture absorption of the textile antenna. Various wearable antennas have been designed using electro-textiles displaying optimal radiation characteristics [21], [29]–[34].

1.2.1 Key Contributions

This thesis will address specific elements of the challenges mentioned above to develop an improved patch antenna for wearable off-body communication systems :

- Preventing frequency detuning by ensuring a low-profile patch miniaturised antenna design. Proposing a trapezoidal geometry for the radiating element results in a significant surface area size reduction for the radiating element when compared to a rectangular patch antenna by up to 39%.
- Adaptation of the Aperture Array Model (AAM) which offers physical insight for the operating mechanism and attempts to predicts the far-field characteristics of the trapezoidal patch antenna in planar and cylindrical configurations.
- Improving the antenna's bandwidth and realized gain following on from miniaturization of the patch antenna. This will be addressed by the proposal of miniaturized Artificial Magnetic Conductor (AMC) cells which are uniquely loaded inside of the dielectric substrate of the patch antenna. The AMC cells are designed to have a zero-degree phase reflection at the resonant frequency of the patch antenna. The backward propagating waves are reflected at the ground plane and constructively interfere with the outward radiating waves. Loading the proposed AMC structure into the trapezoidal patch antenna results in a realized gain enhancement by up to 2.1 dBi, a bandwidth enhancement by 1.23%, as well as further miniaturization of the radiating element of up to 49%.

1.3 Project Approach

One of the requirements of a wearable textile antenna is being low profile. A trapezoidal patch antenna loaded with an artificial magnetic conducting metafilm (TPA-MF) is proposed as a surface area size reduced radiating element. The antenna is designed for off-body communication in body-centric wireless communications systems that operate in the industrial, scientific, and medical (ISM) band at 2.4 GHz.

To begin, the TPA which has been proposed as a miniaturized variant of the rectangular patch antenna is investigated for its suitability as a wearable textile antenna where the antenna is subject to deformation. The dielectric characteristics of the substrate are based on the cotton/polyester sample in [35] which is modelled as a 2.8 mm thick homogenous medium with a dielectric constant of 1.6 and a loss tangent of 0.02.

The aperture array or cavity model for a rectangular patch as presented by T. Bird [36] and C. Balanis [37] is modified to model a trapezoidal patch antenna analytically. A far-field analysis is formulated where all the necessary equations are derived to describe the radiation mechanism that occurs. The equations are implemented in MATLAB to approximate the far-field behaviour. A full wave simulation is performed using the electromagnetic simulation software CST Microwave Studio to benchmark and verify the proposed analytical model.

The non-uniform surface of the human body means that when the antenna is placed conformally on the surface various bending can occur. Depending on the location, the degree to which bending can occur will vary. Therefore, the analytical analysis of the patch antenna is developed to model cylindrical bending. The results are validated by modelling antenna bending in CST Microwave Studio. The simulation software allows a more complete characterisation of antenna bending.

An antenna which is designed for wearable applications will be in proximity to the human body. The antenna will be simulated and investigated in the presence of various body tissue configurations, starting from simple cubic and cylindrical tissue models to more complex body models.

Since the introduction of the surrounding tissue results in a reduction in radiation efficiency and thus gain [4][11][19], the TPA-MF is proposed to improve the antenna's gain. The AMC is designed and simulated in CST and then loaded into the dielectric substrate of the edge fed patch antenna. The body model configurations will be simulated to verify any improvements in performance.

1.4 Thesis Overview

Chapter 1 provides an overview of the project including its aims, objects, and approach.

In Chapter 2, all the relevant parameters to describe the performance of an antenna are listed. Also, the simulation settings using the software CST Microwave Studio are outlined.

Chapter 3 presents a historical overview on the Microstrip Patch Antenna and its design evolutions leading to its more recent applications in wearable systems.

Chapter 4 first focuses on a literature review on miniaturization techniques that have been applied to patch antennas. Following on from this review the TPA is further investigated as a miniaturized alternative to the traditional RPA by analysing the TPA using the Aperture Array Model (AAM). The AAM is an analytical characterization of the TPA which provides intuition and insight on its far-field characteristics. Also, it has not been presented previously in literature for the TPA. Following on from this analysis, a more rigorous characterization of the TPA is presented using the full-wave simulation software CST Microwave Studio.

Chapter 5 first presents a literature review on analytical methods to model patch antenna bending. The AAM previously presented in Chapter 4 to characterize the TPA is extended to the Cylindrical Aperture Array Model (CAAM) which incorporates antenna bending. A full-wave simulation analysis is then performed comparing the radiation performance between the RPA and TPA for E-plane and H-plane bending.

Chapter 6 presents the key novelty of this thesis, that is the trapezoidal patch antenna loaded with an artificial magnetic conducting metafilm (TPA-MF). The chapter introduces the background and theory on the artificial magnetic conductor (AMC). A square ring is chosen as the geometry for the AMC unit cell for its simplicity and being miniaturized. The Generalized Sheet Transition Conditions (GSTCs) and the Effective Medium Model are two methods to characterize the AMCs dielectric response. These characterization methods are compared to each other. This is performed through full-wave simulation analysis. Next the AMC is uniquely implemented into the substrate of the TPA creating the TPA-MF, where its performance is compared to the RPA and TPA. Antenna bending is also investigated. The TPA-MF achieves a miniaturization of the radiating element, a bandwidth, realized gain enhancement and is shown to be a more efficient radiator compared to the TPA.

Chapter 7 presents an introduction to computational human models for specific absorption rate (SAR) compliance testing for antennas. The frequency dependency of the dielectric properties of human tissue are incorporated into the human tissue models. Tissue models of

increasing geometric complexity are introduced. The cylindrical tissue model, homogenous body model, and FEM meshable model incorporate antenna bending, which is investigated for both E and H-plane bending. A comparison is presented for the various configurations which are compared to the antenna simulations performed in free space.

Chapter 8 concludes the thesis by summarizing the key findings, offering recommendations for future work, and reflecting on the research journey.

In Appendix A, the MATLAB codes for the implementation of the AAM and CAAM are provided.

In Appendix B, preliminary S-parameter antenna measurements for the RPA and TPA are provided. The antennas are investigated in a planar and cylindrical configuration to investigate antenna bending.

Appendix C contains the performance parameters of the antenna simulations involving human tissue models from Chapter 7, including their computational costs in terms of simulation time and memory requirements.

Appendix D provides the material definitions for the human tissue models involved in the antenna simulations.

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2 Fundamentals

This chapter provides fundamental background information on the antenna parameters that are used to analyse the antennas' performance throughout this thesis. It also provides the simulation settings that are used within the full-wave simulation software CST Microwave Studio.

2.1 Fundamental Antenna Parameters

To describe the performance of an antenna, definitions of various parameters are presented in this section. Many of these definitions can be found in the IEEE Standard Definitions of Terms for Antennas [1].

2.1.1 Reflection Coefficient and VSWR

The reflection coefficient Γ calculates the ratio between reflected power and total incident power at the input terminals of an antenna [2].

The reflection coefficient Γ is defined as

$$\Gamma = \frac{Z_{in} - Z_0}{Z_{in} + Z_0} \quad (2.1)$$

Where Z_{in} is the input impedance of the port feeding the antenna and Z_0 is the characteristic impedance of the transmission line [3].

The reflection coefficient Γ is unitless and is equivalent to the scattering parameter S_{11} . The scattering parameters relate the waves that are incident on the ports to the waves that are reflected and transmitted through [4]. In antenna analysis where the antenna is excited by a single port, the reflection coefficient Γ is the only scattering parameter of interest since it is a one-port system. For a two-port system as shown in Figure 2.1, a relationship can be established between the reflected waves and the transmitted waves to each port.

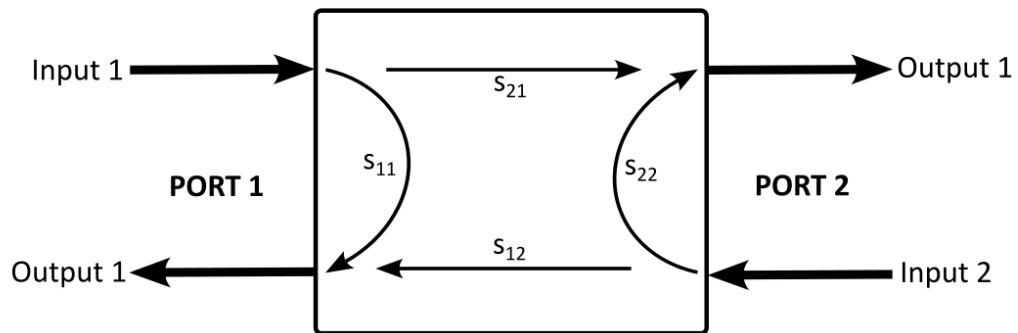


Figure 2.1: Scattering parameters for a two-port network.

In Figure 2.1 S_{11} is the forward reflectance, S_{21} is the forward transmittance, S_{22} is the backward reflectance and S_{12} is the backward transmittance. S_{11} and S_{22} are equivalent to the reflection coefficient Γ defined in (2.1) for their respective ports. By relating the reflected and transmitted waves through a homogenous slab of material, the two-port scattering parameters allows the retrieval of the permittivity and permeability of a material.

The voltage standing wave ratio (VSWR) is related to the reflection coefficient which calculates the ratio between the maximum V_{max} and minimum V_{min} voltage (V) of a standing wave. Both the reflection coefficient Γ and VSWR are equivalent in analysing transmission line efficiency and are closely linked by the formulas below.

$$VSWR = \frac{V_{max}}{V_{min}} = \frac{1 + |\Gamma|}{1 - |\Gamma|} \quad (2.2)$$

$$\Gamma = \frac{VSWR - 1}{VSWR + 1} \quad (2.3)$$

VSWR is also unitless and measured as a ratio relative to the lowest voltage. For example, a 1:1 VSWR means that no reflections occur which means that it is a matched load or ideal system. While a 2:1 VSWR is equivalent to a reflection coefficient Γ of approximately 0.33, which means there is a small amount of reflections. Figure 2.2 shows the point at which minimal reflections occur for various antennas in terms of reflection coefficient S_{11} (dB) and VSWR. The point at which minimal reflections occur is known as the resonant frequency of the antenna.

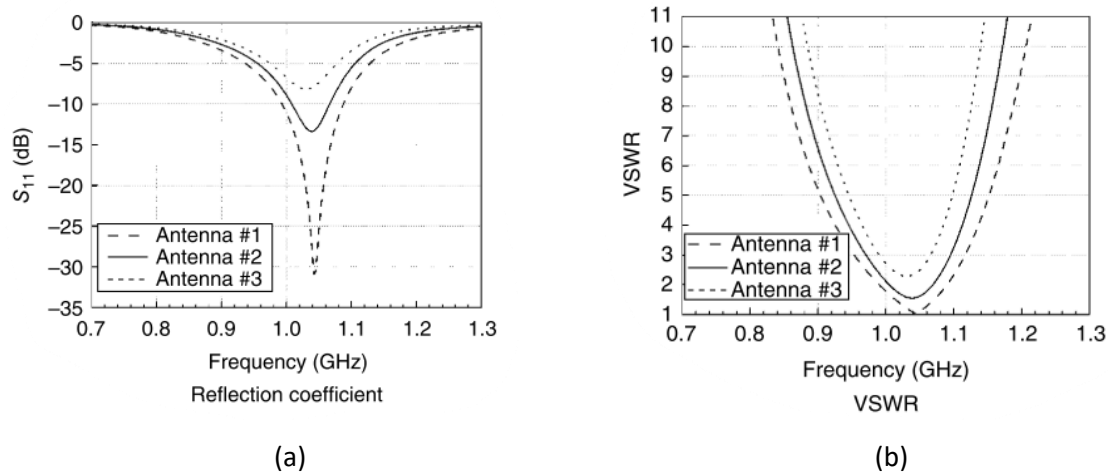


Figure 2.2: A comparison between (a) reflection coefficient and (b) VSWR analysis for antennas [2].

In this research the reflection coefficient measured in dB is preferred since it is better for illustrating the minor changes in the reflection properties of an antenna.

2.1.2 Bandwidth

The impedance bandwidth can be measured from the reflection coefficient graph as the range of frequencies over which the reflection coefficient is small enough for the device to operate as intended [3]. The acceptable values of the reflection coefficient for calculating bandwidth depend on the designers standard and needs of the particular applications [3].

Impedance bandwidth can be calculated from

$$\text{Bandwidth} = \frac{f_u - f_l}{f_0} \times 100\% \quad (2.4)$$

Where f_u and f_l is the upper and lower limit of the frequency band in respect to the reflection coefficient boundary used. f_0 is the resonant or centre frequency located at where the reflection coefficient or VSWR is at its lowest point. Reflection coefficient boundaries of -10dB or -15 dB are commonly used to calculate the bandwidth [5], while [2] mentions that even a boundary of -6 dB is acceptable for an internal multiband antenna. In this research a reflection coefficient boundary of -10 dB is used to compute the bandwidth.

2.1.3 Efficiency

The reflection mismatch efficiency e_r is a dimensionless parameter can be represented as

$$e_r = 1 - |\Gamma|^2 \quad (2.5)$$

The total efficiency e_0 of an antenna can be computed considering the conductor e_c , dielectric e_d and mismatch e_r efficiencies of the antenna. Total efficiency is a dimensionless quantity expressed as

$$e_0 = e_r e_c e_d \quad (2.6)$$

The product of the conductor and dielectric efficiency is also known as the radiation efficiency of the antenna, while incorporating the mismatch efficiency with the radiation efficiency is known as the total radiation efficiency. In this research the total radiation efficiency is the parameter of interest since it encompasses all three of the efficiencies.

2.1.4 Radiation Pattern Parameters

The radiated fields from the antenna can be represented by the radiation pattern it produces. There are three field regions, reactive near-field, radiating near-field and the far-field region. The reactive near field is the region where energy is stored and returned to the antenna [5]. The next region is the radiating near field where the angular field distribution is dependent upon the distance from the antenna [3]. The far-field region is where the angular field distribution is essentially independent of the distance from the antenna and the field components are transverse. This region occurs at a minimum distance of $2D^2/\lambda$, where D is the largest dimension of the antenna and λ is the operating wavelength of the antenna [3]. A visual representation of the various field regions and distributions can be shown in Figure 2.3 and Figure 2.4. The radiated fields from an antenna in this research are analysed only in the far-field.

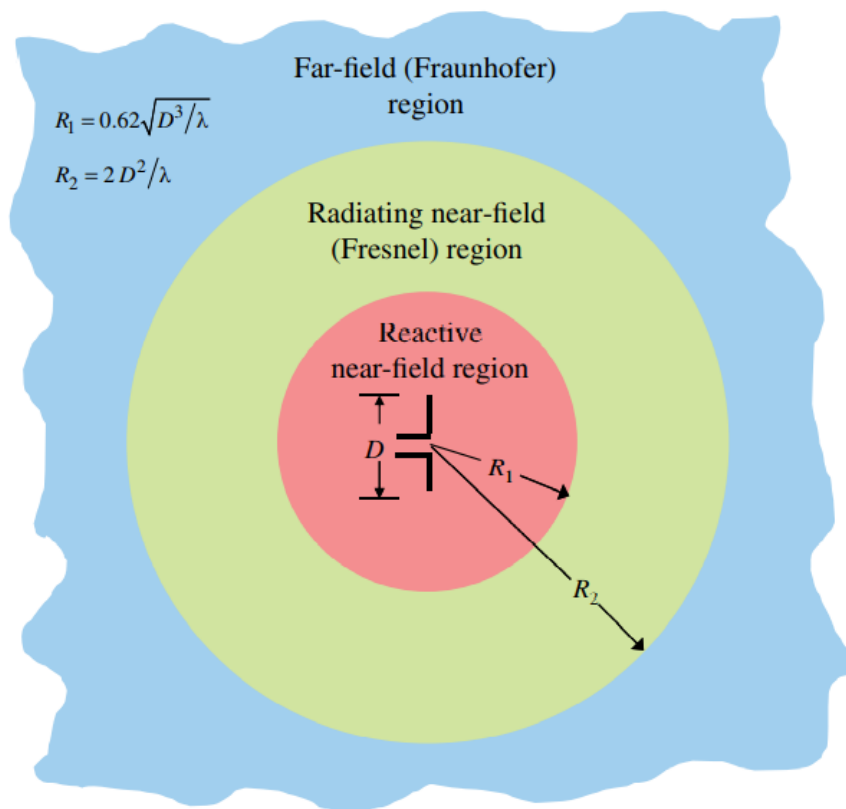


Figure 2.3: Field regions of an antenna [3].

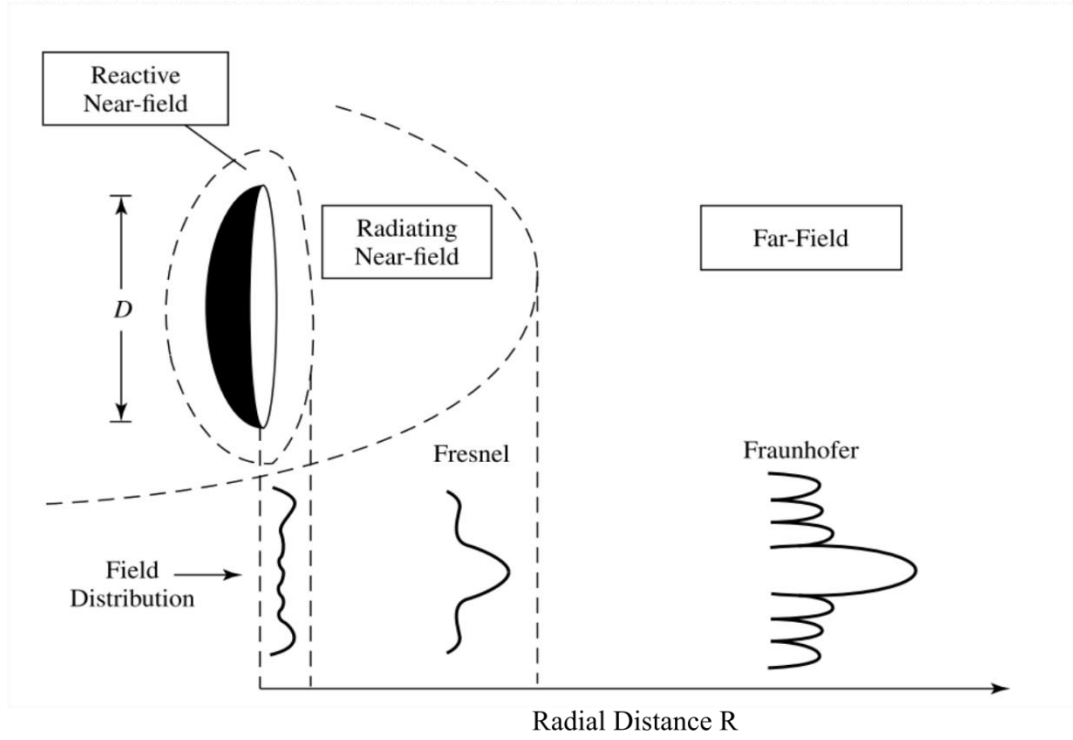


Figure 2.4: Changes of an antennas amplitude pattern shape from the reactive near-field towards the far-field [6].

To characterize the far-field behaviour of an antenna the average power density W_{rad} (W/m²), radiation intensity $U(\theta, \phi)$ (W/unit solid angle), radiated power P_{rad} (W), directivity D (dBi) and realized gain $G_{re}(\theta, \phi)$ (dBi) need to be defined. The average power density is equivalent to the time average Poynting vector which describes the power associated with an electromagnetic wave given by

$$W_{rad} = \frac{1}{2} \text{Re}[\mathbf{E} \times \mathbf{H}^*] \quad (2.7)$$

Where $\mathbf{E}$ is the complex electric field intensity (V/m) and $\mathbf{H}$ is the complex magnetic field intensity (A/m).

The radiated power is the surface integral of the Poynting vector over a large sphere extending into the far-field given by

$$P_{rad} = \oint W_{rad} \cdot d\mathbf{s} = \frac{1}{2} \oint \text{Re}[\mathbf{E} \times \mathbf{H}^*] \cdot d\mathbf{s} \quad (2.8)$$

The radiation intensity is defined as the power radiated from an antenna per unit solid angle [3].

This parameter is obtained by multiplying the average power density W_{rad} by the square of the far-field distance r^2 . The Radiation intensity U is also known as the power pattern [7] defined as

$$U(\theta, \phi) = r^2 W_{rad} \quad (2.9)$$

The directivity is the ratio of the radiation intensity in a given direction from the antenna to the radiation intensity averaged over all directions [3].

The directivity D can be written as

$$D = \frac{U}{U_0} = \frac{4\pi U}{P_{rad}} \quad (2.10)$$

Where U_0 is the radiation intensity of an isotropic source given by

$$U_0 = \frac{P_{rad}}{4\pi} \quad (2.11)$$

The maximum directivity D_{max} can be defined as

$$D_{max} = \frac{4\pi U_{max}}{P_{rad}} \quad (2.12)$$

Realized gain $G(\theta, \phi)$ is a measure that considers the efficiency of the antenna as well as its directional capabilities, combining directivity with total radiation efficiency.

This is defined as

$$G_{re}(\theta, \phi) = e_r e_d e_c D(\theta, \phi) \quad (2.13)$$

In antenna analysis the gain is usually expressed in terms of decibels relative to an isotropic radiator (dBi). The conversion is given by

$$G_{re}(\text{dBi}) = 10 \log_{10}[G_{re}] \quad (2.14)$$

The aim of calculating these parameters is to compute or measure the angular distribution of the radiation pattern over an antenna. Observations of the radiation are made on a large sphere of constant radius extending into the far-field [3] as shown in Figure 2.5.

Radiation patterns are usually displayed in terms realized gain. It may also be found that the radiation pattern is portrayed as a normalized power plot such as in Figure 2.6 and Figure 2.7.

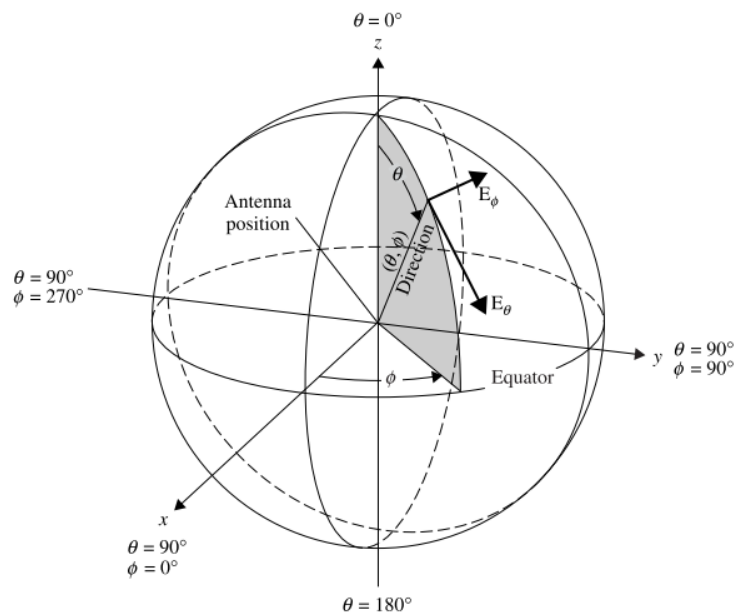
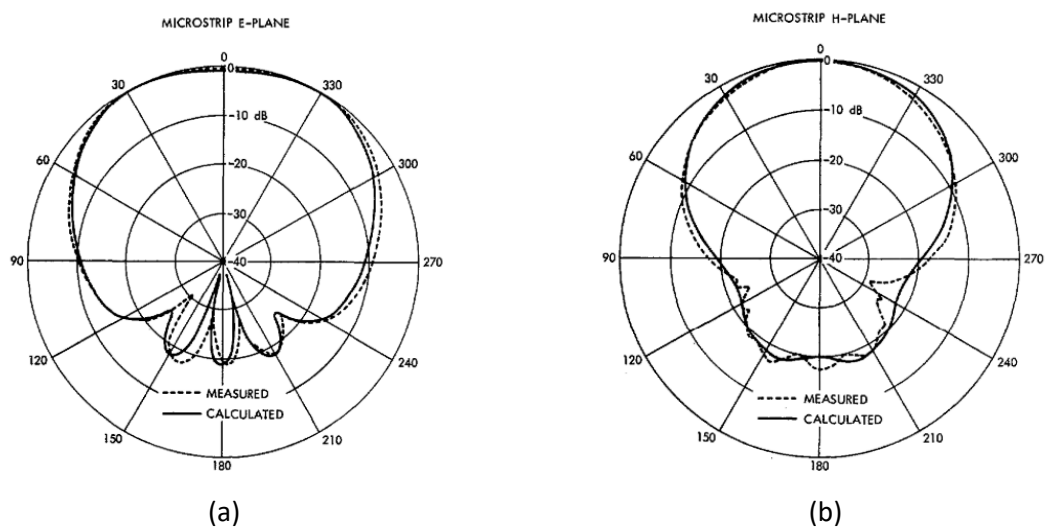


Figure 2.5: Spherical coordinate system geometry for antenna far-field analysis [8].

The radiation pattern can be displayed as a 3D pattern or a 2D planar cut. The 2D planar cuts are usually the choice for antenna analysis since this enables to investigate the radiation properties in different planes. A typical radiation pattern plot for a patch antenna is given in Figure 2.6 below. The E and H-planes of the patch antenna is illustrated in Figure 2.6(c).



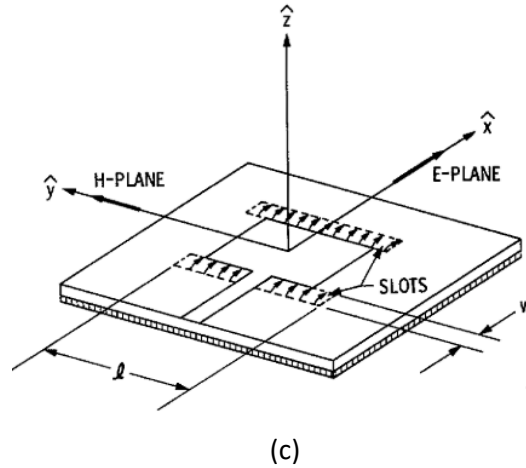


Figure 2.6: 2D normalized radiation power plots for a microstrip patch antenna for the (a) E-plane and (b) H-plane. The antennas geometry, radiating edges modelled as slots and corresponding planes are given in (c) [9].

A directional radiation pattern typically consists of a major, minor, side, and back lobe. A visual representation of the various possible lobes can be shown in Figure 2.7 and Figure 2.8 for a polar and linear radiation plot. The radiation pattern is plotted as a normalized power amplitude pattern. The major lobe occurs at the region where the radiation intensity is at its maximum. A minor lobe is defined as any lobe except the major lobe. This is usually radiation produced in undesired directions. A minor lobe encompasses the side and back lobes of the radiation pattern. The back lobe occurs approximately 180° away from the main lobe of propagation for a directional radiation pattern [3].

The beamwidth is another parameter of interest which provides information on the shape of the radiation pattern. Figure 2.7 and Figure 2.8 show two different types of beamwidths, the half-power beamwidth (HPBW) and the first-null beamwidth (FNBW). The half-power beamwidth (HPBW) is the angle between the two directions in which the radiation intensity is one-half the maximum value in the direction of the maximum lobe of the radiation pattern [1]. The first-null beamwidth (FNBW) is the angular separation between the first nulls of the pattern [3].

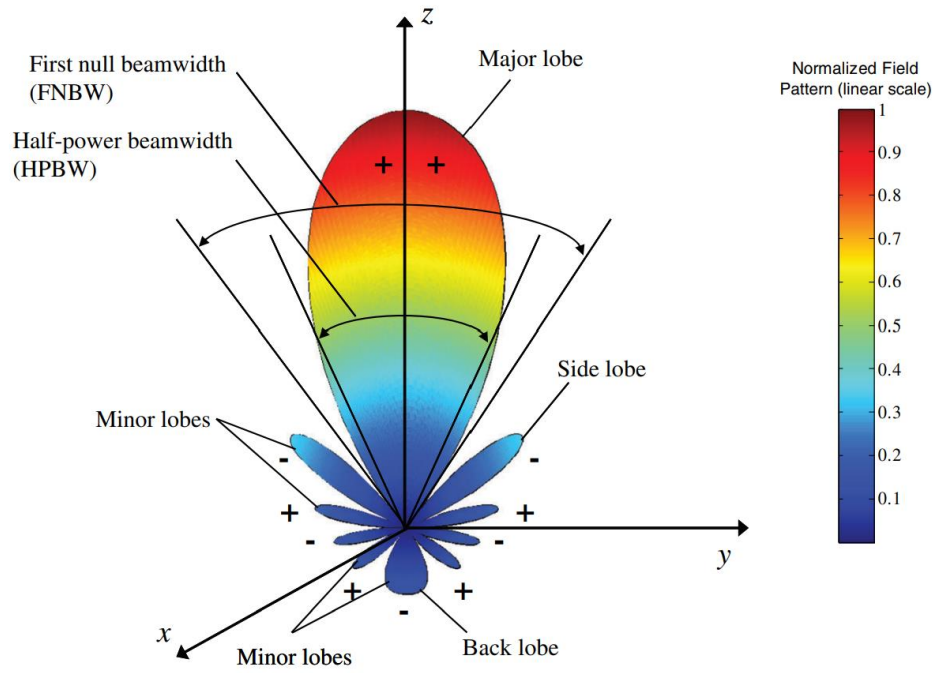


Figure 2.7: Polar radiation power pattern plot showing the various radiation lobes and beamwidth [3].

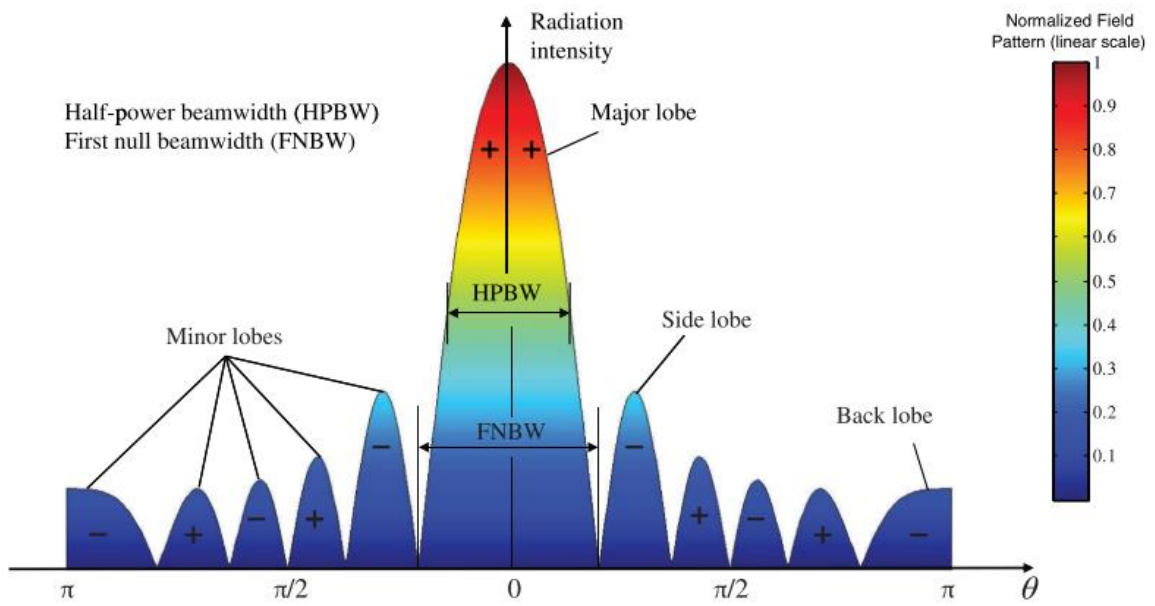


Figure 2.8: Linear radiation power pattern plot showing the various radiation lobes and beamwidth [3].

2.2 Full-Wave Simulations

An accurate full-wave simulation analysis requires [10]:

- Exact material properties definitions for the dielectric and conductive properties of the structure to be simulated,
- An accurate CAD geometry,
- A suitable mesh with mesh refinement techniques applied.
- Suitable choice for the excitation port,
- Appropriate choice of solver,
- Appropriate selection of boundary conditions of the simulation volume,
- Cross verification of results with a different numerical solver where possible.

In this thesis, a computational antenna analysis is performed using the full-wave simulation software CST Microwave Studio. CST is a versatile simulation software, consisting of a variety of EM solvers for a given application. For the type of simulations in this research which focuses on antenna design, solvers such as the frequency and time domain solvers are well suited. The software also consists of a variety of global and local optimizers to study the behaviour of the antenna as its properties are changed. CST is used in leading technology and engineering companies and is argued as the “most powerful general purpose electromagnetic solver for high frequency simulation tasks [11].” Dassault Systèmes, a software company, offers a variety of e-seminars to help users get the most out of CST and has provided a cooperation license for CST Studio Suite to support this work. All simulations are performed using an AMD Ryzen 9 5900X 12-Core Processor with 64GB DDR4 memory. The simulation settings are outlined in this section, focusing on solver options and simulation accuracy.

2.2.1 Excitation

A discrete port is used as excitation for all simulations in this thesis. The impedance element discrete port is used since it allows the calculation of the S-parameters. The port consists of a current source with an inner impedance that emits and absorbs power. This port stimulates an input power of 0.5 W and enables the calculation of correspondent S-parameter, based on the incoming and outgoing time signals [12].

2.2.2 Solver Options

For high frequency full wave simulation analysis, CST offers many solvers depending on the problem type and simulation size. The available numerical methods within CST Studio Suite

include Finite Integration Technique (FIT), Transmission Line Matrix (TLM), Finite Element Model (FEM) and Method of Moments (MoM). CST has a project wizard to help with the selection of the most suitable solver and the best meshing technique based on application areas.

Two solvers are used in this research: the transient or time domain solver and the frequency domain solver. For antenna simulations, the time and frequency domain solvers are used for analysing the antennas scattering parameters, radiation patterns, field distributions and the simulation of dispersive materials.

The time domain solver which is also known as the transient solver is based on the finite integration technique (FIT) [13], which discretizes structures to be simulated over a suitable mesh system using the integral form of Maxwell's equation. The discretized form of Maxwell's equations is known as the Maxwell's Grid Equations.

The frequency domain solver, which is based on the finite elements method (FEM), transforms Maxwell's equations into its phasor form. The equations are related to the transient fields by taking the real part of the product of the field phasor and the time factor.

$$E(t) = \text{Re}\{E(\omega) \cdot e^{j\omega t}\} \quad (2.15)$$

The transient solver in CST uses a hexahedral mesh over a Cartesian grid, while the frequency domain solver can use a hexahedral or tetrahedral mesh. CST documentation recommends tetrahedral meshing of structures when using the frequency domain solver.

For accurate modelling of complex 3D geometries, the transient solver applies the Perfect Boundary Approximation (PBA) technique and the Thin Sheet Technique (TST) for accurate modelling of curved structures and thin PEC materials [12].

2.2.3 Time Domain Convergence Parameters and Mesh Refinement

Time domain solver calculates the development of fields through time at discrete locations and at discrete time samples [12]. An antenna consists of at least one port to excite the structure with an electromagnetic field/signal. A Gaussian pulse is excited for the input signal from the port.

The total energy remaining inside of the simulated structure determines when to stop the simulation. An energy level of -50 dB is set since this allows for extra time for the stimulated fields to decay while also performing a fast simulation.

The auto-regressive (AR) filter is a feature with the time-domain solver which is used to predict the stimulated time signal's next time step, thus speeding up the simulation time. For strongly resonant structures such as an antenna with a high Q-factor, calculating the S-parameters using a Fourier Transform leads to a slow decay of the time signal which can lead to long simulation times [12]. The slow decay causes ripples in the signal decay which is shown to be more apparent in the frequency domain of the signal.

The AR filter predicts the S-parameter result which is used to compare to the actual calculated result. The energy error is computed which compares the time signal to the predicted time signal by the AR filter. If the estimated error outperforms the maximum error, convergence is achieved, and the simulation is stopped. The energy balance is also investigated to observe the quality of the AR filter calculation. The energy balance indicates whether the accuracy of an S-parameter calculation is sufficient. The value for the energy balance should be close to one for most part of the spectrum [12]. If the time signal decays sufficiently in the simulations, the AR filter is not used.

Mesh adjustments are optimized by the adaptive mesh refinement setting which locally refines the regions with high energy density concentrations or field gradients. The simulation is performed numerous times increasing the overall mesh cells by 70% for each subsequent pass. The S-parameters deviation or delta value is compared for each pass. If the deviation between subsequent mesh passes is below a certain limit a convergence is achieved and the simulations halt for further mesh passes. The delta value limit is set to 0.02 by default. An example of the delta convergence using adaptive mesh refinement is shown in Figure 2.9, where a patch antenna is designed and simulated in CST using the time-domain solver with a hexahedral mesh. The graph shows how the delta value decreases with each pass. At the same time, the variation in the S-parameter reduces, as can be seen in Figure 2.10. At 'Mesh Pass = 4' the deviation in the S-parameter when compared to 'Mesh Pass = 3' is minimal with a delta value below 0.02 therefore the mesh refinement process halts.

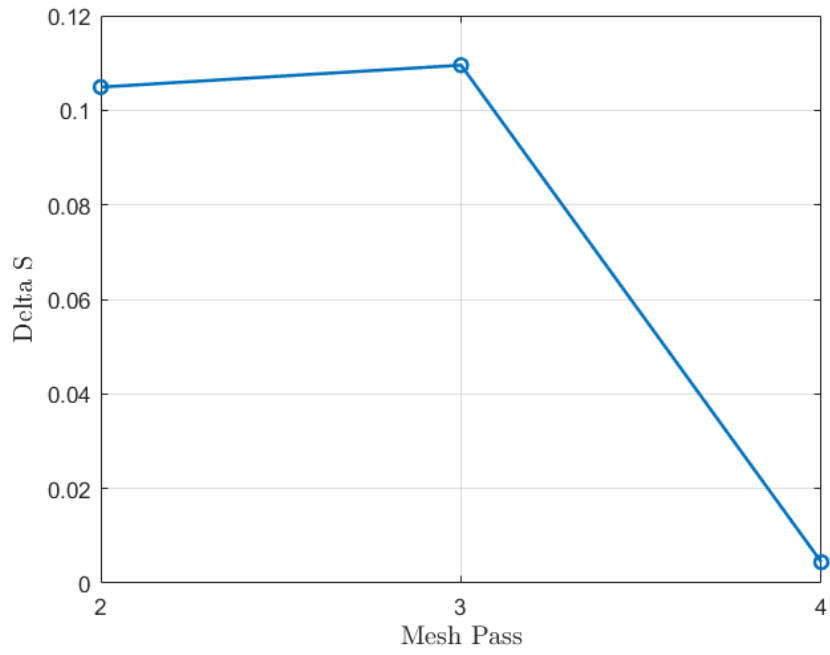


Figure 2.9: Mesh refinement convergence criteria delta using the time-domain solver with a hexahedral mesh.

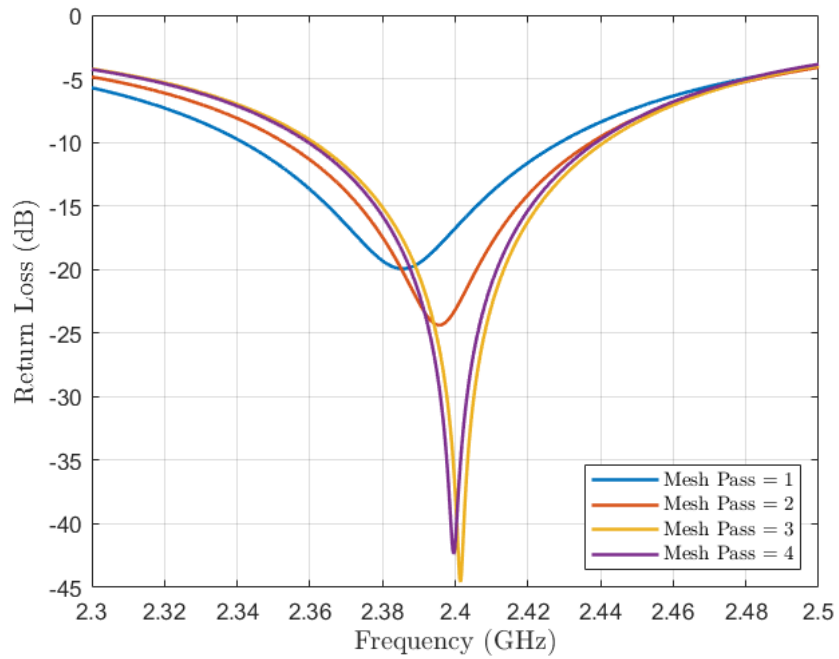


Figure 2.10: Mesh refinement variation in the S-parameters after each subsequent mesh pass.

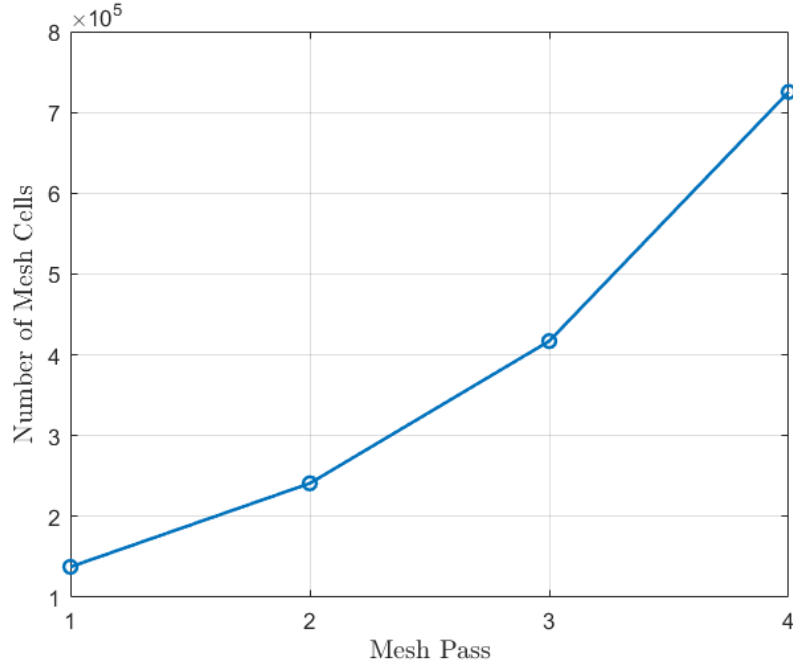


Figure 2.11: Increase in the overall number of mesh cells after each subsequent mesh pass.

Figure 2.11 shows the exponential increase in the mesh cells for each mesh refinement pass. For each pass that the delta does not converge to a value of under 0.02, the mesh cells increase by 70% which are strategically increased over areas where the electromagnetic field concentrations are highest.

2.2.4 Frequency Domain Mesh Refinement

For an accurate frequency domain simulation set up, the general-purpose broadband sweep creates frequency samples over the simulated range. Each frequency sets up a new frequency domain equation system. The general-purpose broadband sweep adaptively selects the frequencies samples to speed up the simulation. The solution to these equations provides the S-parameter result, field distribution and far-field over the simulated structure. The tetrahedral mesh type is chosen for the frequency domain solver, since it is excellent in approximating arbitrary geometries [14]. The tetrahedral mesh creates an unstructured volume of tetrahedrons while the surfaces consist of triangles. Just like for the time-domain solver, the adaptive mesh refinement setting is activated to improve the auto generated mesh during the simulation. A dense mesh is defined before the adaptive mesh refinement technique which optimizes the triangles that form the surface mesh and the tetrahedrons that are generated as the volume mesh. The threshold convergence value for the S-parameter criterion default is set to 0.02 by default. If the maximum difference of the S-parameters

between each adaptive mesh refinement simulation is less than the threshold, mesh refinement halts due to the convergence. The deviation in the S-parameters per mesh refinement is known as the delta value.

2.2.5 Solver and Mesh Structure Selection

The choice of solver for simulations throughout this thesis is dependent on the geometry of the structure to be simulated. For example, modelling antenna bending is best suited being discretized with a tetrahedral mesh since arbitrary geometries can be modelled efficiently [10]. For structures which are electrically large a hexahedral mesh using the time domain solver is more suited [14] since the solver is memory efficient for complex structures [12].

A rectangular patch antenna is designed in Figure 2.12 as an example below to show the difference in how the antenna is discretized using the tetrahedral and hexahedral mesh which is shown in Figure 2.13 and Figure 2.14.

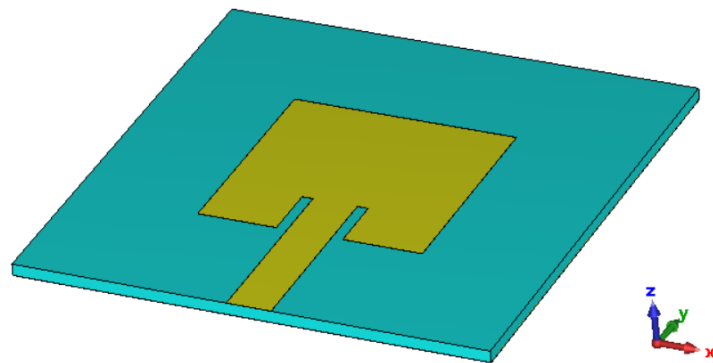
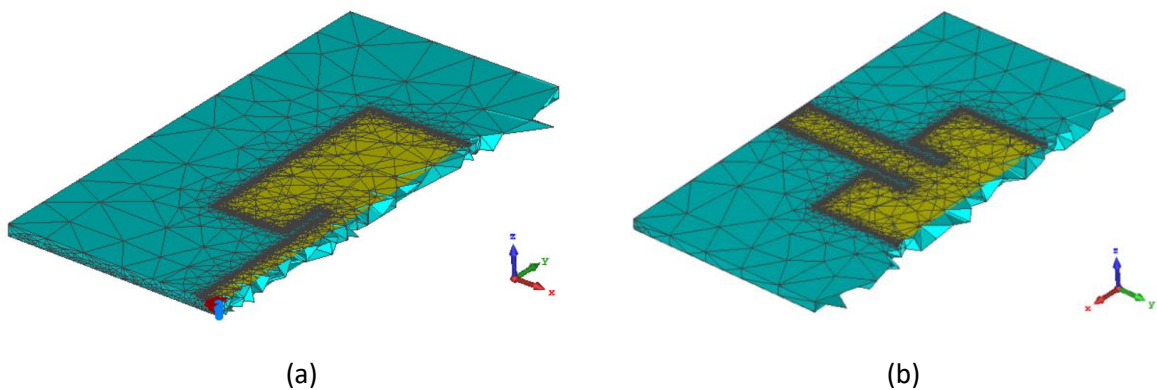
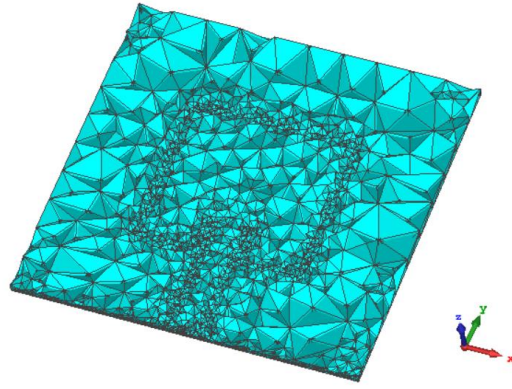


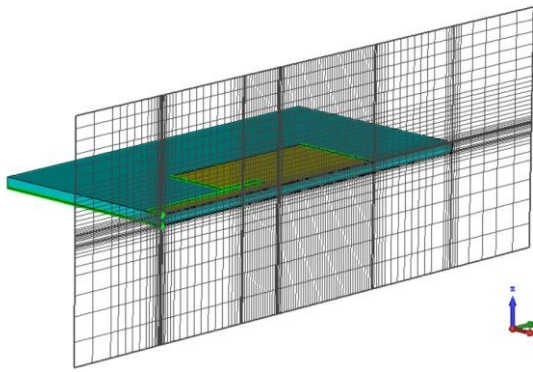
Figure 2.12: Rectangular patch antenna designed in CST Microwave Studio.



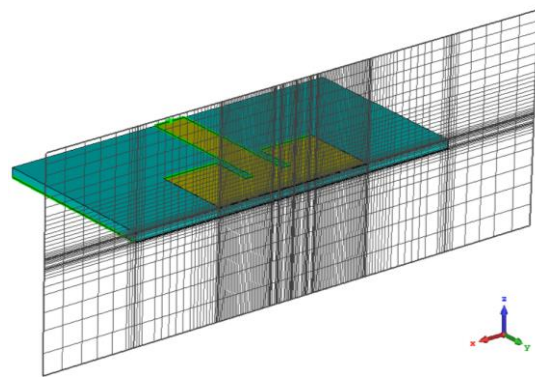


(c)

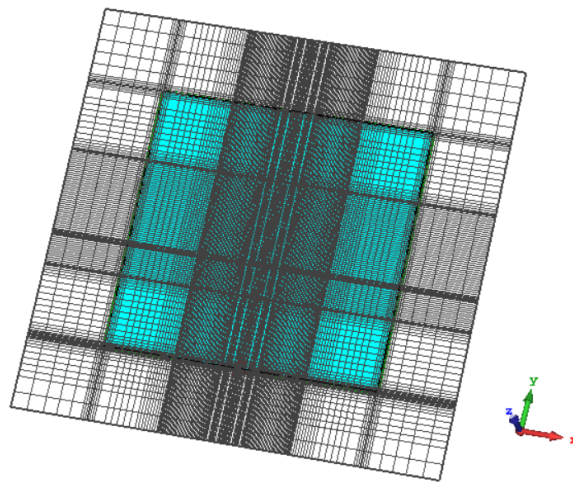
Figure 2.13: Tetrahedral mesh cross-section of a rectangular patch antenna along the (a) ZY plane (b) ZX plane and (c) YX plane.



(a)



(b)



(c)

Figure 2.14: Hexahedral mesh cross-section of a rectangular patch antenna along the (a) ZY plane (b) ZX plane and (c) YX plane.

Cross-sectional planes of the rectangular patch using a tetrahedral mesh are shown in Figure 2.13. The mesh refinement technique has been applied which has strategically

increased the mesh density triangulation at the surface as shown in Figure 2.13(a) and Figure 2.13(b) as well as the volume tetrahedrons shown in Figure 2.13(c) in areas where the antennas field concentrations are high. Figure 2.14 shows the discretization of the antenna model using a hexahedral mesh with the applied mesh refinement technique. The cross-sectional planes are shown in Figure 2.14(a), Figure 2.14(b) and Figure 2.14(c).

2.2.6 Frequency Optimisation

CST Microwave Studio consists of a variety of automatic optimizers which edits the geometry of the simulated structures and fulfils a defined goal. In this research, analytical methods allow the approximation of the resonance behaviour of the microstrip patch antenna. When transferring this information to the simulation model it produces a resonance which is not exact but near the intended design frequency. Using the built-in optimizers within CST can speed up the simulation analysis. For example, an antenna's simulated S-parameter having the least number of reflections at the resonant frequency can be regarded as a goal function for the optimizer. For this goal to be reached, the optimizer must produce a minimum and this minimum level is defined by the user and acts as a convergence criteria. In this example, the defined minimum is reached by altering certain geometric parameters that are sensitive to changing the magnitude of the resonant frequency. A local optimizer converges to the nearest or local minimum, while a global optimizer searches the entire problem space, for instance in the example above, this could be multiple resonances near the intended resonance. In CST there is a choice of a variety of global and local optimizers. The Trust Region Framework optimizer can be used to find either a local or global optimum of a function, but it is used to find a local optimum in the simulations described in this research. Whilst the trust region optimizer is regarded as CST's most robust local optimizer [15], it is chosen for its demonstrated fast goal convergence in this research.

A detailed example of the optimization process using the Trust Region Framework in CST is presented for a rectangular microstrip patch antenna. The patch antenna in Figure 2.12 was designed and simulated in CST to operate at 2.4 GHz. The dimensions of the patch antenna are initially approximated using the transmission line model [3]. Information on the transmission line model design equations can be found in Chapter 4. The front view geometry of the patch antenna is shown in Figure 2.15. The associated approximated dimensions using the Transmission Line Model is shown in Table 2.1.

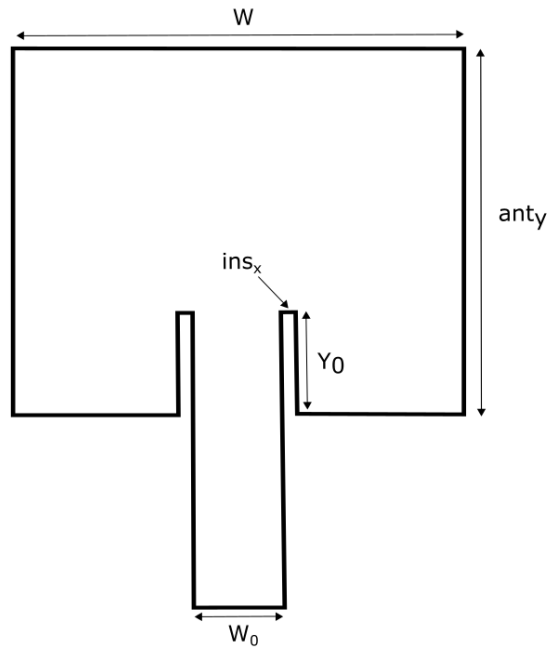


Figure 2.15: Front view geometry of the rectangular patch antenna.

Parameter	Symbol	Transmission Line Model Dimensions (mm)	Full-Wave Optimized Dimensions (mm)
Vertical Patch Length	ant_y	49.4	50.4
Aperture #1 Width	W	54.1	54.1
Inset Recessed Distance	y_0	17.8	17.8
Feedline Width	W_0	10.5	10.5
Substrate Height	H_s	2.8	2.8
Inset Feed Gap Width	ins_x	-	2.53

Table 2.1: Dimensions of the rectangular microstrip patch antenna approximated using the Transmission Line Model.

Using the approximated dimensions from the Transmission Line Model for the CAD model designed in CST Microwave Studio typically leads to a simulated S-parameter result with minimum reflection occurring around 2.4 GHz. To adjust the S-parameter result for a resonance at exactly 2.4 GHz the ‘Trust Region Framework’ local optimizer is used to refine the selected dimensions to adjust the antenna’s resonant frequency and minimize its magnitude at resonance.

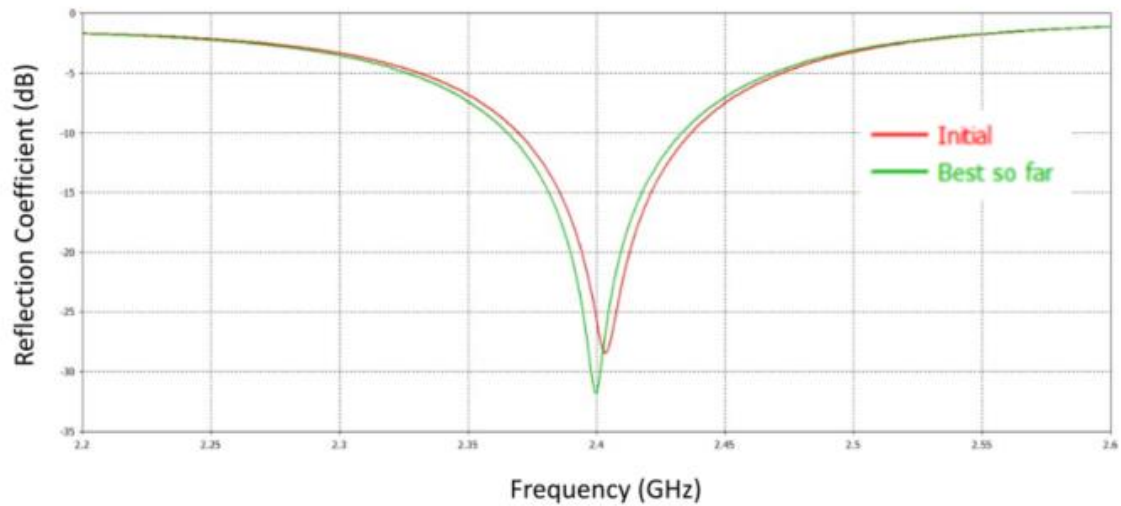


Figure 2.16: Simulated S-parameter result of the rectangular patch antenna using Trust Region Framework optimizer.

The optimization of the rectangular patch antenna's S-parameter result can be seen in Figure 2.16, where the 'Best so far' S-parameter result refers to the antenna with altered dimensions. Perfect convergence of the S-parameter result would occur when the magnitude of the linear S-parameter reaches zero at 2.4 GHz. The linear S-parameter tends towards zero for improvements in the antenna's impedance match. Since a value of zero is not realizable the simulation is manually stopped once the minimal point is at resonance and the magnitude is less than -20 dB which is equivalent to a VSWR of 1.22: 1.

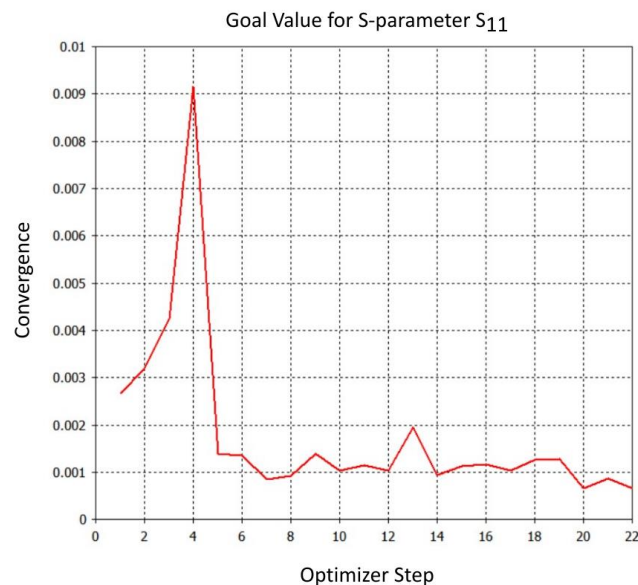
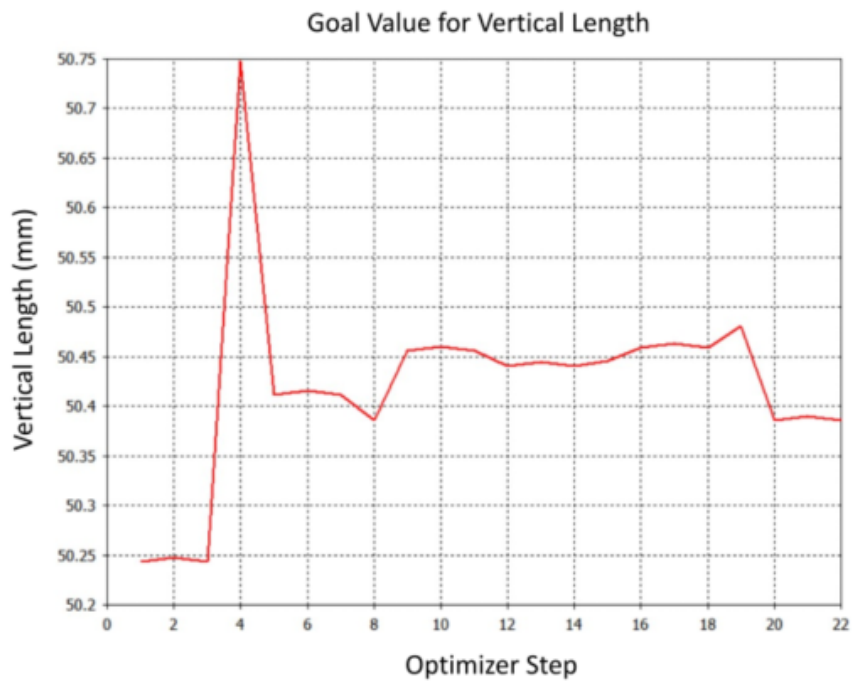


Figure 2.17: Simulation iterations of the rectangular patch antenna using the Trust Region Framework optimizer which is converging closer to the defined goal.



(a)



(b)

Figure 2.18: Optimization of the antenna's (a) vertical length and (b) inset feed gap width.

The Trust Region Framework starts by building a linear model on primary data in a 'trust' region around the starting point [12]. In this example the primary data is the vertical length

of the antenna and the gap width of the inset feed. The vertical length of the antenna controls the frequency at which the resonant occurs and the gap width of the inset feed allows fine tuning of the impedance to obtain a better match between the feedline and the radiating element. Optimizations are performed on the vertical length and gap width of the inset feed as shown in Figure 2.18 which accepts new values if the S-parameter result is superior to their initial or previous values. The optimized dimensions for the vertical length and the gap width of the inset feed are shown in Table 2.1.

2.3 References

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3 Historical Overview

This chapter provides an overview of the historical developments of microstrip patch antennas followed by a review of recent advances in wearable antennas.

3.1 Microstrip Patch Antennas

Microstrip antennas are conformal and low profile. They are "versatile in terms of resonant frequency, polarization, pattern, and impedance" [1]. They also have an advantage of being low cost, low weight, have design flexibility and ease of installation [2]. The low profile characteristics of microstrip antennas allows for their application in systems where aerodynamic profiles need to be maintained such as high-performance aircrafts, spacecrafts, satellites, and missiles [1].

The earliest design can be traced back to the now unclassified symposia proceeding "Microstrip Microwave Antennas" by Deschamps and Sichak in 1953 presented at the Third Symposium on the USAF Antenna Research and Development Program. In the paper, the antenna was presented as a microstrip line which transforms into a widening horn as the radiating element or a parallel plate region by widening the strip line [3].

Research on the antenna was silent up until the early 1970s, where it started to gain popularity in the antenna community. In 1972, Munson presented a microstrip wraparound antenna that produced omnidirectional coverage and showed that it was suitable for low profile conformal applications such as rockets, missiles and aircrafts [4][5]. During the same year, the NASA Langley Research Centre proposed an edge fed linear and circular polarized patch antenna as being an efficient radiator. A brief design procedure for the antennas was presented in [6]. However, one of the issues with the early patch antenna designs was that there was no theory to predict the antenna's input impedance. With the introduction of analytical methods applied to the patch antenna such as the cavity [7] and transmission line model [8], physical insight on the behaviour of the microstrip patch antenna was finally provided.

The 1980s saw major contributions to the advancement of microstrip patch antennas. The introduction of the integral equation technique [9] provided a powerful tool in microstrip patch antenna analysis. It allowed analysis for arbitrary geometries of the radiating element, modelling of feedlines, accurate predictions for the input impedance and radiation behaviour

in the presence of parasitic elements or other antennas. The decade also saw the introduction of microstrip patches being used in arrays to improve antenna gain as well as bandwidth [10].

One of the most notable issues with the microstrip patch antenna is its narrow bandwidth behaviour [1]. Some of the first attempts at increasing the bandwidth involve using a shunt tuning stub on a proximity coupled microstrip feed line [11]. By altering the dimensions of the stub, the bandwidth can be enhanced. Another method involved using parasitic patches next to the radiating edge of a patch antenna [12]. The gap coupled parasitic elements have slightly different dimensions as to not radiate at the same frequencies but near each other. [13] adopted the same technique but used metallic strips placed between the substrate or adjacent to the transmission line, while [14] stacked patches on top of each other.

Another significant issue with the microstrip patch antenna is surface wave propagation with substrates with high dielectric constants. Using a high permittivity substrate has been shown to reduce the dimensions of the antenna [15]. However, using such high permittivity substrates results in narrow bandwidth and surface wave effects that reduce the antennas efficiency and gain [16]. Such substrates also add manufacturing complexity which would not be suitable for low-cost consumer applications.

The 1990s saw advanced developments on computational numerical techniques such as the Finite Difference Time Domain method applied to antenna analysis. This enabled fast and accurate simulations of complex geometries with multiple dielectric interfaces [17]–[19]. The decade also saw the development of optimization techniques to reduce surface wave propagation as well as reducing the size of the antenna for lower frequencies. The use of parasitic elements loaded into the dielectric substrate was previously discussed for its bandwidth enhancement properties. It was also found that when strategically placed at the radiation edges it also mitigated surface waves by reflecting them back into the microstrip cavity [20].

Other methods investigated include the use of shorting posts and hollowed cores or substrates with different dielectric properties under the radiating element [21]. Their research was focused on a variant of the circular patch antenna. Using an annular ring for the radiating element it was found that a certain radius dimension will not excite the TM_0 surface wave. A hollow core slot within the dielectric substrate was attempted in [22] and was shown to be very effective in reducing surface waves when the patch antenna was placed on a

perforated Gallium Arsenide substrate for MMIC applications. The use of a shorting post for a circular and a aperture coupled patch antenna was shown to significantly reduce the overall size of the antenna for radiating at the same resonant frequency [23][24].

The 90s and 2000s saw a drastic increase for microstrip patch antenna applications such as satellite communication services, mobile base stations and mobile handsets [25]. This also resulted in an increase in optimization techniques to improving the antenna's bandwidth, radiation efficiency, reducing surface wave effects and the antenna's size for compact devices.

Such techniques include implementing a combination of the below:

- high permittivity substrates,
- Electromagnetic Bandgap (EBG) structures
- Artificial Magnetic Conductors (AMCs)
- Split Ring Resonators (SRRs)
- metallized vias
- varactor diodes
- geometry alteration to the radiating element

The late 90s saw the introduction of Electromagnetic Bandgap (EBG) structures into microstrip patch antennas. By implementing periodic structures within the dielectric substrate, it is possible to alter the dispersive properties in which certain electromagnetic modes have forbidden frequency regions, which cover all propagation directions. This in turn suppresses the surface waves within the substrate resulting in a reduction of backwards radiation. When the EBG structures are applied in patch antenna array elements, mutual coupling is shown to decrease while the radiation efficiency increases [26]–[31]. The combination of shorting pins and EBG structures was presented in [32] in order to miniaturize the patch as well as reducing the excitation of surface waves. Compared to an ordinary shorted patch antenna, incorporating the EBG ground plane was shown to increase the antenna's gain significantly from 1.2 dBi to 4.4 dBi.

Implementing shorting pins creates fabrication complexity. An alternative aperture-coupled patch antenna is presented in [33] without shorting pins in order to reduce surface waves. The ground plane is replaced by an EBG surface which is used to create a reflection

phase of 0° at the resonant frequency. This behaves like a perfect magnetic conductor reflecting the incident waves at the ground plane with the same phase as the radiating element. However, the design presented has large thickness, due to the dielectric spacer added to ensure the required distance between the aperture and the radiating element.

Another structure which exhibits unconventional electromagnetic properties is the split ring resonator (SRR) [34]. It is known to exhibit negative permeability, and was applied to microstrip antennas due to its subwavelength dimensions at resonance [35].

A variant of the SRR, the complementary split ring resonator (CSRR) etched onto the radiating element and ground plane of a microstrip patch antenna has been shown to produce multi-band operation and miniaturization [36] [37]. The designs proposed in [37] showed that etching a CSRR into the ground plane leads to a reduction in antenna gain due to the ground plane resonance.

To maintain a directive radiation pattern while achieving miniaturization for the patch antenna, loading a CSRR between the ground plane and radiating element was proposed in [38][39]. However, increasing the degree of miniaturization came at a cost by reducing the radiation efficiency.

More recently, altering the ground plane with defected ground structures (DGS) [40]–[42] and loading the patch with metamaterial structures [43]–[45] also provided miniaturization to a large degree while maintaining high radiation efficiency. However, defected ground structures are associated with bi-directional radiation patterns since there is resonance at the ground plane. Also, the metamaterial structures embedded to the substrate lead to a bandwidth reduction.

Another notable patch antenna design is presented in [46]. A U-shape is slotted into the radiating element with varactor diodes placed in the gaps. The varactor diodes allow tuning of the resonant frequency. Antenna miniaturization is achieved by placing metallized vias at the edge of the patch. However, the miniaturization came at a cost of reduced bandwidth as well as creating a complex antenna structure.

The previously discussed miniaturization techniques make the geometry of the patch antenna rather complex. A simpler approach of miniaturization is proposed in [47] where a trapezoidal patch antenna was evaluated using a full wave analysis as a compact alternative to the rectangular patch antenna. It was found that reducing the size of a radiating edge leads to a small reduction of bandwidth. However, the overall performance characteristics is almost

the same compared to a rectangular patch. To mitigate the slight reduction in bandwidth [48] and [49] enhanced the bandwidth of the trapezoidal patch by adding slots onto the radiating element.

3.2 Wearable Microstrip Patch Antennas

More recently there has been a growing interest in using microstrip patch antennas in wearable systems. With unique features such as flexibility and seamless integration, the wearable microstrip patch antenna has application in military, medical, space and emergency service applications [50]. Planar antennas such as the microstrip patch antenna are desirable since they can be placed on the body and radiate perpendicularly from the radiating element, with a ground plane efficiently shielding the human body [51]. Designing wearable antennas can be challenging since they need to be non-invasive, invisible, and robust when integrated into clothing [52].

Body worn antennas can be categorized as wearable antennas, textile antennas and SoftWear antennas [53].

- Wearable antenna - Is a part of clothing. It can either be rigid or flexible, integrated into a garment, a belt, a watch etc. The wearable antenna is optimised to perform near the human body.
- Textile antenna - Conductive fabrics make up the antenna and are integrated with conventional fabrics.
- SoftWear antenna - A combination of wearable and textile antenna properties. Typically a planar structure and flexible in two dimensions [53].

A SoftWear or hybrid textile copper microstrip patch antenna array is proposed in [54]. Since high design parameter accuracy is required to operate at the millimetre waves, the radiating element is designed using a non-textile adhesive copper sheet which is micromachined using a laser system. The patch consists of a cotton dielectric substrate and the 'ShieldIt Super' electro textile is used for the ground plane with the aim to obtain better flexibility. The application of electro-textile was limited only for the ground plane. The radiating element and feeding circuit were created using copper foil due to fabrication accuracy and reproducibility of the foil compared with the electro-textile.

Textile antennas can also be fabricated using non-woven conductive fabrics where the conductive fibres are bonded together by chemical, mechanical or heating processes. An example is the use of inkjet and screen-printing metals such as copper ink sprayed directly on the textile [52]. Copper antennas over time suffer from degradation due to oxidation. In [55] and [56] a low-cost solution for patch antennas is proposed by combining Polydimethylsiloxane (PDMS) which is a waterproof and weatherproof flexible silicone-based dielectric with meshed or perforated thin conducting copper sheets. The antenna can be encased in the proposed PDMS material.

With the use of digitally controlled sewing machines like the Entrepreneur Pro PR1000e [57] and the Tajima [58], complex antenna geometries can be created using conductive threads. However, these machines are not designed for use with conductive threads, therefore a careful selection of the conductive threads has to be made in order to withstand the high tension by the embroidery machine as well as maintaining flexibility and strength [59]. The use of Elektrisola-7 conductive threads is a proposed solution to these issues, providing lower embroidery tension, higher flexibility and geometric precision up to 0.1 mm [60].

Embroidering the conductive elements into the textile substrate allows for an integration that requires no adhesive layers and is considered to be the most suitable for being embedded into clothing [61], however it is known to be susceptible to fraying over time [52]. The electromagnetic properties of various electro-textiles were investigated in [62] by designing a microstrip patch antenna. The study showed that the positions of the conductive threads in the electro-textile is important to mitigate losses. If the conductive threads are placed at the bottom in the nonconductive binder threads as shown in Figure 3.1(a) indicated by the small circles across the electro-textile, the dielectric losses are minimized.



Figure 3.1: Cross sectional schematic of a satin weave fabric microstrip resonator for when (a) the conductive threads are placed underneath and (b) on top of the nonconductive threads [62].

Their research also showed that textiles which exhibit low permittivity and loss tangent values which make such fabrics acceptable for wearable antenna applications. However a careful

selection of the thickness of the dielectric fabric material is required since there is a trade-off between improved radiation efficiency and bandwidth [51]. Furthermore, conductive woven yarns intertwined cause high sheet resistance which results in a drastic decrease in the antenna's gain and radiation efficiency [63]. The findings by I. Locher et al. [63] showed that fabrics which were coated with a metallic layer post the weaving process have conductivity issues since the individual fibres are not completely coated at the overlapping warp and weft yarn points. This creates an inhomogeneous conductive path across the yarns and thus increases the sheet resistance.

Textile antennas have been shown to change their electromagnetic performance in the presence of moisture in the form of humidity and human sweat. M. Scarpello et al. proposed a breathable thermoplastic polyurethane coating for the textile inset-fed rectangular patch antenna which prevents moisture from penetrating the antenna and also allows washability to protect the components from corroding over time [64].

Simulating textile antennas accurately is challenging. Simplified metal grids are simulated in [65] using the full-wave electromagnetic simulation software HFSS based on the FEM. However, this model and simulation does not consider the overlaps that occur between the warp and weft yarns and the sheet resistance issues they raise. A more accurate method of simulating woven fabrics is required not only for higher frequencies but for different types of weaving patterns and stitch density since these are important variables to consider when improving gain and radiation efficiency of a textile antenna.

A major challenge for wearable antennas is deformation such as bending and crumpling. Several articles [54], [63], [74]–[77], [66]–[73] have studied its effect and have shown that it can influence the antennas gain, radiation pattern, bandwidth as well as the resonant frequency. The microstrip patch antenna is known for its narrow bandwidth behaviour [1]. For certain degrees of bending the antenna, the resonant frequency will detune completely where the bandwidth will not include the intended operational frequency anymore. A mitigation for this is proposed in [71] where an F and miniaturized E-shaped textile patch antenna is presented. The study investigated bending as well as non-symmetric crumpling conditions. Both deformations showed resonant frequency detuning. The miniaturized E-shape antenna is shown to be more resilient to bending while having a larger bandwidth. The measured results also showed that placement of the antenna on the arm or a flat area of the body can mitigate bending and crumpling effects. Other attempts to

miniaturize the radiating element for a wearable antenna is shown in [78][79] whereby slits are cut around the edges of the radiating element altering the current path. However, it was also reported that this type of shape modification can introduce out-of-phase radiating edges which can lead to higher dissipation losses [80].

Antenna bending, miniaturisation and bandwidth enhancement are challenges investigated in this thesis and are explored further in Chapter 4, 5 and 6. Due to the simulation complexity of modelling textiles in the patch antenna, in this analysis a simplification is made by modelling the structures as homogenous mediums, since Liu et al. [65] showed through measurements that if the distance between two woven conductive yarns is much smaller than the operating wavelength of the patch antenna, the structure can be modelled as a homogenous medium. This would suggest significant insights may not be gained from modelling the textile's fibres separately in simulations.

Table 3.1 highlights some textile patch antennas that have been previously developed for wearable applications.

Ref	Patch Antenna Configuration	f_{res} (GHz)	ϵ_r	$\tan \delta$	Antenna Size (mm ³)	RESA (mm ²)	Gain (dBi)	Bandwidth (%)	ϵ_0 (%)	Year
[74]	Textile Rectangular-Ring	2.45	1.25	-	78 x 73.5	2548	-	6.9	70	2006
[62]	Textile Rectangular Patch	2.44	1.9	0.0045	115 x 73 x 4	1358	6.6	6.2	78	2008
[64]	Inset-Fed Rectangular Patch	2.45	1.6	0.02	100 x 100 x 2.808	1922	-	-	41.4	2012
[77]	Dual Band Patch with AMC	2.45/5.20	1.2	0.044	100 x 100 x 3	2087	2.5/4	-	>40	2014
[75]	Textile Dual-Band Patch	2.40/5.20	1.3	0.044	50 x 20 x 6.34	200	3.8/6.2	5.51/13.7	-	2015
[66]	Textile Rectangular Patch	2.40	2.1	-	100 x 100 x 1.06	2050	-	1.25	-	2017
[81]	Aperture Coupled Textile Patch	2.45	1.35	0.044	100 x 100 x 3.34	1267	5.42	9.72	50.8	2023

Note: RESA = Radiating Element Surface Area

Table 3.1: Reported textile patch antennas in literature.

3.3 References

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4 Patch Antenna Miniaturization

The aim of antenna miniaturization is to reduce the size of the antenna while maintaining the performance of the antenna as quantified by parameters such as the operating frequency, bandwidth, gain and total radiation efficiency. In wearable applications, smaller antennas are more resilient to bending effects when placed conformally on a curved inhomogeneous structure such as on the human body. Miniaturization can be achieved by reducing a dimension, area or volume of the structure or substructures.

F. Bilotti et al. [1] showed that the Trapezoidal Patch Antenna (TPA) has shown to reduce the overall surface area of the radiating element while remaining to be a half-wavelength resonator and maintaining very close performance characteristics to that of a RPA. Whilst the largest dimension remains to be a half-wavelength, the TPA offers the possibility a surface area size reduction and performance comparable to that of a conventional RPA. The performance of the TPA is explored further in this section for a textile variant suitable for body-centric wireless communications systems that operate in the industrial, scientific, and medical (ISM) band at 2.4 GHz.

The first section of this chapter outlines antenna miniaturization techniques that have been applied to planar antennas in literature. The TPA which has been proposed as a technique to miniaturize the patch antenna's radiating element surface area [1]–[4] is investigated further for its suitability for wearable applications. The simplicity in the design, reduction in surface area and ease to implement on a fabric substrate makes the TPA a good candidate for textile wearable applications.

The characterization of the TPA is presented analytically through the Aperture Array Model (AAM) which has not been shown in previous literature. The focus of this analysis is to investigate the performance degradation of the antenna while converting a RPA to a TPA, in which the size of the antenna's radiating edge is gradually reduced. The key parameters investigated using the aperture array model are the radiation intensity as a far-field plot and maximum directivity.

Following on from the development of the TPA aperture array model, a full wave simulation analysis is performed on the TPA to investigate its far-field behaviour. Other key radiation performance parameters which the aperture array model cannot compute such as

the reflection coefficient, bandwidth, realized gain and the size reduction in the antennas radiating element surface area are also investigated.

4.1 Developments in Antenna Miniaturization

In order to characterize a size reduction of the antenna, the terms primary and secondary reduction are defined [5]. “Primary reduction implies that the antenna system has been reduced in size, either in length, area, or volume.” “Secondary reduction implies that a substructure of the antenna system has been reduced in size, while leaving adjacent structures unchanged (or nearly unchanged).”

The methods of patch antenna miniaturization can be categorised by either using high permittivity substrates, shorting pins/vias, editing the geometrical shape of the radiating patch or ground plane, integrating metamaterial structures within the substrate or a combination of these methods.

The miniaturization of the patch antenna can be achieved using high permittivity substrates since the reduction of the resonant frequency of the patch antenna is directly proportional to the square root of the permittivity of the substrate [6]. High permittivity ceramic substrates such as barium-titanate ($\epsilon_r = 37$) and neodymium-titanate ($\epsilon_r = 85$) implemented into the patch antenna have been investigated by A. Hoorfar et al. [7]. Their research focused on miniaturizing the antenna’s ground plane to show how it affected the far-field pattern. The patch antenna was designed to operate at 1.579 GHz with a length of 15.3 mm and width of 21.8 mm. The size of the ground plane was varied from 15×15 cm to 2.5×2.5 cm. Reducing the size of the ground plane increased the amount of edge diffraction and thus created larger back-lobe radiation, overall creating an omnidirectional radiation pattern. There was also a significant surface wave problem with the neodymium substrate which created a lot of ripples in the radiation pattern and reduced overall efficiency, therefore a larger ground plane was used to minimize the impact of the ripples.

A miniaturized variant of the microstrip patch antenna is the planar inverted-F antenna (PIFA) which is a quarter wavelength radiator. This variant is currently used in many handheld devices. However, the PIFA suffers from low radiation efficiency and reduced gain [8].

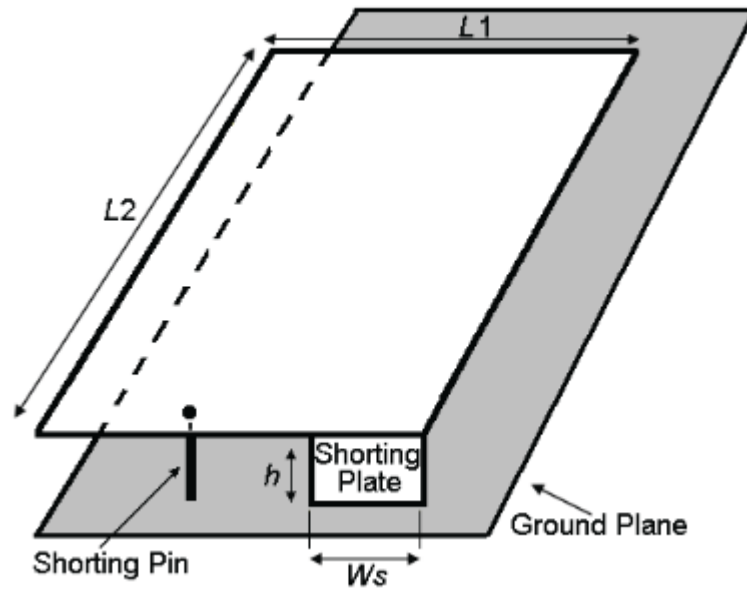


Figure 4.1: Geometry of the Planar Inverted-F Antenna (PIFA) [8].

An added shorting pin located near the feed of a patch antenna creates a strong capacitive coupling effect. The capacitive effect counteracts the inductive nature of the patch antenna below resonance. Therefore the size of the shorted radiating element can be reduced provided there is a significant amount of capacitive coupling [9]. Further miniaturization to the shorted patch antenna can be achieved by strategically etching notches on the radiating element around where the shorting post is located. Altering the dimensions of these notches allow tuning of the input impedance and further miniaturization of the patch due to the capacitance introduced by the additional fringe fields produced by the notches. Therefore, the resonant frequency of the patch antenna is reduced due to the additional capacitance which counteracts the inductive nature of the shorted radiating element [9]. The radiating element of the shorted patch was shown to be 75% smaller than a conventional microstrip patch antenna and 20% smaller than a conventional shorted patch antenna. The shorted patch dimensions given in Figure 4.2 for a resonance at 530 MHz are $L = 44.5$ mm, $W = 48.75$ mm, $L_f = 16$ mm, $W_f = 1$ mm, $L_{fi} = 11.5$ mm, $W_{fi} = 7$ mm, $L_1 = 1$ mm, $W_1 = 3$ mm, $L_2 = 2$ mm, $W_2 = 3$ mm, $x_{ps} = 0.5$ mm, $r_{os} = 1$ mm.

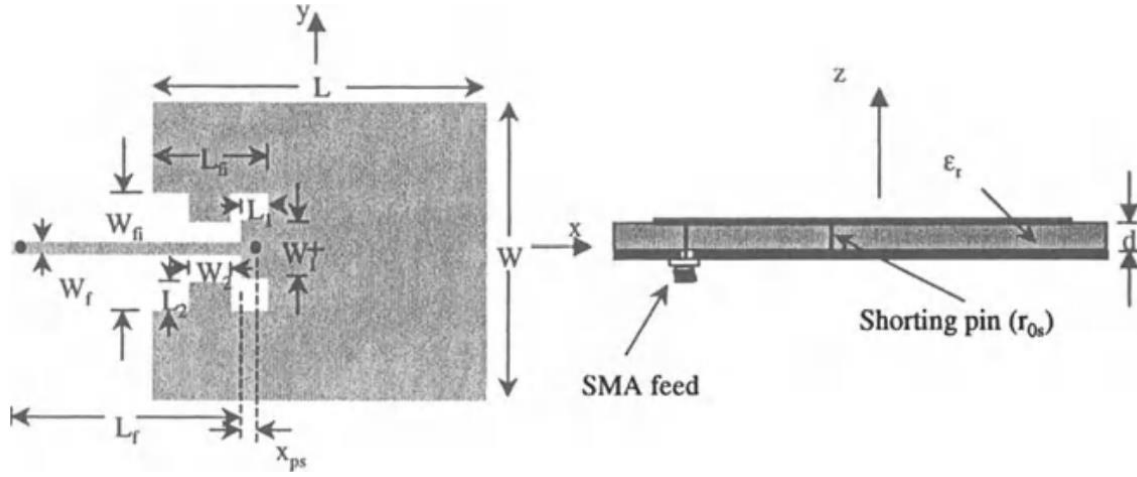


Figure 4.2: Schematic of an edge-fed shorted patch antenna where the dimensions for a resonance at 530 MHz given are $L = 44.5 \text{ mm}$, $W = 48.75 \text{ mm}$, $L_f = 16 \text{ mm}$, $W_f = 1 \text{ mm}$, $L_{fi} = 11.5 \text{ mm}$, $W_{fi} = 7 \text{ mm}$, $L_1 = 1 \text{ mm}$, $W_1 = 3 \text{ mm}$, $L_2 = 2 \text{ mm}$, $W_2 = 3 \text{ mm}$, $x_{ps} = 0.5 \text{ mm}$, $r_{0s} = 1 \text{ mm}$ [9].

Altering the radiating element geometry of the shorted patch antenna into a spiral has been shown to miniaturize the patch antenna to a large degree. The spiral geometry increases the current path within the area confined, creating an electrically larger structure. Careful consideration into how many bends the spiral geometry consists of was shown to maximize the degree of miniaturization [9].

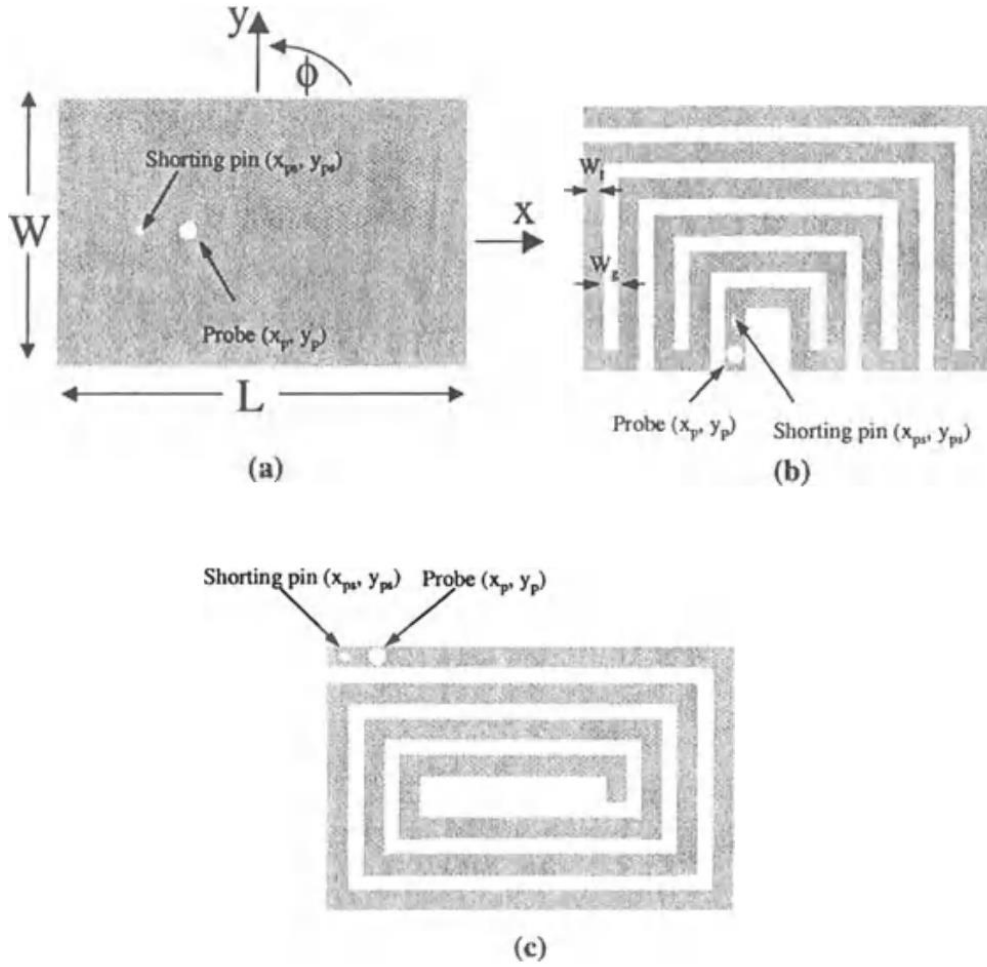


Figure 4.3: Schematic of the radiating element of two types of printed rectangular spiral patch antennas where (a) has a resonance of 1.9 GHz (b) has a resonance of 715 MHz and (c) 440 MHz all with dimensions $\epsilon_r = 1.07$, $d = 10$ mm, $L = 17$ mm, $W = 25$ mm, $w_t = 1.2$ mm, $w_g = 1$ mm, feed probe radius = 0.6 mm, shunting pin radius = 0.325 mm [9].

The designs from Figure 4.2 and Figure 4.3 suffer from very narrow bandwidths, low gain and omni-directional radiation patterns.

The introduction of metamaterial structures such as the Split Ring Resonator (SRR) [10], known for exhibiting unconventional electromagnetic properties such as negative permeability and subwavelength dimensions at resonance [11], has recently found its application in microstrip antenna design. A variant of the SRR, the Complementary Split Ring Resonator (CSRR) etched onto the radiating element and ground plane of a microstrip patch antenna has been shown to produce multi-band operation and miniaturization [12][13]. The designs proposed by V. Kumar et al. [13] showed that etching a CSRR into the ground plane leads to a reduction in antenna gain due to the ground plane resonance. To maintain a

directive radiation pattern while achieving miniaturization for the patch antenna, loading a CSRR between the ground plane and radiating element is proposed by R. O. Ouedraogo [6] and X. W. Dai et al. [14]. However, increasing the degree of miniaturization came at a cost by reducing the radiation efficiency.

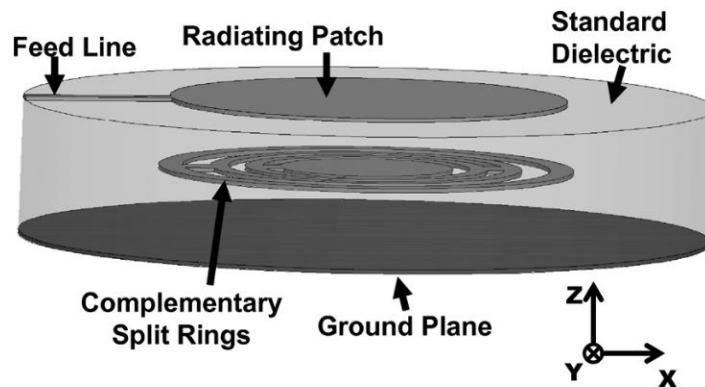


Figure 4.4: Miniaturized patch antenna with CSRR loaded into substrate [6].

More recently, altering the ground plane with defected ground structures (DGS) [15]–[17] and loading the patch with metamaterial structures [18]–[20] also provide miniaturization to a large degree while maintaining high radiation efficiency. However, defected ground structures are associated with bi-directional radiation patterns since there is resonance at the ground plane while the metamaterial structures embedded to the substrate suffer from a bandwidth reduction. Bi-directional radiation patterns are not ideal for antennas used in wearable applications since this is shown to increase the specific absorption rate (SAR) to unacceptable levels. Maintaining a specific separation distance between the skin and the antenna for such antennas results in acceptable SAR levels depending on the type of antenna [16][21].

Another notable miniaturized patch antenna design is presented by A. Boukarkar et al. [22]. A U-shape is slotted into the radiating element with varactor diodes placed in the gaps. The varactor diodes allow tuning of the resonant frequency. Antenna miniaturization is achieved by placing metallized vias at the edge of the patch. However, the miniaturization came at a cost of reduced bandwidth, gain as well as creating a complex antenna structure. Miniaturizing an antenna using components such as varactor diodes [22][23] and amplifiers [24] results in antenna miniaturization to a large degree but also suffer from reduced antenna gain.

The previously discussed primary miniaturization techniques make the geometry of the patch antenna rather complex or suffer from bandwidth and radiation efficiency degradation. Wheeler [25], Chu [26] and Harrington [27] found that there was a reciprocal relationship between an antenna's bandwidth and size. It was found that high radiation efficiency antennas require the antenna to be narrow band or rather have a high Q factor, while low radiation efficiency allows for broadband behaviour and thus a small Q factor. This relationship is known as the Chu-Harrington limit. Miniaturizing an antenna whether it is either from a primary or secondary reduction scales according to the Chu-Harrington limit. The size of the antenna needs to be carefully designed as to not degrade the antenna's radiation performance. A disadvantage of patch antennas are their low radiation efficiency and very narrow frequency bandwidth which is typically a few percent [28]. Effective miniaturization manages to maintain the bandwidth, radiation efficiency and radiation pattern of the antenna to that of the non-miniaturized antenna. The various miniaturized antenna's reported in Table 4.1 show that the greatest miniaturized results in a reduction in the antenna's gain, bandwidth or radiation efficiency.

A simpler approach for patch antenna miniaturization is proposed by F. Bilotti, C. Vegni and M. Pan et al. [1]–[3] where a trapezoidal patch antenna was analyzed and fabricated as a compact alternative to the rectangular patch antenna. It was found that reducing the size of a radiating edge leads to a small reduction of the antenna's bandwidth. F. Bilotti et al. [1] showed that altering the slope angle of the sides of the trapezoidal radiating element did not significantly change its resonant frequency. A slope angle of 90° corresponds to a RPA operating at 5.95 GHz. Decreasing the slope angle causes negligible change to the resonant frequency until a slope angle of 65° where the resonant frequency starts to increase upwards to approximately 6.15 GHz at a slope angle of 55° . However, the overall performance characteristics are almost the same compared to a rectangular patch. To mitigate the slight reduction in bandwidth, C. Vegni [2] and M. Pan et al. [3] enhanced the bandwidth of the trapezoidal patch by adding slots onto the radiating element. The trapezoidal patch antenna can be thought of as a 'secondary reduction' technique [5], where the radiating element size is decreased rather than all the components of the antenna.

As seen in the literature discussed in this section so far, patch antenna miniaturization has its pros and cons. When trying to improve one performance parameter it may come at a cost in reducing another. A summary of the antenna configurations, dimensions and

performance properties are presented in Table 4.1. For some of the miniaturized antennas reported, their bandwidths were not stated, however the graphs for the S_{11} parameters are shown to be extremely narrow over the resonant frequency being smaller than 20 MHz. From analysing Table 4.1, attempts to miniaturize the patch antenna using primary reduction methods lead to drastic degradation in the antennas gain.

Ref	Year	Antenna configuration	Antenna Size mm ³	Radiator Size mm ²	f_{res} (GHz)	ϵ_r	$\tan \delta$	BW (%)	e_0 (%)	Gain (dBi)
[7]	2001	Rectangular Patch	25.4 × 25.4 × 1.27	15.32 × 21.77	1.56	37	0.003	0.001	-	-
[7]	2001	Rectangular Patch	50.8 × 50.8 × 2.54	11 × 9.65	1.468	85	0.002	0.002	-	-
[9]	2003	Edge-Fed Shorted Patch	100 × 100 × 1.59	44.5 × 48.75	0.53	2.20	-	-	-	1
[9]	2003	Shorted Spiral Patch (Fig 4.3a)	40 × 40 × 10	25 × 17	0.715	1.07	-	-	-	0.3
[9]	2003	Shorted Spiral Patch (Fig 4.3b)	40 × 40 × 10	25 × 17	0.440	1.07	-	-	-	-
[12]	2012	Patch with CSRR	-	18 × 23	4.19 4.80	2.43	-	-	17.9 64.8	-0.11 5.85
[6]	2012	Patch with CSRR	R = 46.2 H = 2.34	R = 11.5	2.45	2.33	0.0012	0.004	28.1	-1.28
[14]	2013	Patch with CSRR	-	30 × 20	2.25 3.59 4.30	2.65	0.0017	-	-	-
[15]	2016	Patch with DGS	20 × 18.8 × 0.762	10 × 8.8	2.43 5.2	3.48	0.004	1.65 ^[1] 23 ^[1]	30 81	-1.7 2.4
[16]	2020	Monopole with EBG & DGS	60 × 60 × 2.4	20 × 16	2.4	1.05	0.0003	32.1	72.3	6.45
[29]	2019	CPW-fed Monopole with AMC	102 × 68 × 1.8	27 × 34	4.8	1.08	0.008	33.3	-	6.12
[18]	2021	CRLH antenna with AMC	25 × 25 × 0.1	19 × 7.8	2.45	2.9	0.0025	1.67	87	2.34
[19]	2015	CPW-fed Monopole with Metamaterial	20 × 37 × 0.2	4.8 × 8	2.60 6.22	3.7	0.025	29.3 6.9	93 78	-0.8 4.3
[21]	2020	Dual-band Monopole	19 × 12 × 0.508	4 × 10	2.45 5.85	2.2	0.0009	5.71 3.76	86 91	2 3.2
[22]	2017	Tunable Triple Band Patch	40 × 40 × 1.96	15 × 25	2.33 3.21 4.99	2.55	0.002	2.27 2.21 1.6	62.6 63.3 66.8	2.72 2.76 2.87
[1]	2001	Trapezoidal Patch	-	20.4 × 20	5.95	3.38	-	4.43	-	-
[3]	2000	Slotted Trapezoidal Patch	100 × 100 × 1.6	42 × 37.3	1.8	4.4	-	6.61	-	3

R: Radius of circular patch

H: height of circular patch

^[1] – 6dB Bandwidth

Table 4.1: Comparison of reported miniaturized planar antennas in literature.

4.2 Proposed Inset-Fed Trapezoidal Patch Antenna

The Trapezoidal Patch Antenna (TPA) is proposed as a design consideration for patch antenna applications in low-profile wearable applications since the decrease in the radiating elements size will be shown to outweigh the marginal decrease in the antenna's radiation performance. To begin, the TPA is modelled as a dual-aperture array since the apertures formed across the widths of a RPA account for most of the radiation [28]. This is to determine whether two radiating apertures are adequate to characterize the far-field behaviour, while ignoring the effects of the non-radiating edges located along the diagonal lengths of the trapezoid. The two radiating apertures are shown in Figure 4.5. The quad-aperture array model incorporates the previously ignored non-radiating edges and characterises them as partial radiators which contribute to the total electric field of the TPA. The dual and quad-aperture array models are compared to full-wave simulations to determine which characterization is more appropriate.

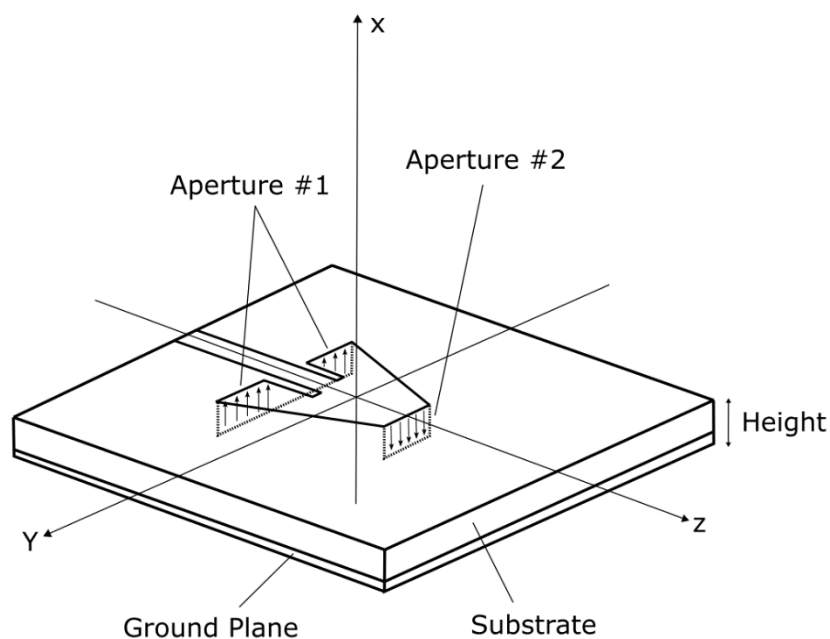


Figure 4.5: Trapezoidal Patch Antenna Geometry.

The TPA is designed to operate for wearable applications in the 2.4 GHz ISM band. The antenna model comprises of an inset-fed copper radiating patch, a cotton dielectric substrate and a copper ground plane. The antenna is fed by a $50\ \Omega$ microstrip line where the width of the line is 10.5 mm.

When a wave propagates down the microstrip line towards the radiating element

there is a change of width where the strip line meets the first radiating slot as shown where Aperture #1 is located in Figure 4.5. The wave will continue to travel to the top end of the patch where the second radiating slot is. A large reflection occurs at this discontinuity and the wave reflects along the length of patch creating fringing radiation at both ends of the antenna [30]. The fields radiated by both slots produce a resultant effect in the far-field.

C. Balanis [28] gives an interpretation of the radiation mechanism of the patch antenna in terms of charge distribution and current density. Energizing the patch results in a charge distribution located above and below the radiating element as well as on top of the ground plane as shown in Figure 4.6. The current density $\mathbf{J}_t$ and $\mathbf{J}_b$ gives rise to the the fringe fields. The like charges above and below the radiating element cause a repulsion which pushes charge off the edge giving rise to the fringing fields. At the same time there is an attractive mechanism between the charges located below that patch and on top of the ground plane. This attraction is the dominant mechanism and causes a large concentration of charge located just above the ground plane.

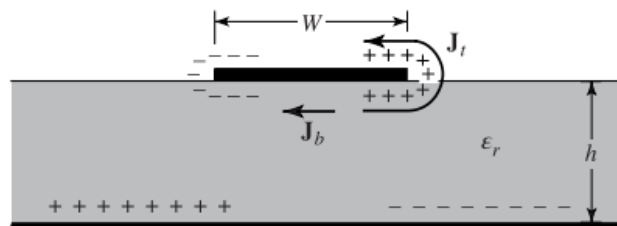


Figure 4.6: Charge distribution and current density formed on the microstrip patch antenna [28].

4.3 Aperture Array Model

Using this model, a radiation far-field characteristic comparison will be made between the RPA and TPA. The apertures of the patch antenna can be seen in Figure 4.5 where aperture #1 has a width of W_1 and aperture #2 has a width of W_2 . For a RPA the aperture widths correspond to when $W_1 = W_2$, and for a TPA $W_2 \ll W_1$.

To begin, some background theory of a radiating aperture is presented. Then a radiated power analysis is presented to predict the antenna's far-field pattern and directivity. The far-field distance r is given by a commonly used minimum distance observation criterion of $2d^2/\lambda$ [28]. Where d is the maximum dimension of the patch antenna and λ is the operating wavelength.

4.3.1 Theory and Analysis of an Aperture Antenna

To approximate the fields radiated by an aperture at the far-field, it is first required to define a source within the aperture, which is represented the current densities $\mathbf{J}_1$ and $\mathbf{M}_1$. Using the ‘field equivalence principle’, the fields that are radiated outside of the closed surface aperture can be modelled.

To model the fields that are outside of the aperture, an equivalent problem is formed. In the equivalent problem as shown in Figure 4.7(b), the original sources $\mathbf{J}_1$ and $\mathbf{M}_1$ are removed and it is assumed that there exist fields $\mathbf{E}$ and $\mathbf{H}$ inside the surface, and fields $\mathbf{E}_1$ and $\mathbf{H}_1$ outside of the surface. For these fields to exist within and outside the surface S , they must satisfy the boundary conditions on the tangential electric and magnetic field components [28].

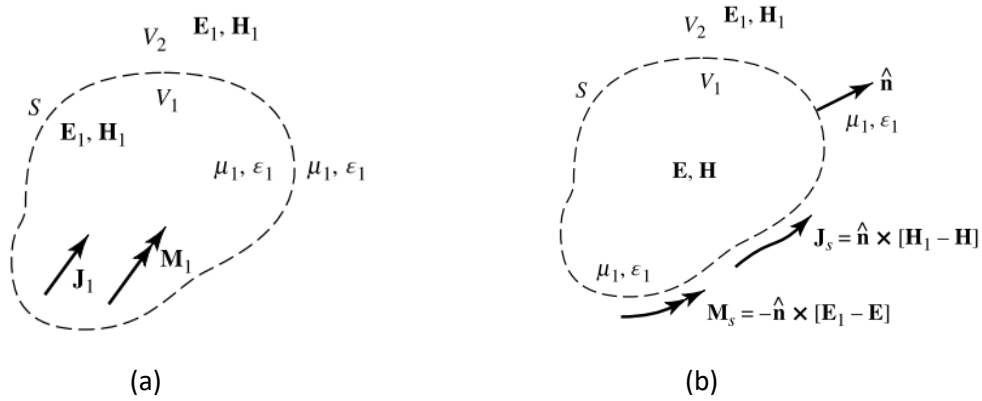


Figure 4.7: The fields, current densities, and material properties for the (a) actual and (b) equivalent aperture model [28].

Therefore, on the surface S there must exist the equivalent sources which radiate into free space as

$$\mathbf{J}_s = \hat{\mathbf{n}} \times [\mathbf{H}_1 - \mathbf{H}] \quad (4.1)$$

$$\mathbf{M}_s = -\hat{\mathbf{n}} \times [\mathbf{E}_1 - \mathbf{E}] \quad (4.2)$$

Equivalent problems can be found only requiring the magnetic or electric current densities. The fields $\mathbf{E}$ and $\mathbf{H}$ within the closed surface can be assumed to be zero since they are not of interest. This simplifies the current densities to

$$\mathbf{J}_s = \hat{\mathbf{n}} \times \mathbf{H}_1 \quad (4.3)$$

$$\mathbf{M}_s = -\hat{\mathbf{n}} \times \mathbf{E}_1 \quad (4.4)$$

This simplification of the field is known as Love’s Equivalence Principle [28]. In which a null field is produced within the surface.

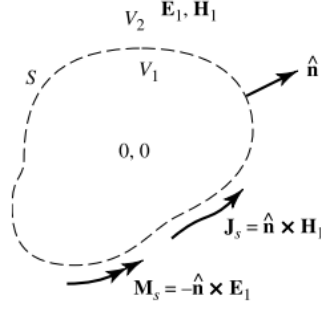


Figure 4.8: Love's equivalent model [28].

To find the magnetic and electric fields generated by an electric current source $\mathbf{J}$ and a magnetic current $\mathbf{M}$, the inhomogeneous vector wave equations in the form of the magnetic and electric auxiliary potential functions $\mathbf{A}$ and $\mathbf{F}$ are required to be solved, which are given by

$$\nabla^2 \mathbf{A} + k^2 \mathbf{A} = -\mu \mathbf{J} \quad (4.5)$$

$$\nabla^2 \mathbf{F} + k^2 \mathbf{F} = -\varepsilon \mathbf{M} \quad (4.6)$$

The magnetic auxiliary potential function $\mathbf{A}$ for an electric source $\mathbf{J}$ is related to the electric $\mathbf{E}$ and magnetic $\mathbf{H}$ fields by

$$\mathbf{E} \approx -j\omega \mathbf{A} \quad (4.7)$$

$$\mathbf{H} \approx -j \frac{\omega}{\eta} \hat{\mathbf{a}}_r \times \mathbf{A} \quad (4.8)$$

For a magnetic source $\mathbf{M}$ they are related to the electric $\mathbf{E}$ and magnetic $\mathbf{H}$ fields by

$$\mathbf{H} \approx -j\omega \mathbf{F} \quad (4.9)$$

$$\mathbf{E} \approx -j\omega\eta \hat{\mathbf{a}}_r \times \mathbf{F} \quad (4.10)$$

The vector wave equations (4.5) and (4.6) have solutions in the form of

$$\mathbf{A} = \frac{\mu}{4\pi} \iiint_V \mathbf{J} \frac{e^{-jkR}}{R} dv' \quad (4.11)$$

and

$$\mathbf{F} = \frac{\varepsilon}{4\pi} \iiint_V \mathbf{M} \frac{e^{-jkR}}{R} dv' \quad (4.12)$$

Where v' refers to the volume of the source, $k^2 = \omega^2 \mu \varepsilon$ and R is the distance from a point in the source to the observation point.

If the electric $\mathbf{J}$ or magnetic $\mathbf{M}$ current source represent linear densities the solutions to the wave equation can reduce to surface integrals as

$$\mathbf{A} = \frac{\mu}{4\pi} \iint_S \mathbf{J} \frac{e^{-jkR}}{R} ds' \quad (4.13)$$

$$\mathbf{F} = \frac{\varepsilon}{4\pi} \iint_S \mathbf{M} \frac{e^{-jkR}}{R} ds' \quad (4.14)$$

If the sources are electric $\mathbf{I}_e$ or magnetic $\mathbf{I}_m$ currents the solution can be reduced to line integral forms as

$$\mathbf{A} = \frac{\mu}{4\pi} \int_C \mathbf{I}_m \frac{e^{-jkR}}{R} dl' \quad (4.15)$$

$$\mathbf{F} = \frac{\varepsilon}{4\pi} \int_C \mathbf{I}_e \frac{e^{-jkR}}{R} dl' \quad (4.16)$$

The far-field observation point R can be approximated for phase variations by

$$R \approx r - r' \cos\psi \quad (4.17)$$

For amplitude variations it is approximated by

$$R \approx r \quad (4.18)$$

Where r' is the distance occupied by the source and ψ is the angle between r and r' as shown in Figure 4.9. Using the solution where the auxiliary potentials current source is represented as a linear density, the far-field equations for the magnetic and electric auxiliary potential function can be formed.

Assuming the current density resides on the surface of the source, $\mathbf{A}$ and $\mathbf{F}$ will take the form

$$\mathbf{A} = \frac{\mu}{4\pi} \iint_S \mathbf{J} \frac{e^{-jkR}}{R} ds' \approx \frac{\mu e^{-j\beta r}}{4\pi r} \mathbf{N} \quad (4.19)$$

$$\mathbf{F} = \frac{\varepsilon}{4\pi} \iint_S \mathbf{M} \frac{e^{-jkR}}{R} ds' \approx \frac{\varepsilon e^{-j\beta r}}{4\pi r} \mathbf{L} \quad (4.20)$$

Where $\mathbf{N}$ and $\mathbf{L}$ are known as the space factors, which determines the electric and magnetic current densities along the source observed at the farfield.

$$\mathbf{N} = \iint_S \mathbf{J}_s e^{j\beta r' \cos\psi} ds' \quad (4.21)$$

$$\mathbf{L} = \iint_S \mathbf{M}_s e^{j\beta r' \cos\psi} ds' \quad (4.22)$$

Expanding this in a rectangular coordinate system yields

$$\mathbf{N} = \iint_S (\hat{\mathbf{a}}_x J_x + \hat{\mathbf{a}}_y J_y + \hat{\mathbf{a}}_z J_z) e^{j\beta r' \cos\psi} ds' \quad (4.23)$$

$$\mathbf{L} = \iint_S (\hat{\mathbf{a}}_x M_x + \hat{\mathbf{a}}_y M_y + \hat{\mathbf{a}}_z M_z) e^{j\beta r' \cos\psi} ds' \quad (4.24)$$

The product of the element factor and the space factor give the total radiated field. The element factor refers to the term outside of the brackets in (4.34), (4.35), (4.37) and (4.38) along with any term that can be factored out of the integral for (4.39-4.42) [28].

In the far-field region only the θ and ϕ components are dominant while the radial component is assumed to be negligible.

The electric fields in terms of their auxiliary potentials can be expressed as

$$(E_A)_\theta \approx -j\omega A_\theta \quad (4.25)$$

$$(E_A)_\phi \approx -j\omega A_\phi \quad (4.26)$$

$$(E_F)_\theta \approx -j\omega\eta F_\theta \quad (4.27)$$

$$(E_F)_\phi \approx j\omega\eta F_\phi \quad (4.28)$$

and for the magnetic fields

$$(H_F)_\theta \approx -j\omega F_\theta \quad (4.29)$$

$$(H_F)_\phi \approx -j\omega F_\phi \quad (4.30)$$

$$(H_A)_\theta \approx j\omega \frac{A_\phi}{\eta} \quad (4.31)$$

$$(H_A)_\phi \approx -j\omega \frac{A_\theta}{\eta} \quad (4.32)$$

Thus, the total electric $\mathbf{E}$ and magnetic $\mathbf{H}$ fields can be written as the superposition between the electric and magnetic auxiliary potential fields due to an electric $\mathbf{J}$ and magnetic $\mathbf{M}$ current density source.

$$E_r \approx 0 \quad (4.33)$$

$$E_\theta \approx -\frac{jke^{-jkr}}{4\pi r} (L_\phi + \eta N_\theta) \quad (4.34)$$

$$E_\phi \approx \frac{jke^{-jkr}}{4\pi r} (L_\theta - \eta N_\phi) \quad (4.35)$$

$$H_r \approx 0 \quad (4.36)$$

$$H_\theta \approx -\frac{jke^{-jkr}}{4\pi r} \left(N_\theta - \frac{L_\theta}{\eta} \right) \quad (4.37)$$

$$H_\phi \approx -\frac{jke^{-jkr}}{4\pi r} \left(N_\phi + \frac{L_\phi}{\eta} \right) \quad (4.38)$$

L_ϕ , L_θ , N_θ and N_ϕ can be obtained by transforming (4.23) and (4.24) to a spherical coordinate system. Doing so yields

$$N_\theta = \iint_S [J_x \cos \theta \cos \phi + J_y \cos \theta \sin \phi - J_z \sin \theta] e^{jkr' \cos \psi} ds' \quad (4.39)$$

$$N_\phi = \iint_S [-J_x \sin \phi + J_y \cos \phi] e^{jkr' \cos \psi} ds' \quad (4.40)$$

$$L_\theta = \iint_S [M_x \cos \theta \cos \phi + M_y \cos \theta \sin \phi - M_z \sin \theta] e^{jkr' \cos \psi} ds' \quad (4.41)$$

$$L_\phi = \iint_S [-M_x \sin \phi + M_y \cos \phi] e^{jkr' \cos \psi} ds' \quad (4.42)$$

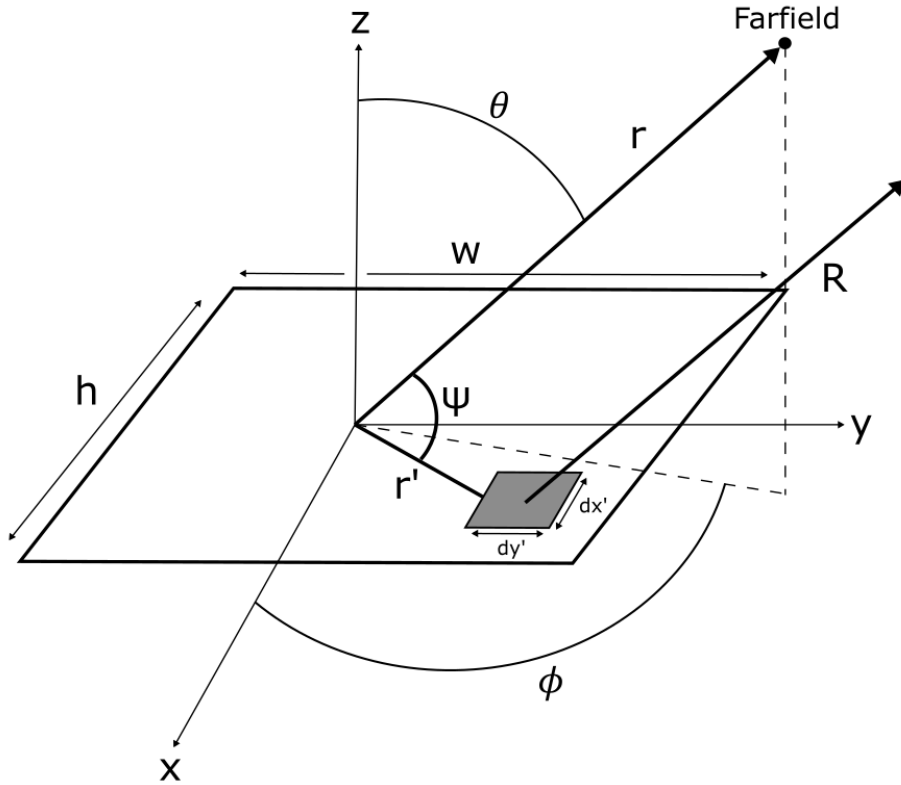


Figure 4.9: Far-field equivalent diagram for an aperture.

For a rectangular aperture the far-field equivalent diagram is shown in Figure 4.9 which gives a visual representation of the rectangular and spherical coordinate system that is used to analyse the fields radiated from the aperture. In the xy plane the current and magnetic densities J_x, J_y, M_x, M_y will be non-zero.

For this analysis, the electric field across the aperture E_a is modelled as a uniform field, thus

$$E_a = \begin{cases} \hat{a}_x E_0, & -h \leq x \leq h, -\frac{W}{2} \leq y \leq \frac{W}{2} \\ 0, & \text{elsewhere} \end{cases} \quad (4.43)$$

Since the aim of this analysis is to apply it to a patch antenna model in the later section with a ground plane, the equivalent sources are in the form

$$\mathbf{M}_s = -2(\hat{a}_z \times \hat{a}_x)E_0 = +2\hat{a}_y E_0 \quad (4.44)$$

$$\mathbf{J}_s = 0 \quad (4.45)$$

Where the magnetic current density $\mathbf{M}_s$ is doubled to account for the presence of the ground plane by the image theory [30]. The electric current density $\mathbf{J}_s$ on the radiating patch and the ground plane are very small and thus assumed to be negligible. If the apertures were modelled with an electric current density source $\mathbf{J}_s$, the reflected fields from the ground plane (PEC) would have opposite phase to the incident waves from the aperture and thus destructively interfere with each other. Thus, the fields or space waves from the apertures in the presence of a ground plane are predominantly produced by the magnetic currents $\mathbf{M}_s$. This is shown in Figure 4.10 where (b) represents the image source due to the ground plane and (c) is the combination of both magnetic sources.

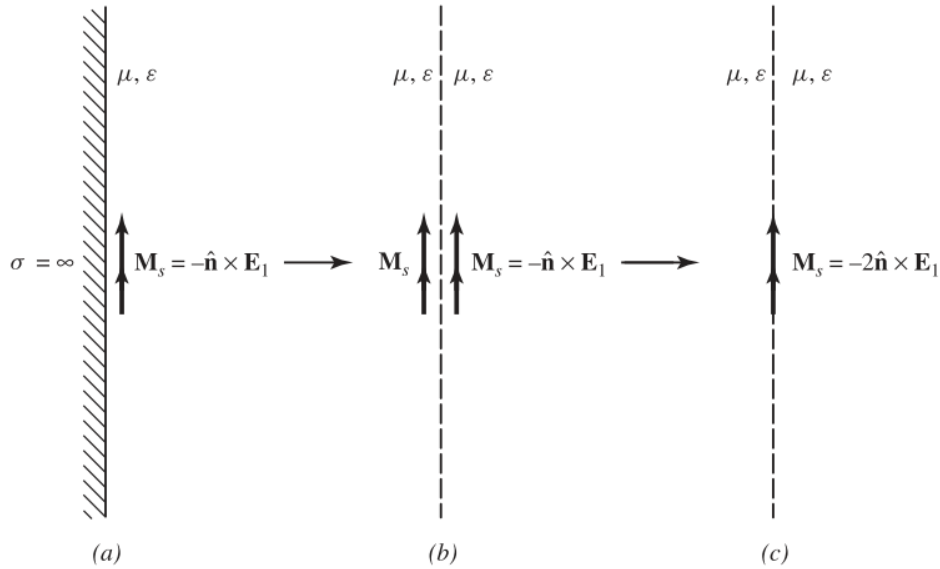


Figure 4.10: Equivalent models for a magnetic source near a ground plane or a perfect electric conductor [31].

For a magnetic source $\mathbf{M}_s$, the space factor components $N_\phi = N_\theta = 0$. Therefore only L_ϕ and L_θ need to be computed to express the electric and magnetic field components $E_\theta, E_\phi, H_\theta$ and H_ϕ . L_θ from (4.41) can be simplified to

$$L_\theta = \cos \theta \sin \phi \iint_S M_y e^{jkr' \cos \psi} dx' dy' \quad (4.46)$$

The differential path of $r' \cos \psi$ from the origin of the aperture to the farfield is

$$r' \cos \psi = \mathbf{r}' \cdot \hat{\mathbf{a}}_r = x' \sin \theta \cos \phi + y' \sin \theta \sin \phi \quad (4.47)$$

Therefore,

$$L_\theta = \cos \theta \sin \phi \int_{-W/2}^{W/2} \int_{-h/2}^{h/2} M_y e^{jk(x' \sin \theta \cos \phi + y' \sin \theta \sin \phi)} dx' dy' \quad (4.48)$$

$$L_\theta = 2hWE_0 [\cos \theta \sin \phi \text{sinc } X \text{sinc } Y] \quad (4.49)$$

where

$$Y = \frac{kh}{2} \sin \theta \cos \phi \quad (4.50)$$

$$X = \frac{kW}{2} \sin \theta \sin \phi \quad (4.51)$$

L_ϕ is calculated as

$$L_\phi = \cos \phi \int_{-W/2}^{W/2} \int_{-h/2}^{h/2} M_y e^{jk(x' \sin \theta \cos \phi + y' \sin \theta \sin \phi)} dx' dy' \quad (4.52)$$

$$L_\phi = 2hWE_0 [\cos \phi \text{sinc } X \text{sinc } Y] \quad (4.53)$$

Therefore, the fields radiated at the far-field can be represented as

$$E_\theta = -j \frac{hWkE_0 e^{-jkr}}{2\pi r} [\cos \phi \text{sinc } X \text{sinc } Y] \quad (4.54)$$

$$E_\phi = j \frac{hWkE_0 e^{-jkr}}{2\pi r} [\cos \theta \sin \phi \text{sinc } X \text{sinc } Y] \quad (4.55)$$

$$H_\theta = -\frac{E_\phi}{\eta} \quad (4.56)$$

$$H_\phi = \frac{E_\theta}{\eta} \quad (4.57)$$

In this section the analytical foundations for analysing the fields that are radiated by an aperture antenna are formed. This forms the foundations for the next section where the patch antenna is modelled as a dual-aperture array to approximate the far-field behaviour.

4.3.2 Aperture Array Model for the Trapezoidal Patch Antenna

Assuming the total field across the apertures is uniform from (4.43), the equivalent sources are in the form

$$\mathbf{M}_s = 2\hat{a}_y \times E_0 \quad (4.58)$$

$$\mathbf{J}_s = 0 \quad (4.59)$$

As before, the magnetic current density $\mathbf{M}_s$ is doubled to account for the presence of the ground plane by the image theory. The electric current density $\mathbf{J}_s$ on the radiating patch and

the ground plane are very small and thus assumed to be negligible. The electric and magnetic field components are given by (4.54-4.57).

Since the patch is split into 2 radiating apertures, the overall field of the patch for both θ and ϕ components are given by the superposition of the two radiating slots at the far-field as

$$E_{\Sigma\theta} = E_{\theta_1}^{W_1} + E_{\theta_2}^{W_2} \quad (4.60)$$

$$E_{\Sigma\phi} = E_{\phi_1}^{W_1} + E_{\phi_2}^{W_2} \quad (4.61)$$

Figure 4.11 outlines the geometry of the aperture array and the parameters used to model the combined fields of both apertures at the far-field. The radial distance R_1 and R_2 are defined as the distance from aperture 1 of width W_1 and aperture 2 of width W_2 to the far-field observation point. Both apertures are symmetrically positioned from the origin with a length of $\pm L/2$ along the z-axis.

The radial distance R_1 from aperture 1 is calculated as

$$R_1 = \sqrt{x^2 + y^2 + \left(z - \frac{L}{2}\right)^2} \quad (4.62)$$

Using the binomial expansion on the z term to the first 2 terms, and substituting $z = r \cos \theta$, the radial distance R_1 can be simplified to

$$R_1 \approx r - \frac{L}{2} \cos \theta \quad (4.63)$$

Repeating this procedure for aperture 2, the radial distance R_2 becomes

$$R_2 \approx r + \frac{L}{2} \cos \theta \quad (4.64)$$

Dielectric effects are ignored in the far-field analysis. A dielectric material can be incorporated into the aperture array analysis using the analysis presented by C. Knop et al. [32]. However, the far-field pattern is minimally affected when the height h of the dielectric-covered aperture is thin ($h/\lambda = 0.1$) or medium-sized ($h/\lambda = 0.4$). Therefore, is ignored in this analysis.

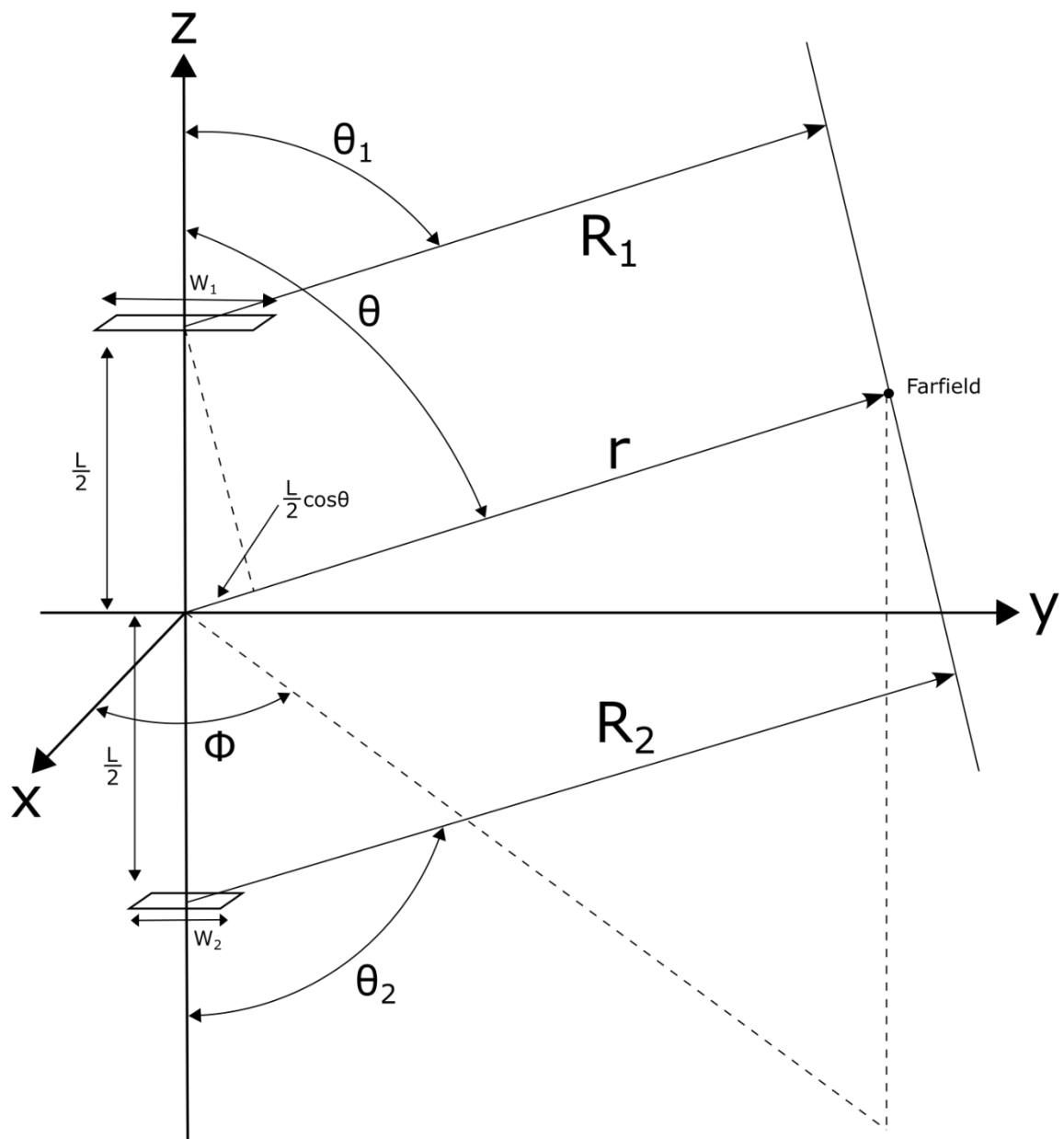


Figure 4.11: Dual aperture array geometry and far-field approximations.

Based on the geometry of Figure 4.11, by using (4.58), (4.34) and (4.35) the electric field components for one of the radiating apertures of the TPA can be solved as

$$E_{\theta} = \frac{-jk e^{-jkr}}{4\pi r} \cos \phi \int_{-\frac{W}{2}}^{\frac{W}{2}} \int_{-\frac{h}{2}}^{\frac{h}{2}} 2E_0 e^{jk(x' \sin \theta \cos \phi + y' \sin \theta \sin \phi)} dx' dy' \quad (4.65)$$

$$E_{\theta} = \frac{-jk}{4\pi} 2hE_0 \cos \phi \operatorname{sinc}\left(\frac{1}{2}kh \sin \theta \cos \phi\right) W \operatorname{sinc}\left(\frac{1}{2}kW \sin \theta \cos \phi\right) \frac{e^{-jkr}}{r} \quad (4.66)$$

$$E_{\phi} = \frac{jk e^{-jkr}}{4\pi r} \cos \theta \sin \phi \int_{-\frac{W}{2}}^{\frac{W}{2}} \int_{-\frac{h}{2}}^{\frac{h}{2}} 2E_0 e^{jk(x' \sin \theta \cos \phi + y' \sin \theta \sin \phi)} dx' dy' \quad (4.67)$$

$$E_{\phi} = \frac{jk}{4\pi} 2hE_0 \cos \theta \sin \phi \operatorname{sinc}\left(\frac{1}{2}kh \sin \theta \cos \phi\right) W \operatorname{sinc}\left(\frac{1}{2}kW \sin \theta \sin \phi\right) \frac{e^{-jkr}}{r} \quad (4.68)$$

To account for the radiation produced from each aperture, the array factor term is computed by substituting (4.63) and (4.64) only in the phase term which is e^{-jkr} . Using both (4.54) and (4.55) and approximating the amplitude terms as $\frac{1}{R_1} \approx \frac{1}{R_2} \approx \frac{1}{r}$, the total E-field contribution from both apertures $E_{\Sigma \theta}$ is given by

$$\begin{aligned} E_{\Sigma \theta}^{W_{12}} = & \frac{-jkhE_0}{2\pi} \cos \phi \operatorname{sinc}\left(\frac{1}{2}kh \sin \theta \cos \phi\right) \\ & \left(W_1 \operatorname{sinc}\left(\frac{1}{2}kW_1 \sin \theta \sin \phi\right) e^{\frac{jkL}{2} \cos \theta} \right. \\ & \left. + W_2 \operatorname{sinc}\left(\frac{1}{2}kW_2 \sin \theta \sin \phi\right) e^{\frac{-jkL}{2} \cos \theta} \right) \frac{e^{-jkr}}{r} \end{aligned} \quad (4.69)$$

and for $E_{\Sigma \phi}$ is given by (4.61) as

$$\begin{aligned} E_{\Sigma \phi}^{W_{12}} = & \frac{jkhE_0}{2\pi} \cos \theta \sin \phi \operatorname{sinc}\left(\frac{1}{2}kh \sin \theta \cos \phi\right) \\ & \left(W_1 \operatorname{sinc}\left(\frac{1}{2}kW_1 \sin \theta \sin \phi\right) e^{\frac{jkL}{2} \cos \theta} \right. \\ & \left. + W_2 \operatorname{sinc}\left(\frac{1}{2}kW_2 \sin \theta \sin \phi\right) e^{\frac{-jkL}{2} \cos \theta} \right) \frac{e^{-jkr}}{r} \end{aligned} \quad (4.70)$$

Since the height of patch is much smaller than the wavelength the $\operatorname{sinc}\left(\frac{1}{2}kh \sin \theta \cos \phi\right)$ term can vanish.

To characterize the far-field behaviour, the average power density W_{rad} , radiation intensity $U(\theta, \phi)$, radiated power P_{rad} and Directivity D need to be computed and are defined in Chapter 2. Calculations start once a suitable aperture width W , separation distance L between the apertures and resonant frequency f_r is defined.

To calculate the aperture width W and separation length L between the apertures, the dielectric constant of the patch antenna's substrate as well as the height of the substrate must be known prior. To study how well the proposed antenna works for wearable textile applications, the substrate is modelled with the dielectric properties of a textile.

The textile dielectric substrate is modelled as a homogenous medium. The dielectric characteristics of the substrate are based on the cotton/polyester sample in [33] which is modelled as a homogenous medium. The substrate is 2.8 mm thick, has a dielectric constant ϵ_r of 1.6 and a loss tangent $\tan \delta$ of 0.02. It was shown in [34] that the thickness of the fabric substrate is important to consider when designing wearable patch antennas. A fabric which is less than 1 mm thick results in poor radiation efficiency and gain. Therefore, a thick sample was chosen.

The value for the width of the aperture W is approximated using

$$W = \frac{c}{2f_r} \sqrt{\frac{2}{\epsilon_r + 1}} \quad (4.71)$$

since this is shown to produce good radiation efficiencies [35], where f_r is the resonant frequency and c is the speed of light.

The separation length L between each aperture is approximated using [28]

$$L = \frac{1}{2f_r \sqrt{\epsilon_{reff}} \sqrt{\mu_0 \epsilon_0}} - 2\Delta L \quad (4.72)$$

Where ϵ_{reff} is the effective dielectric constant to account for the fringing fields that propagate in the substrate and air from the patch antenna [28] and is given by

$$\epsilon_{reff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left[1 + 12 \frac{h}{W} \right]^{-\frac{1}{2}} \quad (4.73)$$

ΔL from (4.72) corresponds to the extended incremental length of the patch. The fringing fields produced by each aperture of the microstrip patch make the antenna look electrically larger than its length L . The extended incremental length ΔL is approximated by [36]

$$\Delta L = 0.412 h \frac{(\epsilon_{reff} + 0.3) \left(\frac{W}{h} + 0.264\right)}{(\epsilon_{reff} - 0.258) \left(\frac{W}{h} + 0.8\right)} \quad (4.74)$$

Since the antenna's intended operational frequency is at 2.4 GHz, the width and length can be approximated as $W_1 = 54.8$ mm and $L = 47.2$ mm. By computing (2.9) and varying the width of the smaller aperture $W_2 = 54.8$ mm to 5 mm, the normalized radiation intensity is plotted for both the E-plane and H-plane in dB. From Figure 4.5 the E-plane is the cut along the xz plane while the H-plane is the cut along the xy plane.

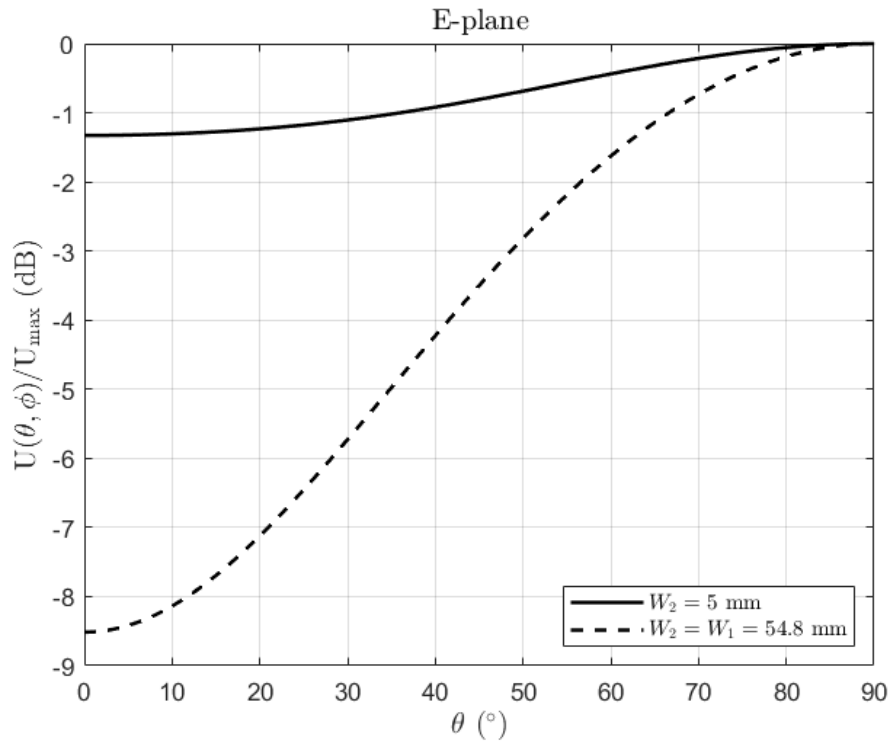


Figure 4.12: E-plane normalized linear far-field plot in terms of the radiation intensity U of a RPA where $W = W_1$ and a TPA where $W_2 = 5$ mm.

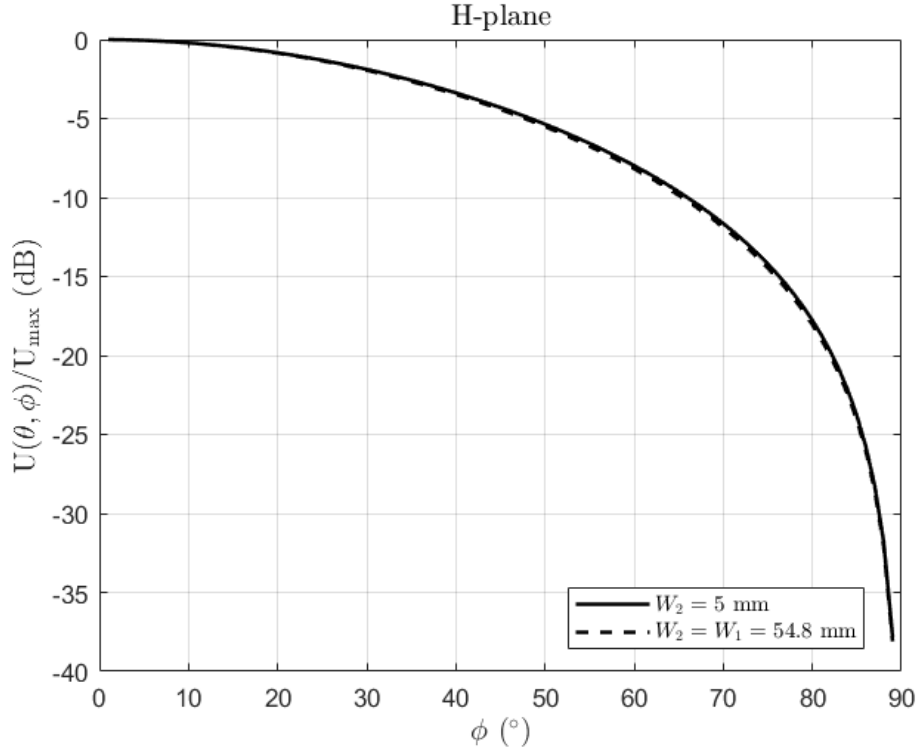


Figure 4.13: H-plane normalized linear far-field plot in terms of the radiation intensity U of a RPA where $W = W_1$ and a TPA where $W_2 = 5$ mm.

Comparing the two normalized radiation intensity plots some predictions can be made on miniaturizing the RPA to a TPA. For a TPA where $W_2 = 5$ mm, the antenna's farfield pattern in the E-plane becomes less directive whereby the beamwidth can be seen to increase since the radiation intensity has increased for the angles outside of where the maximum occurs at $\theta = 90^\circ$. The shape of the H-plane radiation intensity pattern in Figure 4.13 remains unaltered by reducing the width of Aperture #2.

By computing (2.10) the antenna's maximum directivity can be plotted to investigate how it changes by gradually reducing the width W_2 of Aperture #2.

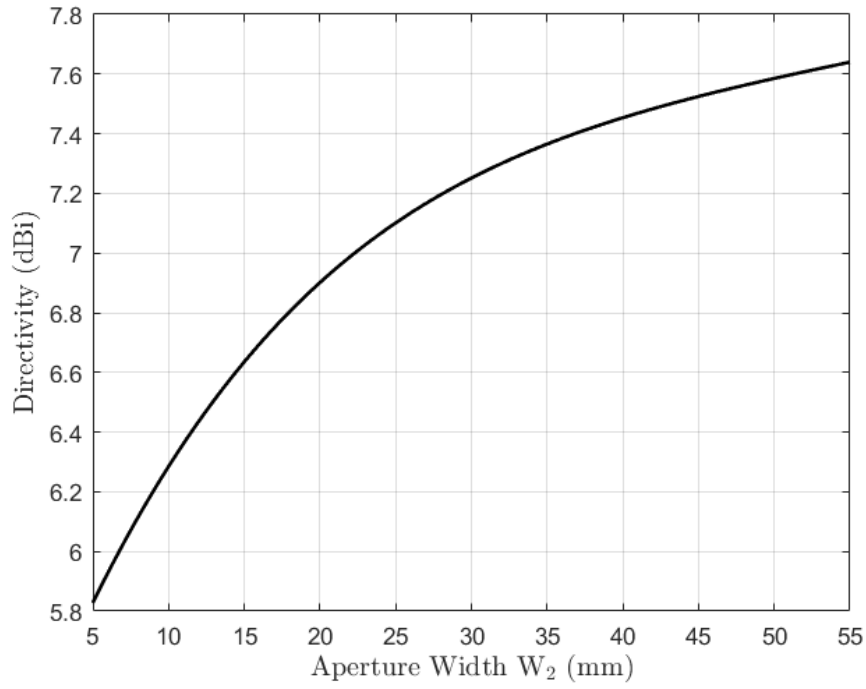


Figure 4.14: Parametric analysis modelling the patch antenna as a dual-aperture array by investigating the change in maximum directivity at $\theta = 90^\circ$ by varying the width W_2 of aperture #2.

For a RPA where $W = W_1 = W_2 = 54.8$ mm the directivity is at a maximum of 7.64 dBi. By reducing the aperture width W_2 and keeping the separation length L between the apertures constant, the directivity reduces to 5.83 dBi at $W_2 = 5$ mm. Increasing the separation length L can compensate for the reduction in directivity but only marginally.

From the Dual-AAM, it can be shown that the cost of miniaturizing the width of a single aperture of the patch antenna results in a decrease in the antenna's radiated power and directivity. The analysis predicts a 1.81 dBi decrease in directivity from an aperture width of 54.8 mm to 5 mm. The analysis can be taken further to compute gain or realized gain provided the conductor, dielectric and mismatch losses are obtained through measurements.

The decrease in the directivity for small aperture widths W_2 occurs due to most of the radiation propagation occurring at only a single radiating edge. The model does not account for the radiation that increasingly propagates at the diagonal lengths of the trapezoid as the aperture width W_2 becomes smaller. To account for the field propagation that occurs along the diagonal lengths L_d of the trapezoidal patch, two additional apertures need to be implemented into the model.

For a RPA, the apertures along the lengths L_d are regarded as non-radiating apertures. This is because the fields along these apertures almost cancel each other out completely in the principal planes, thus the non-radiating apertures are typically neglected in patch antenna analysis. The E-field along the non-radiating apertures have a co-sinusoidal taper, where the maximum E-field occurs at the edges across the length as shown in Figure 4.16. For a TPA, it is predicted that the magnitude of the E-field across the non-radiating apertures increases due to introducing a horizontal component to the aperture since it is placed diagonally across the yz plane.

In this section, non-radiating apertures are added to the aperture array analysis to account for the radiation along the edges of the TPA. Also, a new characterization method is presented for the TPA that incorporates the partial radiation that propagates along the diagonal lengths L_d which has both z and y components.

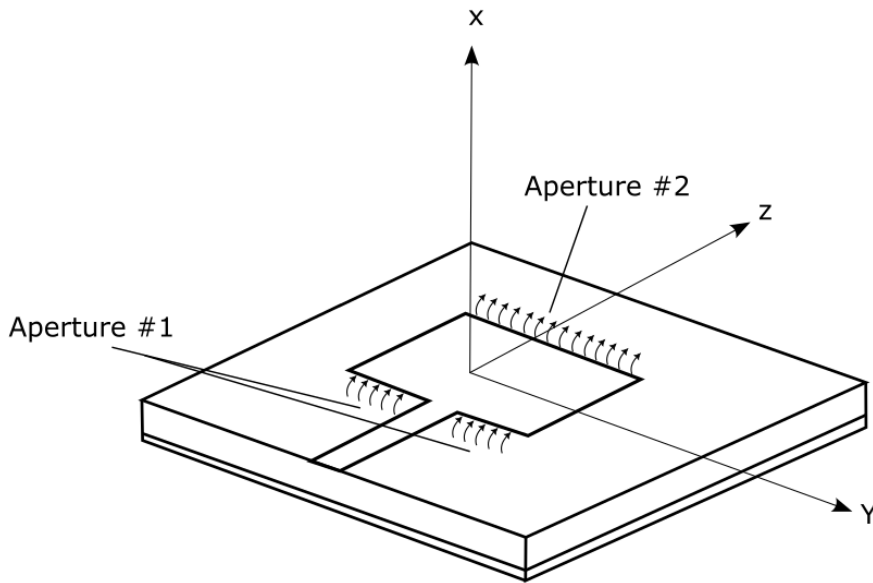


Figure 4.15: Rectangular patch antenna geometry showing the electric fringe fields at each aperture.

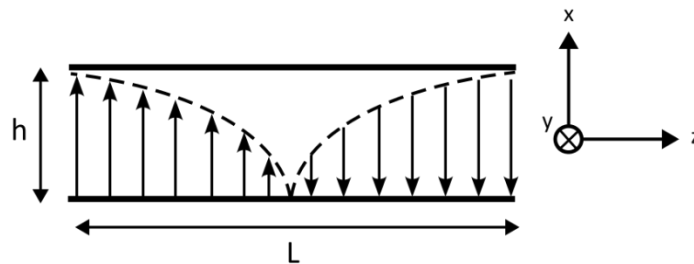


Figure 4.16: Non-radiating edges along the length of the rectangular patch antenna.

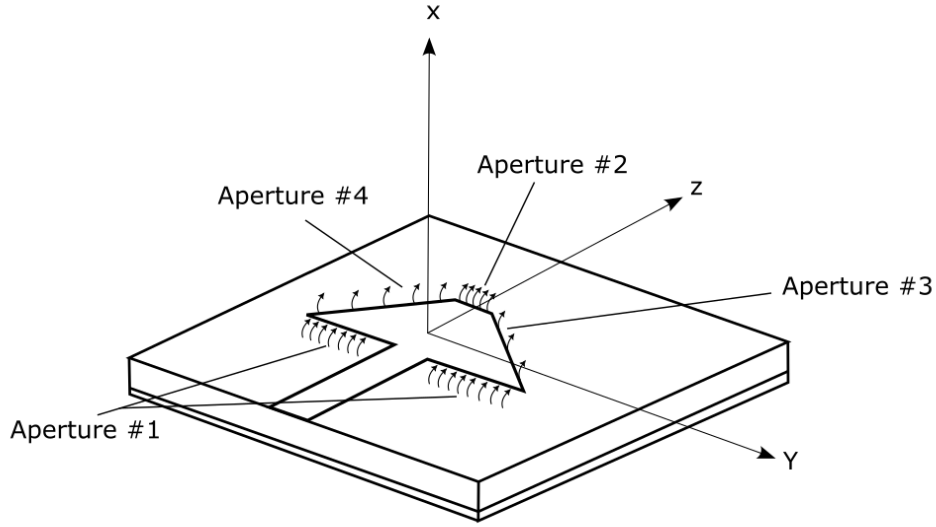


Figure 4.17: Trapezoidal patch antenna geometry showing the electric fringe fields at each aperture.

Figure 4.15 and Figure 4.17 show a comparison between the electric fringe fields that propagate for a RPA and a TPA. For a RPA, the electric fields predominantly propagate at Aperture #1 and #2.

The apertures formed along the lengths of the RPA are known as the non-radiating apertures. The E-field at these apertures take form as a co-sinusoidal taper

$$\mathbf{M}_s = -2\hat{a}_z \times E_0 \cos\left(\frac{\pi}{L}z\right) \quad (4.75)$$

$$\mathbf{J}_s = 0 \quad (4.76)$$

Since the magnetic current density propagates along the length or z axis, there is no M_z component to compute L_ϕ in (4.42), thus $E_\theta = 0$. This leaves only a E_ϕ component given as

$$E_\phi = \frac{-jke^{-jkr} \sin \theta}{4\pi r} \int_{-\frac{L}{2}}^{\frac{L}{2}} \int_{-\frac{h}{2}}^{\frac{h}{2}} 2E_0 \cos\left(\frac{\pi}{L}z'\right) e^{jk(x' \sin \theta \cos \phi + z' \cos \theta)} dx' dz' \quad (4.77)$$

$$E_\phi = \frac{-jke^{-jkr} E_0 \sin \theta}{2\pi r} \int_{-\frac{L}{2}}^{\frac{L}{2}} \int_{-\frac{h}{2}}^{\frac{h}{2}} \cos\left(\frac{\pi}{L}z'\right) e^{jk(x' \sin \theta \cos \phi + z' \cos \theta)} dx' dz' \quad (4.78)$$

$$E_\phi = \frac{-jke^{-jkr} E_0 \sin \theta}{2\pi r} \int_{-\frac{h}{2}}^{\frac{h}{2}} e^{jk(x' \sin \theta \cos \phi)} dx' \int_{-\frac{L}{2}}^{\frac{L}{2}} \cos\left(\frac{\pi}{L}z'\right) e^{jk(z' \cos \theta)} dz' \quad (4.79)$$

$$E_{\phi} = \frac{-jkh e^{-jkr} E_0 \sin \theta}{2\pi r} \left(\text{sinc} \left(\frac{1}{2} kh \sin \theta \cos \phi \right) \right) \left(\frac{2\pi L^2 \cos \left(\frac{1}{2} kL \cos \theta \right)}{\pi^2 - k^2 L^2 \cos^2 \theta} \right) \quad (4.80)$$

$$E_{\phi} = -jkhL^2 E_0 \sin \theta \left(\text{sinc} \left(\frac{1}{2} kh \sin \theta \cos \phi \right) \right) \left(\frac{\cos \left(\frac{1}{2} kL \cos \theta \right)}{\pi^2 - k^2 L^2 \cos^2 \theta} \right) \frac{e^{-jkr}}{r} \quad (4.81)$$

The non-radiating elements are placed along the y-axis as shown in Figure 4.18.

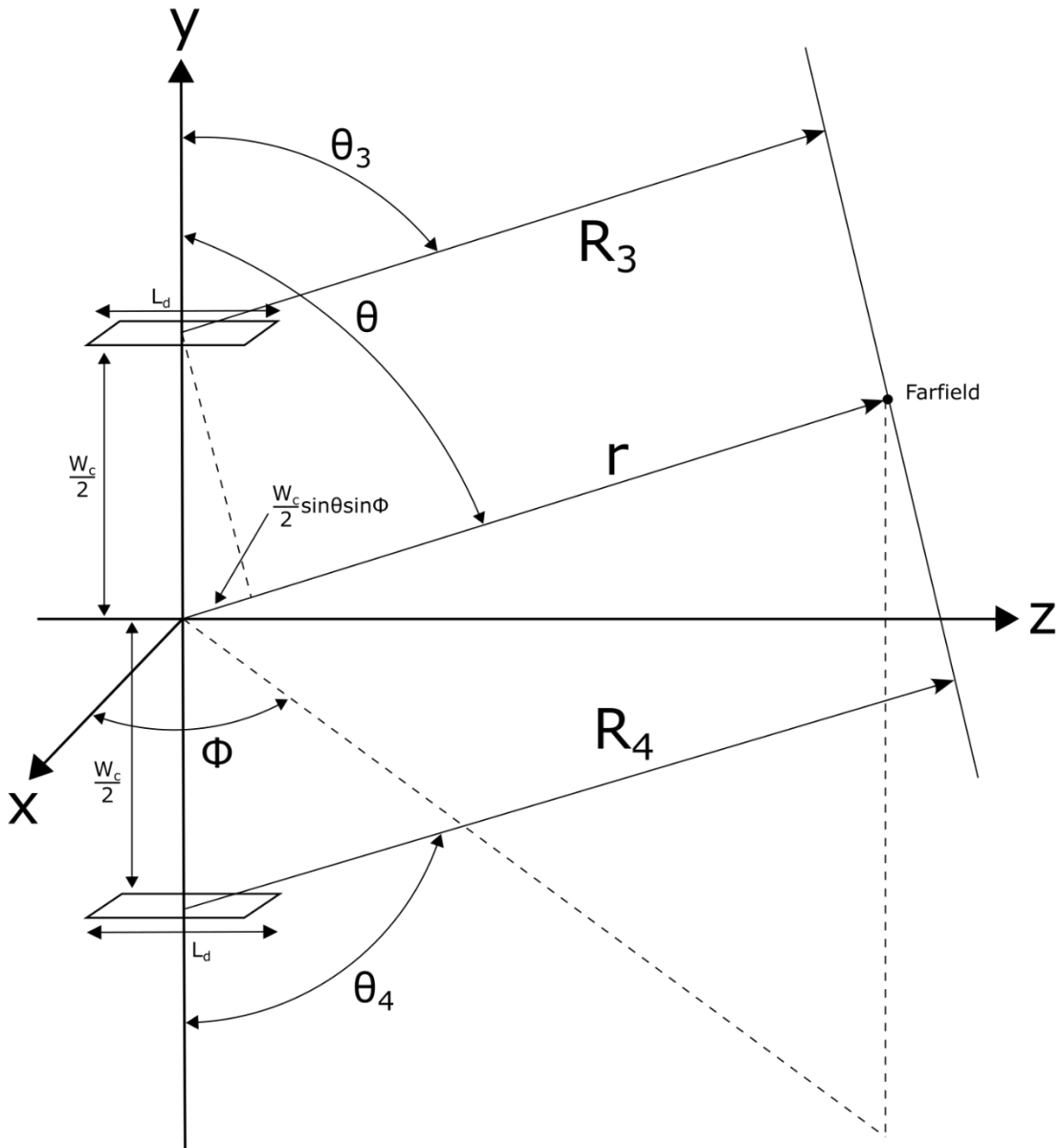


Figure 4.18: Aperture Array Model analysis and far-field approximation for Aperture #3 and #4 with lengths of L_d .

The array factor term AF which combines the radiation from both apertures together is transformed from the z axis to become

$$AF_y = \frac{e^{-jk\left(r - \frac{W_c}{2} \cos(\cos^{-1}(\sin \theta \sin \phi))\right)}}{r} + \frac{e^{-jk\left(r + \frac{W_c}{2} \cos(\cos^{-1}(\sin \theta \sin \phi))\right)}}{r} \quad (4.82)$$

which simplifies to

$$AF_y = \frac{e^{-jkr}}{r} \left(2 \cos \left(\frac{1}{2} k W_c \sin \theta \sin \phi \right) \right) \quad (4.83)$$

where $\cos^{-1}(\sin \theta \sin \phi)$ is the angle between the y axis of the array and the radial vector from the origin to the observation point. Therefore, the total field across both non-radiating apertures can be expressed as

$$E_{\Sigma\phi}^L = -jkhL^2 E_0 \sin \theta \left(\operatorname{sinc} \left(\frac{1}{2} kh \sin \theta \cos \phi \right) \right) \left(\frac{\cos \left(\frac{1}{2} kL \cos \theta \right)}{\pi^2 - k^2 L^2 \cos^2 \theta} \right) \left(2 \cos \left(\frac{1}{2} k W_c \cos(\cos^{-1}(\sin \theta \sin \phi)) \right) \right) \frac{e^{-jkr}}{r} \quad (4.84)$$

For the TPA the electric fields are predicted to propagate predominantly at Apertures #1 and #2 with partial electric field propagation at Apertures #3 and #4 due to the introduced horizontal components of the magnetic current density along the lengths of the patch. A relationship must be formed to link the aperture dimensions together to the field equations.

The front view geometry of the TPA is shown in Figure 4.19, where each length of the trapezoid corresponds to an equivalent radiating aperture. L_d is the diagonal length of the trapezoid, L_v is the vertical length or height of the trapezoid and W_{ext} is the width extension distance that varies with an angle of α as the width W_2 of the antenna changes. Using the Pythagoras theorem, a relationship can be formed between the diagonal length L_d of Aperture #3 and #4, vertical length L_v and the widths W_1 and W_2 of Aperture #1 and #2.

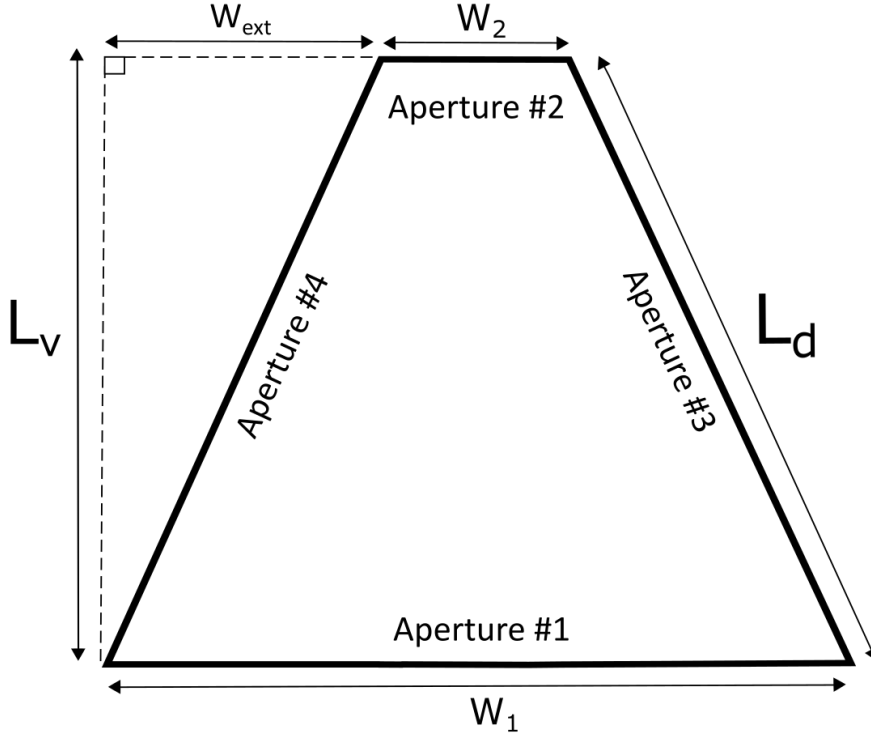


Figure 4.19: Front view trapezoidal patch geometry modelled as a quad-aperture array.

The diagonal lengths of the equilateral trapezoid can be calculated as

$$L_d = \sqrt{(L_v)^2 + (W_{\text{ext}})^2} \quad (4.85)$$

where

$$W_{\text{ext}} = \frac{1}{2}(W_1 - W_2) \quad (4.86)$$

The fields propagated from Apertures #1 and #2 remain the same as for the Dual-AAM. To form an expression for the electric fields propagated from Aperture #3 and #4, two additional apertures are required for each of Aperture #3 and #4. A vertical aperture along the length L_v and a horizontal aperture along the width extension W_{ext} . The vertical apertures have their electric field component in the form of (4.84), while the horizontal aperture has its electric field components in the form of (4.66) and (4.68). The additional apertures are represented in Figure 4.18 to approximate the fields at the far-field.

The combined electric field produced from Aperture #3 and #4 for both θ and ϕ components are given by the superposition of the two radiating slots at the far-field as

$$E_{\Sigma\theta}^{L_v} = 0 \quad (4.87)$$

$$E_{\Sigma\phi}^{L_v} = E_{\phi_3}^{L_v} + E_{\phi_4}^{L_v} \quad (4.88)$$

The radial distance R_1 and R_2 are defined as the distance from Aperture #3 and #4 of length L_v to the arbitrary far-field observation point. Both apertures are symmetrically positioned from the origin with a length of $\pm W_c/2$ along the z-axis, where W_c is the separation distance between the middle of apertures #3 and #4 defined as

$$W_c = \frac{1}{2}(W_1 + W_2) \quad (4.89)$$

The horizontal field components for Apertures #3 and #4 are modelled the same as for the Dual-AAM, resulting in electric field components given as

$$E_\theta = \frac{-jk}{4\pi} 2hW_c E_0 \cos \phi \operatorname{sinc}\left(\frac{1}{2}kh \sin \theta \cos \phi\right) \operatorname{sinc}\left(\frac{1}{2}kW_c \sin \theta \cos \phi\right) \frac{e^{-jkr}}{r} \quad (4.90)$$

$$E_\phi = \frac{jk}{4\pi} 2hW_c E_0 \cos \theta \sin \phi \operatorname{sinc}\left(\frac{1}{2}kh \sin \theta \cos \phi\right) \operatorname{sinc}\left(\frac{1}{2}kW_c \sin \theta \sin \phi\right) \frac{e^{-jkr}}{r} \quad (4.91)$$

The electric field components are symmetric about the z-axis and therefore the array factor term can be simplified to

$$AF_z = \frac{e^{-jkr}}{r} \left(2 \cos\left(\frac{1}{2}kw_c \cos \theta\right) \right) \quad (4.92)$$

Therefore, the total electric field components due to the horizontal components of Apertures #3 and #4 is given by

$$E_{\Sigma\theta}^{W_c} = \frac{-jk}{2\pi} 2hW_c E_0 \cos \phi \operatorname{sinc}\left(\frac{1}{2}kh \sin \theta \cos \phi\right) \operatorname{sinc}\left(\frac{1}{2}kW_c \sin \theta \cos \phi\right) \cos\left(\frac{1}{2}kw_c \cos \theta\right) \frac{e^{-jkr}}{r} \quad (4.93)$$

$$E_{\Sigma\phi}^{W_c} = \frac{jk}{2\pi} 2hW_c E_0 \cos \theta \sin \phi \operatorname{sinc}\left(\frac{1}{2}kh \sin \theta \cos \phi\right) \operatorname{sinc}\left(\frac{1}{2}kW_c \sin \theta \sin \phi\right) \cos\left(\frac{1}{2}kw_c \cos \theta\right) \frac{e^{-jkr}}{r} \quad (4.94)$$

The total electric field components for the TPA can be expressed as the sum of all apertures using (4.69) (4.70) (4.84) (4.93) and (4.94)

$$E_{\Sigma\theta}(\theta, \phi) = E_{\Sigma\theta}^{W_{12}} + E_{\Sigma\theta}^{L_v} + E_{\Sigma\theta}^{W_c} \quad (4.95)$$

$$E_{\Sigma\phi}(\theta, \phi) = E_{\Sigma\phi}^{W_{12}} + E_{\Sigma\phi}^{L_v} + E_{\Sigma\phi}^{W_c} \quad (4.96)$$

Equations (4.95) and (4.96) form the total electric field for the proposed Quad-AAM. To observe the behaviour of the radiation pattern at the far-field, the normalized radiation intensity is plotted for both the E-plane and H-plane in Figure 4.20 and Figure 4.21 by computing (2.9) and varying the width of a single aperture from $5 \text{ mm} \leq W_2 \leq 54.8 \text{ mm}$.

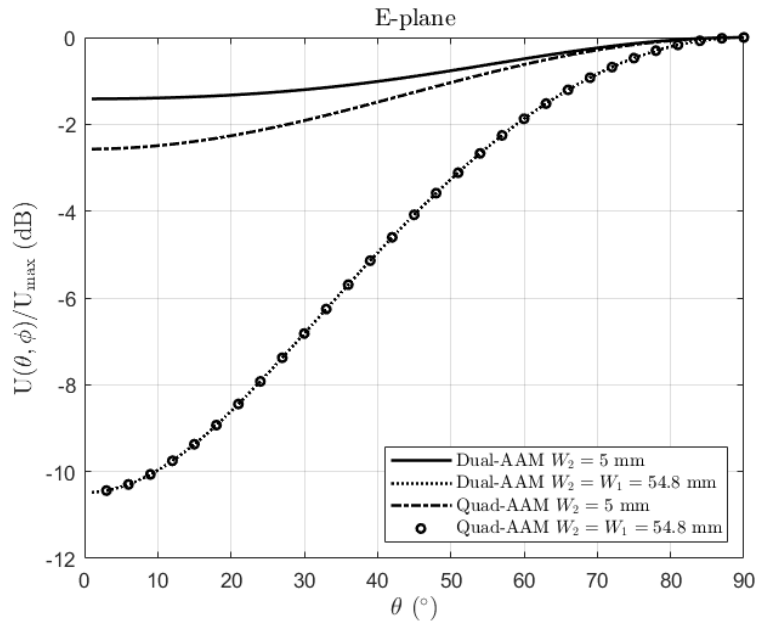


Figure 4.20: E-plane far-field plot in terms of the normalized radiation intensity U comparing the Dual and Quad-AAM for a RPA and TPA.

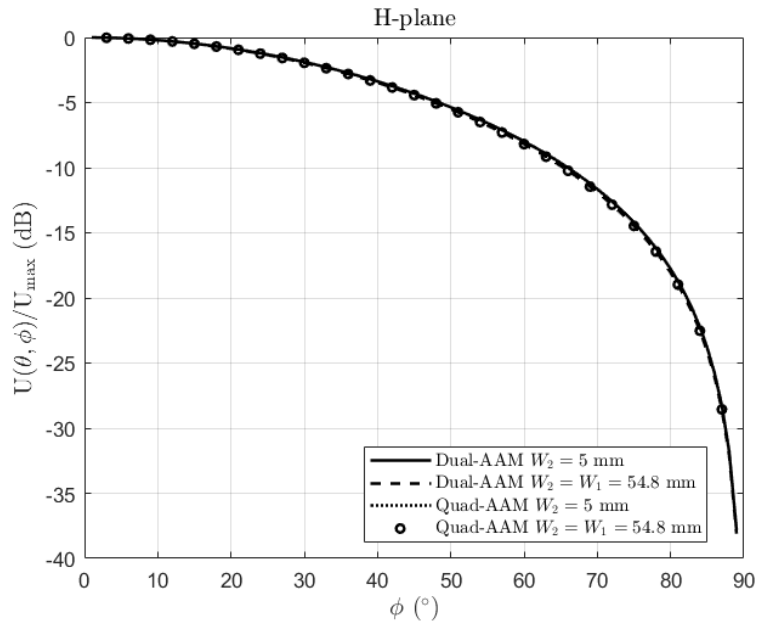


Figure 4.21: H-plane far-field plot in terms of the normalized radiation intensity U comparing the Dual and Quad-AAM for a RPA and TPA.

The two normalized radiation intensity plots show that for a TPA where $W_2 = 5$ mm that the antenna's farfield pattern in the E-plane becomes less directive than that of a RPA where $W_2 = 54.8$ mm. This is consistent with the Dual-AAM, however, when compared to the Quad-AAM the magnitude of the normalized radiation pattern outside of where the maximum occurs is predicted to be less than that of the Dual-AAM. This implies that the directivity will not reduce as much as what the Dual-AAM predicts. Just like for the Dual-AAM, the Quad-AAM predicts that the shape of the H-plane far-field pattern remains unaltered by reducing the aperture width W_2 .

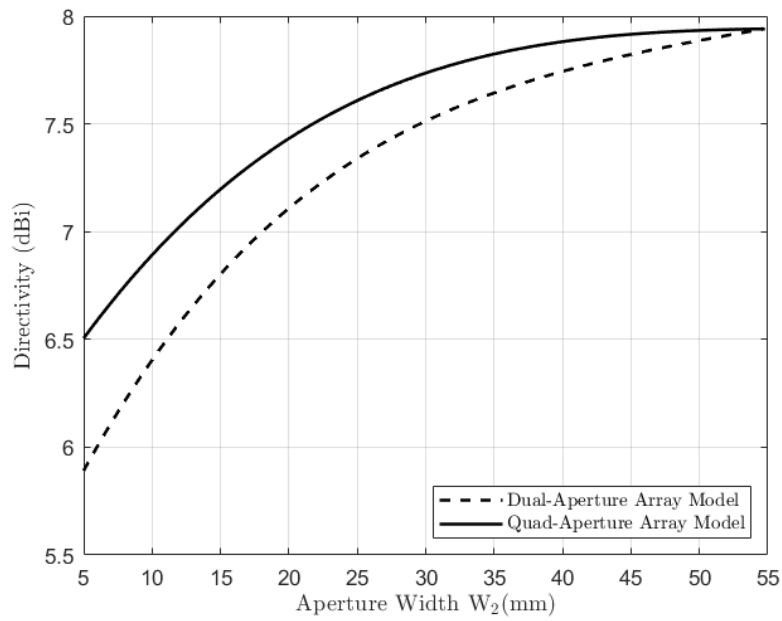


Figure 4.22: Comparison between the Dual and Quad-AAM analysis calculating the directivity of the patch antenna varying the aperture width W_2 .

By computing (2.10) the antenna's maximum directivity can be plotted to investigate how it changes by gradually reducing the width W_2 of aperture #2. This result is plotted in Figure 4.22 which compares the result to the Dual-AAM. It can be shown in the Quad-AAM that for smaller aperture widths of W_2 the directivity does not decrease in as much compared to the Dual-AAM. Both aperture array models show a decrease in directivity as the aperture width W_2 is reduced. However, the Quad-AAM maintains a higher directivity than the Dual-AAM even when the aperture width W_2 is reduced. For both models, the antenna's directivity decreases rapidly when the aperture width W_2 is reduced below 25 mm. At $W_2 = 5$ mm, the difference in directivity between the models is approximately 0.6 dBi.

To summarise, the aim of this section was to develop a simple analytical model to characterize the far-field characteristics of a TPA. The parameters of interest for this analysis were the antenna's far-field pattern and directivity. The analysis was implemented in MATLAB as shown in Appendix A. The Quad-AAM is proposed as a suitable analytical method for characterizing the TPA in the far-field. In the next section, the Quad-AAM is validated against a full-wave simulation analysis.

4.4 Full Wave Simulation Analysis

To compare the results produced by the AAM, the TPA is modelled and simulated in CST Microwave Studio. The time and frequency domain solvers are used to compare simulations to verify the results. Before designing the TPA, an inset-fed rectangular patch antenna was developed to act as a performance benchmark. The inset-feed offers a simple method for impedance matching compared to an impedance matching network involving fixed or tunable inductors and capacitors. It requires etching out a small portion of the radiating element where the transmission line intersects the radiating element. The depth and width of the inset control the input impedance.

The width W and vertical length L_v of the RPA are computed using (4.67) and (4.68). Using the analytical transmission line model as presented by Balanis in [28], the width of the feedline W_0 and the inset feed recessed depth y_0 can be approximated.

The width of the feed line W_0 is calculated using [31]

$$Z_c = \frac{120\pi}{\sqrt{\epsilon_{reff}} \left[\frac{W_0}{h} + 1.393 + 0.667 \ln \left(\frac{W_0}{h} + 1.444 \right) \right]} \quad (4.97)$$

Which is valid for $W_0/h > 1$. Z_c is the characteristic impedance of the microstrip feedline which is set to 50 Ω .

Thus, an approximate inset feed recessed depth y_0 can be calculated as [28]

$$R_{in} = \frac{1}{2(G_1 \pm G_{12})} \cos^2 \left(\frac{\pi}{L} y_0 \right) \quad (4.98)$$

Where the calculation for the conductance of a slot G_1 and the mutual conductance between two slots G_{12} can be found in [28].

For a substrate height of 2.8 mm, the calculated parameters using the Transmission Line Model are shown in Table 4.2.

Parameter	Symbol	Transmission Line Model Dimensions (mm)
Vertical Patch Length	L_v	49.4
Aperture #1 Width	W_1	54.1
Aperture #2 Width	W_2	54.1
Inset Recessed Distance	y_0	17.8
Feedline Width	W_0	10.5
Substrate Height	H_s	2.8
Inset Feed Gap Width	G_{tp}	-

Table 4.2: Dimensions obtained using the Transmission Line Analysis for a RPA for operation at 2.4 GHz.

The RPA is designed and simulated in CST Microwave Studio using the time domain and frequency domain solver. Table 4.2 shows the analytical dimensions for the RPA. Comparing this to the simulated results in Table 4.3 and Table 4.4 for when $W_1 = W_2$, there is shown to be a decrease within a few millimetres for the length L_v and width W_1 to the analytical dimensions obtained using the transmission line model. However, there is a large difference of 4.3 mm between the calculation of the inset recessed distance and the simulated results. This can be explained due to the transmission line not considering the width of the inset feed G_{tp} , since through the simulations it is shown that it has a significant effect on the antennas input impedance. Adjusting both the inset feed gap width G_{tp} and the inset recessed distance y_0 the antennas input impedance can be fine-tuned to match 50Ω .

A ground plane width and length of 110 mm is chosen since a ground plane size of approximately $\lambda/2$ is considered to be the critical ground plane size for input impedance matching [37]. Simulating the antenna with these dimensions results in suitable shielding resulting in a minimal radiating back lobe in the radiation pattern. The width W_2 of aperture #2 is varied from $W_2 = W_1 = 52$ mm to $W_2 = 5$ mm while W_1 is kept constant to see the effects of transforming the RPA to a TPA has on its radiation characteristics. The values given in Table 4.3 and Table 4.4 are the simulated results for varying the aperture width W_2 using the frequency and time domain solver.

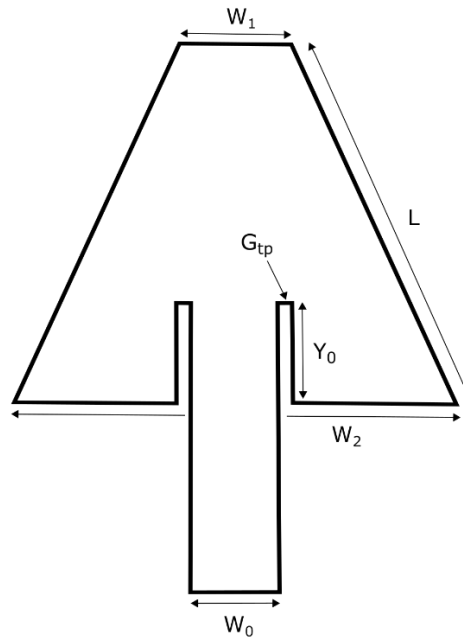


Figure 4.23: Front view geometry of the Trapezoidal Patch Antenna.

The width W_2 of aperture #2 is varied from $W_2 = W_1 = 52$ mm to $W_2 = 5$ mm while W_1 is kept constant to see the effects of transforming the RPA to a TPA has on its radiation characteristics. The values given in Table 4.3 and Table 4.4 are the simulated results for varying the aperture width W_2 using the frequency and time domain solver.

W_2 (mm)	f_{res} (GHz)	L_v (mm)	L_d (mm)	G_{tp} (mm)	y_0 (mm)	BW (%)	D (dBi)	G_{re} (dBi)	e_0 (%)	RESA (mm ²)	Area Size Reduction %
5	2.4	53.34	58.29	2.67	11.8	2.36	7.74	4.47	47.1	1748.6	38.9
10	2.4	51.54	55.66	2.53	11.8	2.35	7.78	4.62	48.3	1839.2	35.7
15	2.4	50.80	54.06	2.53	11.4	2.48	7.77	4.66	48.9	1948.8	31.9
20	2.4	50.30	52.78	2.53	11.4	2.51	7.80	4.82	50.3	2060.3	28.0
25	2.4	50.02	51.81	2.53	11.4	2.53	7.85	4.98	51.7	2176.8	23.9
30	2.4	50.02	51.22	2.53	11.4	2.58	7.89	5.11	52.6	2301.8	19.5
35	2.4	50.00	50.72	2.53	11.4	2.68	7.96	5.28	53.8	2426.1	15.2
40	2.4	50.13	50.49	2.53	11.4	2.73	7.99	5.39	54.9	2556.4	10.6
45	2.4	50.24	50.37	2.53	11.4	2.76	8.07	5.56	56.1	2686.6	6.09
52	2.4	50.23	-	2.50	11.6	2.80	8.14	5.73	57.7	2860.9	0

Note: RESA = Radiating Element Surface Area

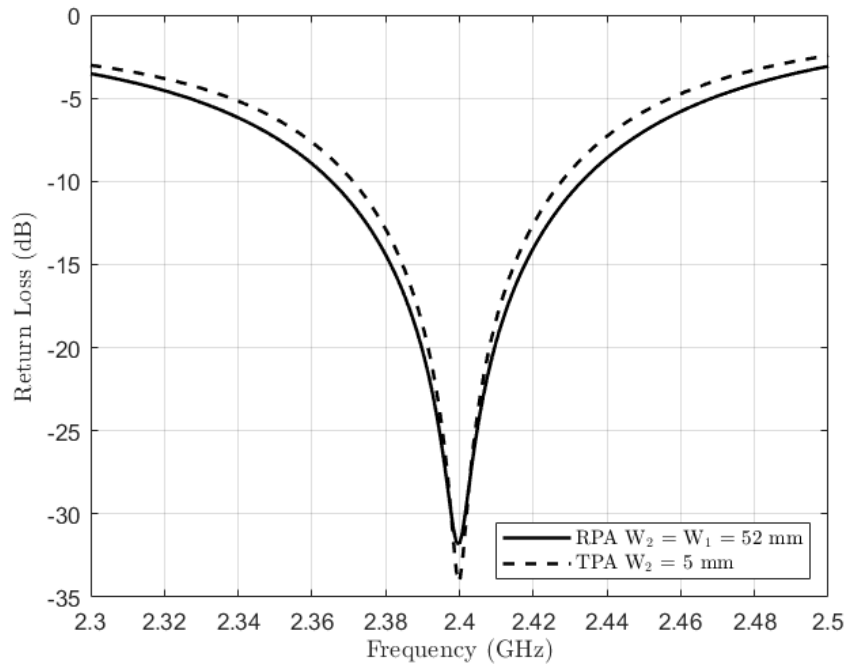
Table 4.3: A comparison of the simulated dimensions and radiation performance parameters of the patch antenna by varying the aperture width W_2 using the frequency domain solver with a tetrahedral mesh.

W_2 (mm)	f_{res} (GHz)	L_v (mm)	L_d (mm)	G_{tp} (mm)	y_0 (mm)	BW (%)	D (dBi)	Gain (dBi)	e_0 (%)	RESA (mm ²)	Area Size Reduction %
5	2.4	51.95	57.02	2.23	13.0	2.59	8.03	4.94	49.1	1721.3	38.5
10	2.4	50.36	54.56	2.15	13.0	2.63	8.03	5.01	49.9	1812.6	35.2
15	2.4	49.46	52.80	2.12	13.0	2.69	8.04	5.12	50.1	1913.4	31.6
20	2.4	49.06	51.60	1.95	13.0	2.82	8.05	5.22	52.1	2029.2	27.5
25	2.4	49.00	50.84	1.94	13.0	2.78	8.05	5.31	53.1	2150.5	23.2
30	2.4	48.9	50.16	1.93	13.0	2.88	8.10	5.44	54.3	2270.9	18.9
35	2.4	48.7	49.43	2.53	13.0	2.96	8.22	5.69	55.8	2367.8	15.4
40	2.4	48.8	49.21	2.53	13.0	3.07	8.26	5.80	56.8	2495.9	10.8
45	2.4	49.0	49.12	2.53	13.0	3.19	8.30	5.92	57.8	2624.7	6.21
52	2.4	49.0	-	2.53	13.0	3.30	8.38	6.08	58.9	2798.4	0

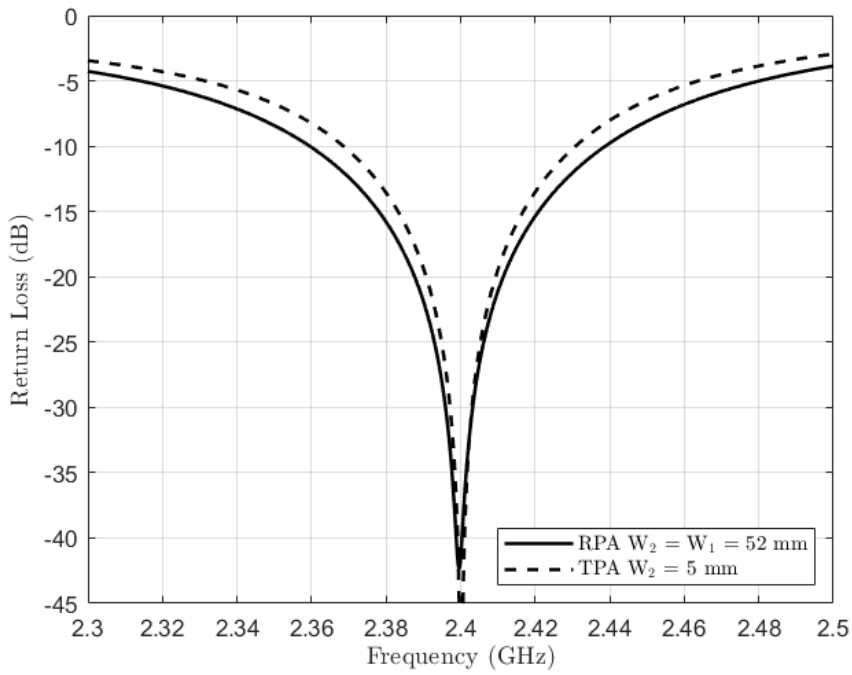
Note: RESA = Radiating Element Surface Area

Table 4.4: A comparison of the simulated dimensions and radiation performance parameters of the patch antenna by varying the aperture width W_2 using the time domain solver with a hexahedral mesh.

Marginal resonant frequency detuning is shown to occur when the width W_2 would decrease while all the other parameters remained the same. Small adjustments were made to the vertical length L_v , inset feed gap G_{tp} and inset feed recessed depth y_0 to maintain operation at 2.4 GHz with minimal mismatch. As shown in Table 4.3 and Table 4.4, these dimension changes were small.

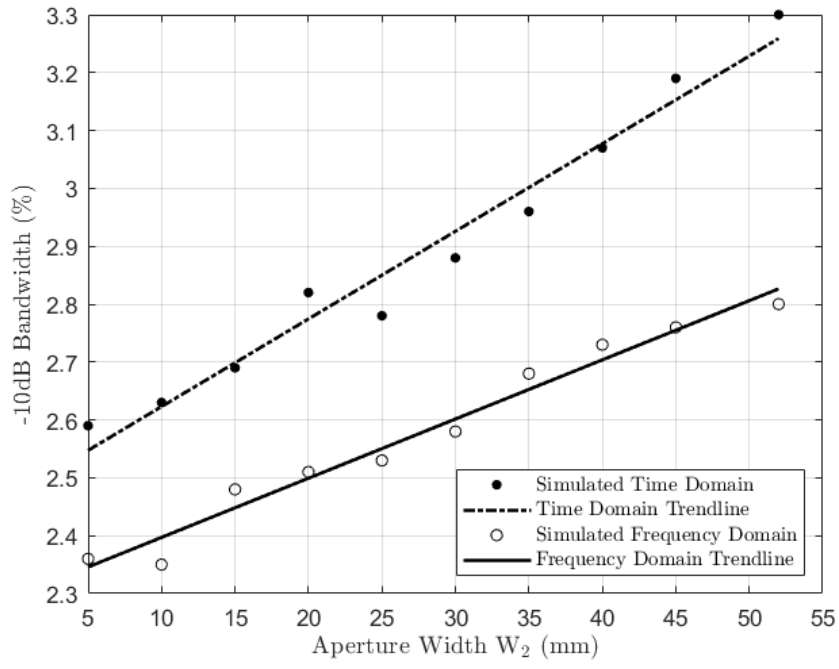


(a)

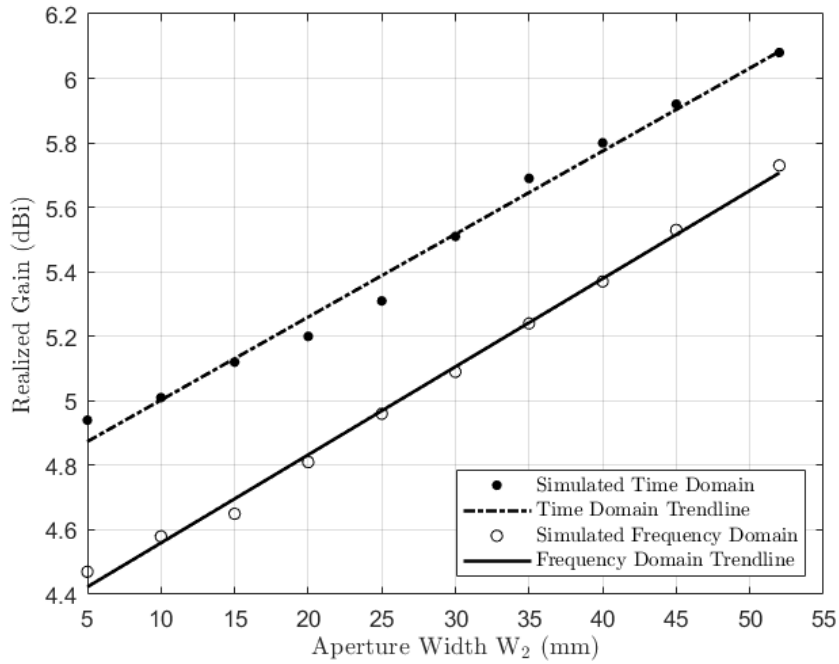


(b)

Figure 4.24: Comparison of simulated reflection coefficients for the TPA and RPA using the (a) frequency domain solver and (b) time domain solver.



(a)



(b)

Figure 4.25: Comparison between the time domain and frequency domain solver in CST for (a) bandwidth and (b) realized gain while varying the aperture width W_2 .

The simulated results using the frequency domain solver show that a linear relationship can be established between the decrease in aperture width W_2 and the decrease in realized gain and bandwidth. In Figure 4.25 the realized gain reduces by a total of 1.26 dBi from an aperture width W_2 of 52 mm to 5 mm. The difference in the -10 dB bandwidths can be seen

in Figure 4.24 in which the TPA results in a 10.7 MHz reduction. Figure 4.26 shows the variation in aperture width W_2 and the reduction in the surface area of the radiating element. A surface area size reduction of 38.9 % is achieved reducing the aperture width W_2 to 5 mm.

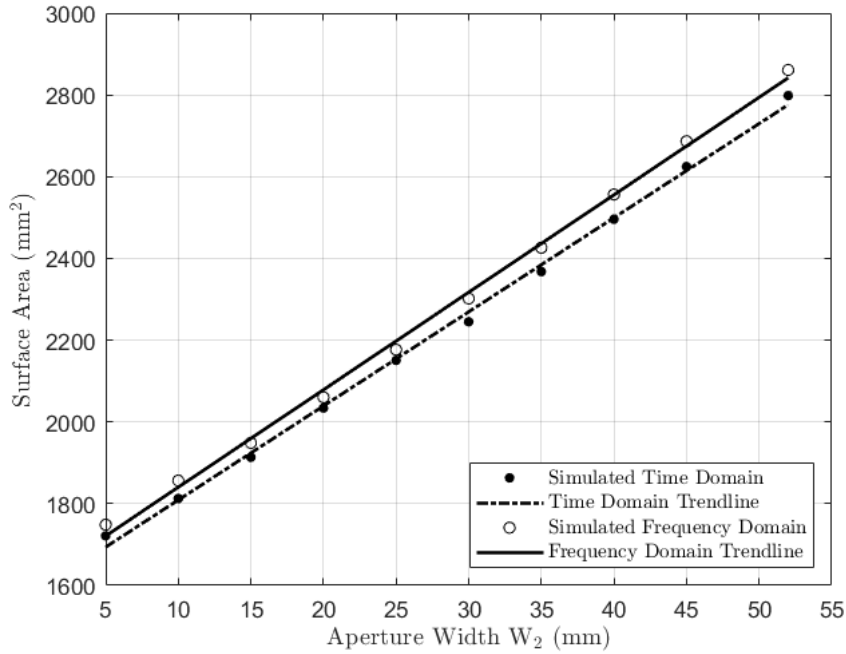


Figure 4.26: Change in the antenna's radiating element surface area while varying the aperture width W_2 and maintaining a resonant frequency of 2.4 GHz.

The simulations are repeated using the time-domain solver with a hexahedral mesh. Table 4.4 summarizes the results. There is a slight dimensional difference for the inset feed recessed distance y_0 , which required an increase to 13 mm to match the antennas input impedance back to 50 Ω . There are also some notable differences with the simulated bandwidth and gain. The time-domain solver predicts a 12 MHz increase for $W_2 = 52$ mm and a 5.6 MHz increase for $W_2 = 5$ mm when compared to the frequency domain solver. The difference in realized gain is marginal with the time-domain solver predicting a 0.35 dBi increase for $W_2 = 52$ mm and a 0.47 dBi increase for $W_2 = 5$ mm when compared to the frequency domain solver. The surface area size reduction agrees well with the frequency domain results resulting in a 38.5% size reduction when comparing a trapezoidal radiating element to a rectangular one.

The simulated far-field plots are shown in Figure 4.27 and Figure 4.28 which compares the results obtained using the time and frequency domain solvers. The analytical result using the Quad-AAM is also compared alongside. The far-field cuts are plotted in terms of

normalized electric field. For the simulated E-plane cuts, the maximum magnitude is translated $+90^\circ$ to match the analytical AAM far-field coordinate system for a direct comparison.

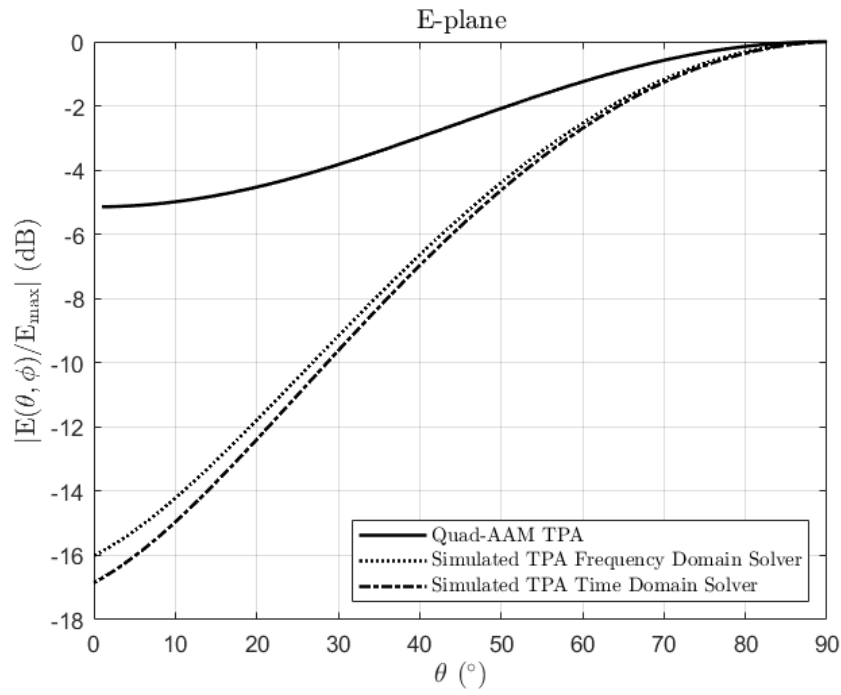
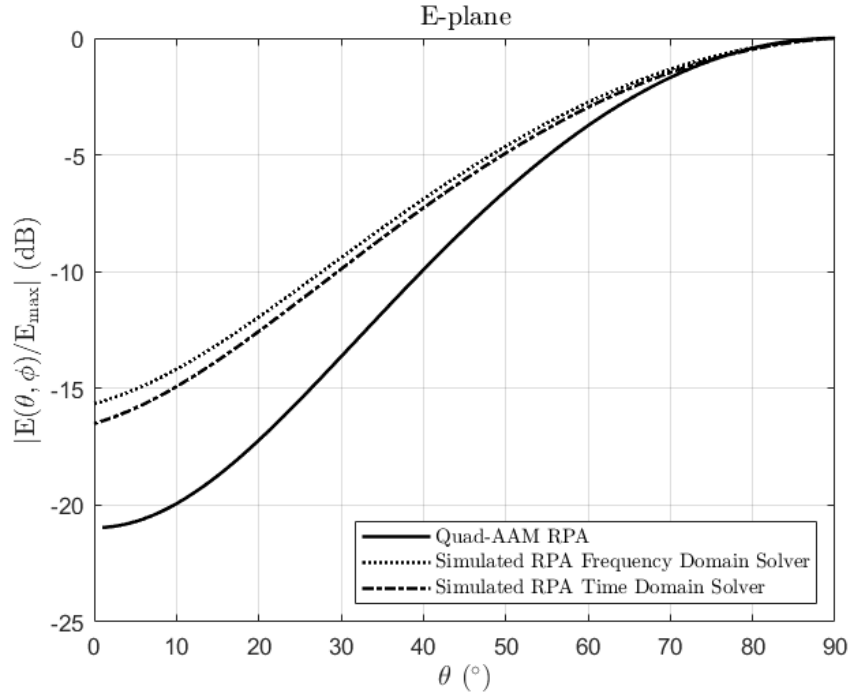
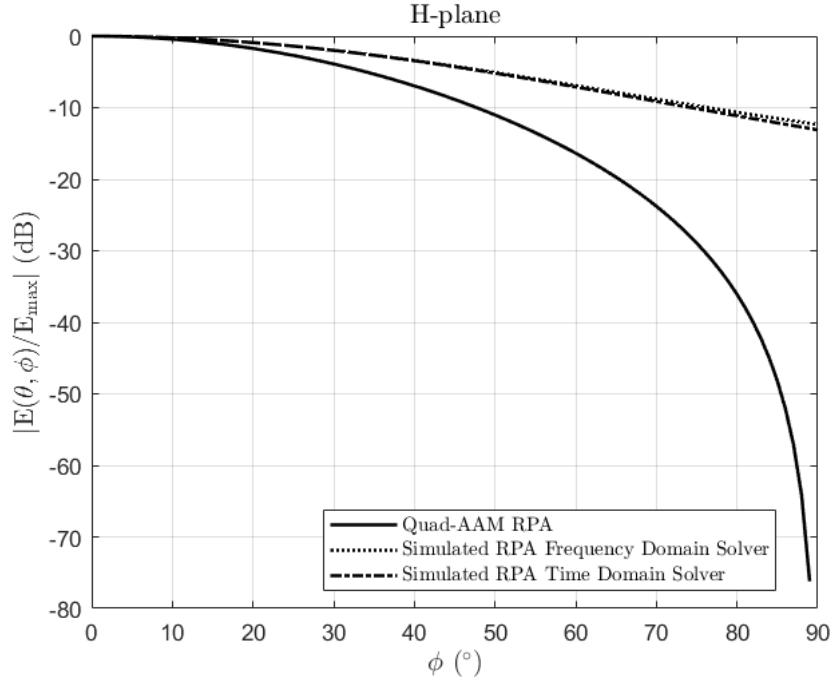
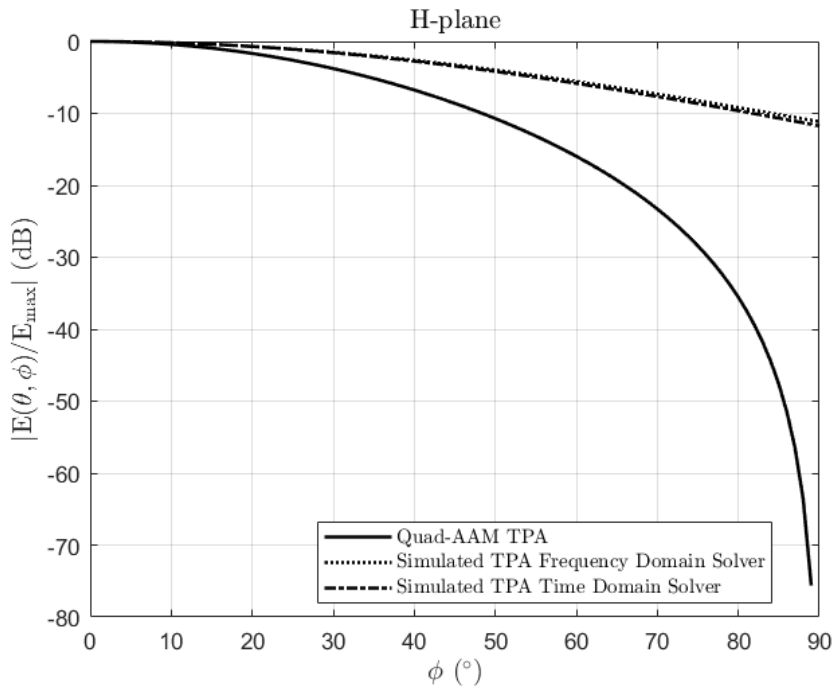


Figure 4.27: E-plane radiation pattern cut comparison for the (a) RPA and (b) TPA between the analytical Quad-AAM and full-wave simulations using the time and frequency domain solver.



(a)



(b)

Figure 4.28: H-plane radiation pattern cut comparison for the (a) RPA and (b) TPA between the analytical Quad-AAM and full-wave simulations using the time and frequency domain solver.

For both the E and H-plane, the simulations using both solvers agree well with each other. There is however a difference between the Quad-AAM analysis and the simulation results. For both the E and H-plane, the RPA and TPA have radiation patterns which are almost

identical. The Quad-AAM predicts that the magnitude of the E-plane decreases, which means that radiated power is distributed more broadly which implies that directivity decreases. However, the simulated far-field pattern does not change from the RPA to the TPA, unlike the Quad-AAM, which exceeds expectations in this regard. The reason for this is due to the adjustment of the vertical length of the antenna. Converting the RPA to the TPA a vertical length increase adjustment is made to counter the frequency detune. Whilst this a marginal overall increase by up to 3 mm in L_v from reducing W_2 to 5 mm, the effective length of L_v will be larger and thus the diagonal apertures contribute more to the overall radiated power and thus directivity for a smaller aperture width W_2 . Currently the AAM does not account for resonant frequency detuning and assumes that the structure resonates at 2.4 GHz for all aperture widths.

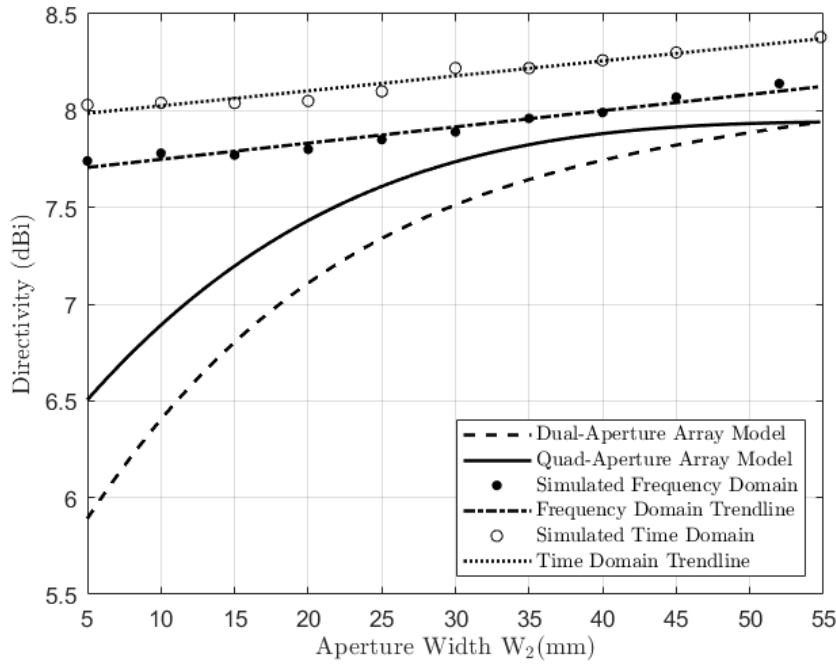


Figure 4.29: Comparison between the proposed analytical models and the simulated analysis calculating the directivity of the antenna varying the aperture width W_2 .

The simulated maximum directivity of the antenna for the time and frequency domain solver is plotted in Figure 4.29. Both simulated values for directivity are consistent with each other showing a small linear decrease. Using the time domain solver to vary the antennas aperture width W_2 to 5 mm, the total reduction in the antenna's directivity is 0.4 dBi, while using the frequency domain the directivity decreases a total of 0.35 dBi.

Comparing the simulated directivity results with the analytical analysis it can be shown that the Quad-AAM is consistent with the simulated frequency domain trendline up till approximately $W_2 = 30$ mm where the directivity starts to diverge for decreasing the aperture width W_2 . This is similar for the Dual-AAM but it starts to diverge at approximately $W_2 = 40$ mm. The Quad-AAM predicted while reducing W_2 , the radiation produced by the diagonal radiating apertures would compensate for the size reduction. The simulation analysis shows that the maximum directivity of the antenna at smaller aperture widths supersedes the performance to the analytical models. The explanation for this divergence is the same explanation mentioned earlier due to the countering of the frequency detuning by increasing the vertical length of the antenna L_v , which limits the decrease in directivity for smaller values of W_2 .

Despite the subtle differences in the simulated results the correlation trends in the performance parameters remain the same.

4.5 Conclusion

Overall, the simulated analysis in CST showed that the radiating element's surface area can be reduced by up to 39% when compared to a RPA with a marginal decrease in the antennas realized gain, directivity, bandwidth, and radiation efficiency. Reducing the size of the patch antennas radiating element is shown to be a key design consideration for low-profile wearable textile patch antennas used in on-body to off-body communication links. The decrease in the radiating elements size supersedes the marginal decrease in the antenna's radiation performance.

A partial textile prototype for the RPA and TPA was created and analysed. These are preliminary results as the measurements were taken using a NanoVNA and are reported in Appendix B.1. Further investigations are required by performing measurements using a VNA as well as far-field measurements in the anechoic chamber to validate the full-wave simulation results. However, the preliminary measured results in Appendix B.1 show an agreement with the degree of surface area miniaturization and bandwidth reduction when converting the RPA to a TPA.

4.6 References

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5 Investigating Patch Antenna Bending Effects

In this chapter cylindrical bending of the patch antenna is investigated. Following a literature review, an extension to the Aperture Array Model is presented for the Trapezoidal Patch Antenna as an aperture array on a conducting cylinder. A far-field analysis is performed to investigate the antenna's directivity for various bending angles.

Finally, a full wave simulation analysis using CST Microwave Studio is performed on the trapezoidal patch antenna to validate the analytical model and further investigate antenna bending in detail. The key parameters investigated in this analysis are shift in resonant frequency, directivity, radiation efficiency, realized gain and bandwidth.

5.1 Literature

Analytical models in literature for patch antenna bending are outlined and discussed in this section. Treating the patch antenna as a cylindrical cavity forms the basis of analysing patch antenna bending to analytically obtain key parameters such as the antenna's resonant frequency, radiation pattern, radiation efficiency, input impedance and Q-factor [1]–[5]. In the Cylindrical Cavity Model the patch antenna is modelled as a sector of an annulus between two concentric circular cylinders as shown in Figure 5.1 and Figure 5.2.

Some of the earliest work on analytical modelling of patch antenna bending is presented by C. Krowne [1] in 1981. C. Krowne used the cylindrical cavity model and a circuit equivalent model to produce an expression for the antenna's radiation efficiency. A more rigorous analytical analysis was presented by C. Krowne in 1984 on patch antenna bending [2] which focused on varying the patch antenna's dielectric substrate thickness while bending the antenna to investigate how the resonant behaviour changes. It was found that curvature changed the resonant frequency by less than about 5% for a substrate thickness equal to one-tenth of the radius curvature [2].

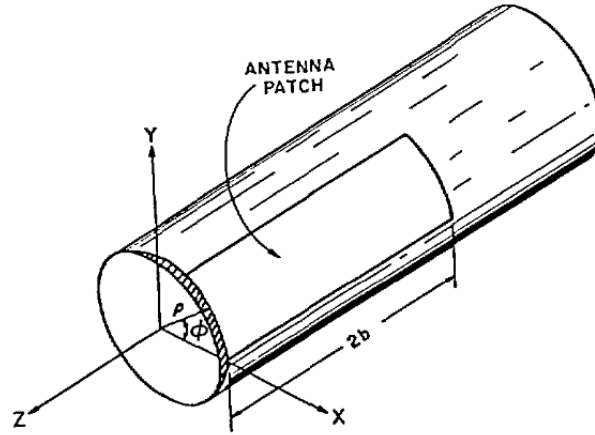


Figure 5.1: Perspective view of a rectangular patch antenna modelled as a cylindrical-rectangular cavity [2].

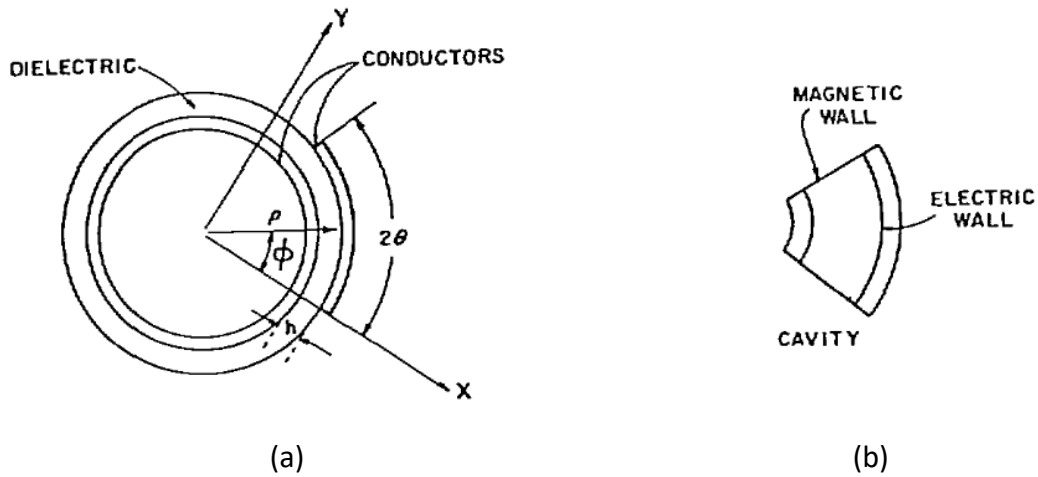


Figure 5.2: Cross-sectional view of (a) the cylindrical-rectangular cavity model and (b) the isolated sector that forms the patch antenna showing the associated boundary conditions [2].

The geometry of the patch antenna modelled in the cylindrical-rectangular cavity model is shown in Figure 5.1 and Figure 5.2, where 2θ is the bending angle along the ϕ component, $2b$ is the length of the radiating element and h is the height of the substrate. Figure 5.2(b) shows the boundary conditions that are applied to the patch. The conductive radiating element and ground plane are represented as electric walls, while the surrounding four walls of the substrate are represented as magnetic walls. The radius of the concentric cylinders along the ρ component is a and $a + h$ as shown more clearly in Figure 5.8(a) which is based on the same model.

By solving the electromagnetic vector wave equation (Helmholtz) in cylindrical coordinates with the appropriate boundary conditions, the resonant frequency of the patch antenna can be expressed for both TM^z and TE^z modes as [2]

$$(f_r)_{mli} = \frac{1}{2\sqrt{\mu\epsilon}} \sqrt{k_{mi}^2 + \left(\frac{l\pi}{2b}\right)^2} \quad (5.1)$$

Where k_{mi} is the radial wavenumber along the ρ component, which is solved for TE^z modes through

$$J'_v(k_{mi}a)Y'_v(k_{mi}(a+h)) - J'_v(k_{mi}(a+h))Y'_v(k_{mi}a) = 0 \quad (5.2)$$

and for TM^z modes using

$$J_v(k_{mi}a)Y_v(k_{mi}(a+h)) - J_v(k_{mi}(a+h))Y_v(k_{mi}a) = 0 \quad (5.3)$$

Where J_v and Y_v is the Bessel function of the first and second kind. The order of the Bessel function v is given by

$$v = \frac{m\pi}{2\theta}, \quad m = 1, 2, 3, \dots \quad (5.4)$$

The fundamental TE^z mode corresponds to when $m = 1$, $l = 0$ and $i = 1$, and for the fundamental TM^z mode when $m = 0$, $l = 1$ and $i = 1$.

In the late 1980s, Dahele et al. [3] took the analysis further by computing the patch antenna's radiation pattern and input impedance which was then compared to measured results.

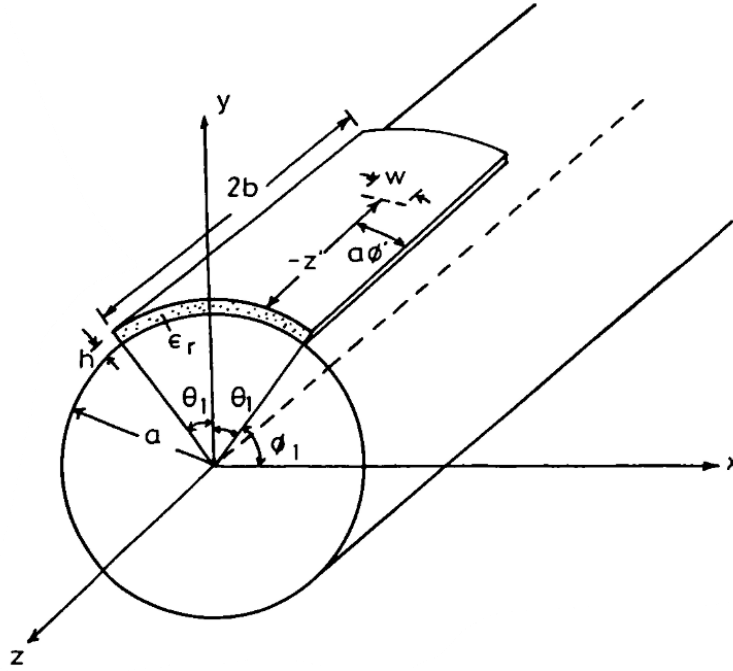


Figure 5.3: Geometry of the cylindrical-rectangular patch antenna [3].

The geometry of the Cylindrical Cavity Model in [3] is given in Figure 5.3 which is almost the same as that given in Figure 5.1. The difference in this model is that the antenna is fed by a coaxial cable, which is represented as a current ribbon of width w to calculate the antenna's input impedance. In this analysis the derived resonant frequencies for the TM_{mn} modes of the antenna are expressed as

$$(f_r)_{mn} = \frac{c}{2\sqrt{\epsilon_r}} \sqrt{\left(\frac{m}{2(a+h)\theta_1}\right)^2 + \left(\frac{n}{2b}\right)^2} \quad (5.5)$$

The resonant frequencies are simpler to compute using equation (5.5) when compared to equation (5.1). (5.1) requires the computation of the transcendental equations (5.2) and (5.3). Equation (5.5) was shown to be in agreement with the measured results in [3] for the TM_{01} mode showing an unchanged resonant frequency for varying the bending angle. The TM_{10} mode was not investigated through measurements for validation.

The analytical analysis for the radiation pattern in [3] is based on the analysis presented by Roger F. Harrington [6] in 1961 on apertures in conducting cylinders. At the magnetic walls of the patch antenna the currents are said to radiate radially in the presence of the cylindrical surface [3], therefore for the TM_{10} mode the electric field components are expressed as

$$E_\phi = \left(-\frac{2E_0h}{a\pi^2r}\right) \frac{\sin(k_0b \cos \theta)}{k_0 \cos \theta} e^{-jk_0(r+b \cos \theta)} \sum_{p=0}^{\infty} \frac{\epsilon_p j^p \cos(p\theta_1)}{H_p^{(2)'}(k_0a \cos \theta)} \cos\left[p\left(\phi - \frac{\pi}{2}\right)\right] \quad (5.6)$$

$$E_\theta \approx 0 \quad (5.7)$$

and for the TM_{01} mode

$$E_\theta = \left(j\frac{2E_0h\theta_1}{\pi^2r}\right) \frac{\cos\left(\frac{\pi}{2\sqrt{\epsilon_r}} \cos \theta\right)}{\sin \theta} e^{-jk_0(r+b \cos \theta)} \times \sum_{p=0}^{\infty} \frac{\epsilon_p j^p}{H_p^{(2)'}(k_0a \cos \theta)} \frac{\sin(p\theta_1)}{p\theta_1} \cos\left[p\left(\phi - \frac{\pi}{2}\right)\right] \quad (5.8)$$

$$E_\phi \approx 0 \quad (5.9)$$

where r is the radial distance at the farfield, $H_p^{(2)}$ is the hankel function of the second kind, $H_p^{(2)'}$ is its derivative and $\epsilon_p = 1$ when $p = 0$, else $\epsilon_p = 2$.

Equation (5.8) is computed in Figure 5.4 which is compared to measured results. The results show that bending the antenna over a small cylinder of radius 'a' result in larger back lobe radiation in the H-plane. The authors mention that similar results are obtained for the E-plane but is not reported in their research.

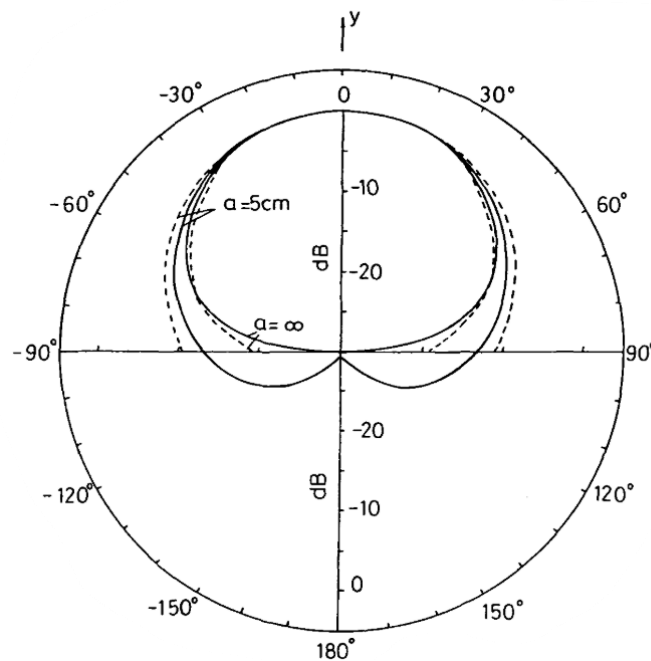


Figure 5.4: Normalized radiation pattern comparison between the theoretical analysis (solid curves) and measured results (broken curves) for TM_{01} mode [3].

In 1989, Kwai-Man Luk et al. [4] performed a similar analysis but for thin substrates much smaller than the operating wavelength and radius of curvature when bending. Their analytical analysis includes the investigation of the antenna's radiation pattern, input impedance and Q-factor.

The TM_{10} and TM_{01} modes for the E and H plane were analysed for their radiation patterns which is shown in Figure 5.5 and Figure 5.6. The results for the TM_{10} mode showed that there was significant back lobe radiation for large angles of bending. It was also found that increasing the dielectric constant of the substrate while bending the antenna would increase the back lobe radiation. The radiation patterns for the TM_{01} mode remain directive while having significantly less radiation at the back lobe when compared to the TM_{10} modes. This is expected since (5.5) implies that for TM_{01} modes the antenna's resonant frequency is not affected by antenna bending.

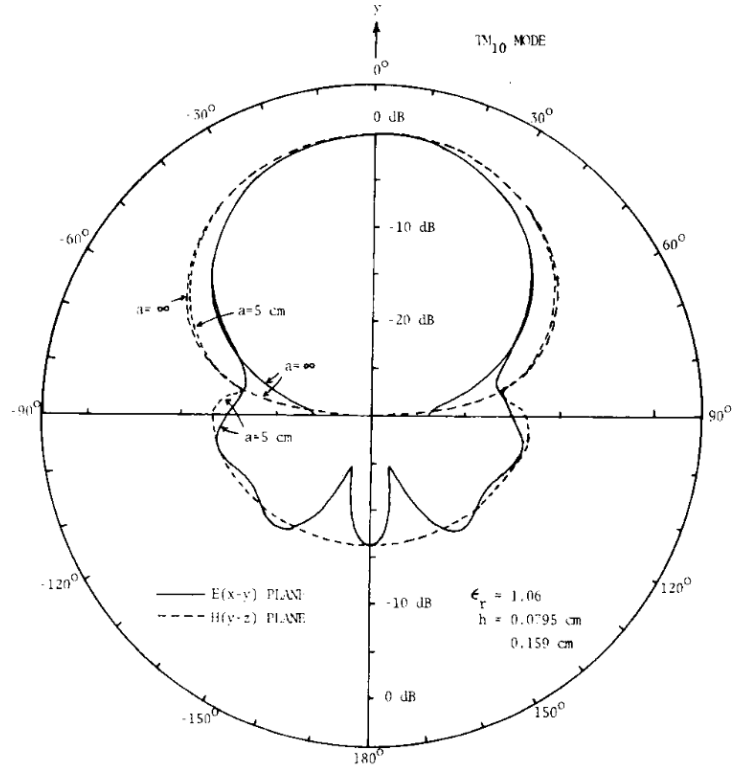


Figure 5.5: Normalized power pattern plot showing the E and H plane for the TM_{10} mode with $\epsilon = 1.06$ [4].

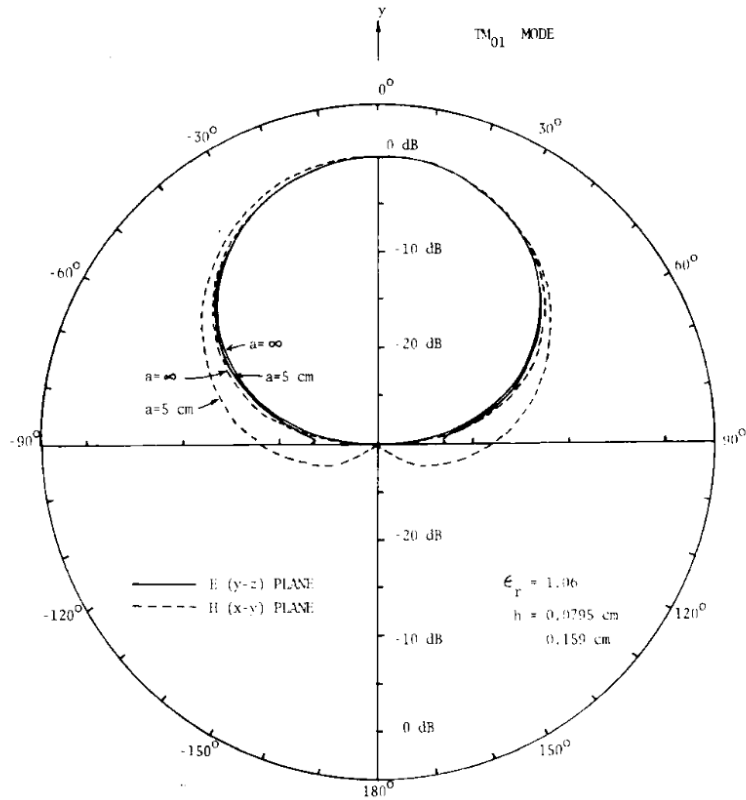


Figure 5.6: Normalized power pattern plot showing the E and H plane for the TM_{01} mode with $\epsilon = 1.06$ [4].

It was also found that for the TM_{01} mode the antenna's Q-factor decreases (BW increase) as the bending angle increases. However, for the TM_{10} mode, the Q-factor increases (BW decrease) as the bending angle increases. This was tested for two cases where the antenna's substrate has a dielectric constant of $\epsilon_r = 2.32$ and 9.8.

The input impedance of the antenna is also shown to be affected by antenna bending. The input impedance is calculated analytically by assuming a coaxial feed to the patch antenna which is modelled as a current ribbon which is represented by the three dotted lines in Figure 5.3 as an arc length w . The analysis and derivation for the input impedance can be found in [4]. The results show that for the TM_{10} mode the resonant resistance increases as the bending angle increases. However, it decreases for a $\epsilon_r = 1.06$. For the TM_{01} mode the resonant resistance decreases as the bending angle increases.

More recently modelling and investigating antenna bending has been of great interest due to the development of wearable textile antennas. In 2016, Sanjari et al. [7] derived an equation for tensile strain in bending operations for textile patch antennas. Tensile strain causes stretching or elongation to occur during antenna bending. Thus, they incorporated this term into the cylindrical cavity model to derive the resonant frequencies for E and H-plane bending. It was found that incorporating tensile strain into the cylindrical cavity model, to see the change in the resonant frequency during bending, agreed well with simulations and measurements. The results show that for large bending angles along the E-plane the resonant frequency decreases if elongation effects are considered as shown in Figure 5.7, where f_0 is the resonant frequency and Δf is the change in resonant frequency due to bending radius.

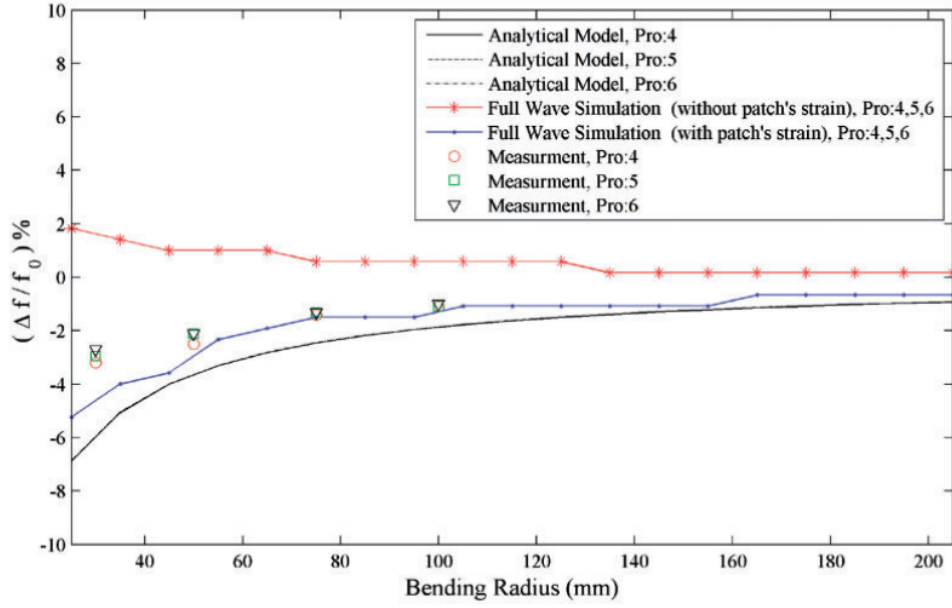


Figure 5.7: Percent change of the resonant frequency while varying the bending radius along the textile patch antenna's length [7].

Following their research, in 2018 L. Song and Y. Rahmat-Samii [5] produced a comprehensive analysis of textile patch antenna bending using the cylindrical cavity model, a lumped element circuit model, a full wave simulation analysis and measurements from a textile patch antenna prototype. The cylindrical cavity model analysis presented in [5] is based on the work by C Krowne [2]. The radius of the concentric cylinders is a and $a + h$ as shown in Figure 5.8(a), where W and L is the width and length of the rectangular patch and h is the height of the cavity. The radiator and ground plane are modelled as perfect electric conductors (PEC), while the edges of the patch are modelled as magnetic walls (PMC).

Figure 5.8(b) and Figure 5.8(c) show the antenna bent along the E and H plane. Antenna bending in this configuration is defined as the angle 2θ of the sector that is formed between the edges of the radiating element. The orientation of which bending occurs is along the z-axis.

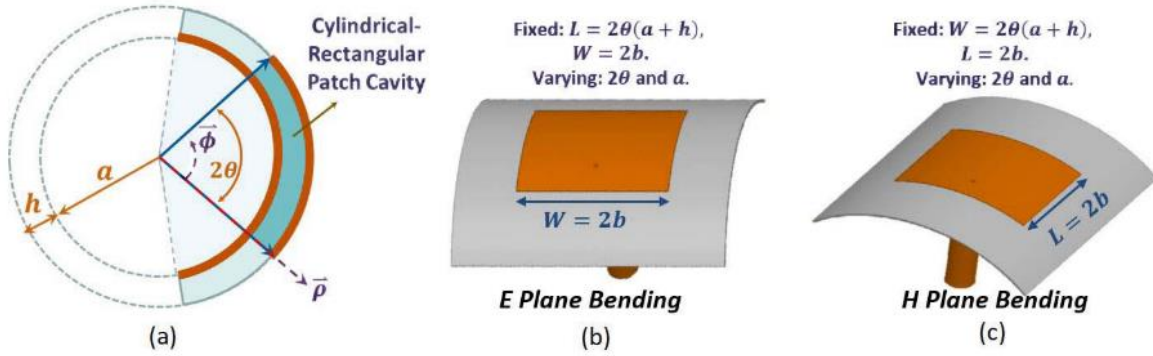
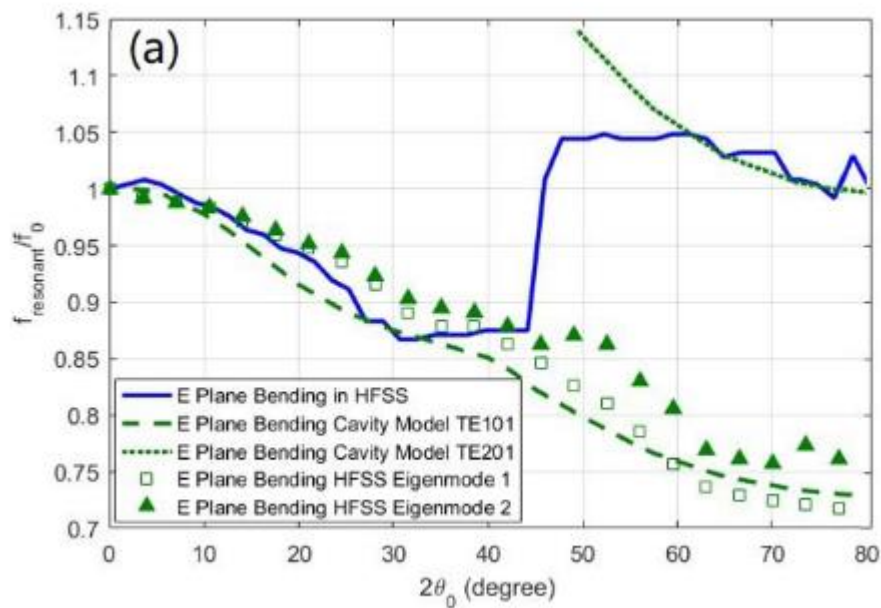


Figure 5.8: Cylindrical-rectangular cavity model showing the (a) geometry of the patch antenna bent on a cylindrical surface and the bending configuration for (b) E-plane bending and (c) H-plane bending [5].

The results in Figure 5.9 show that for E-plane bending the cylindrical-rectangular cavity model predicts that the resonant frequency decreases with increasing bending angles. When compared to the full wave and measured results, the results only agree with the cavity model up to a bending angle of 40° . For angles greater than 40° the resonant frequency begins to shift upwards as shown in Figure 5.9(a) and Figure 5.10. For H-plane bending the resonant frequency is not shifted significantly as bending angle increases. The measured, simulated and cavity model results shown in Figure 5.9(b) and Figure 5.10 agree with each other for H-plane bending.



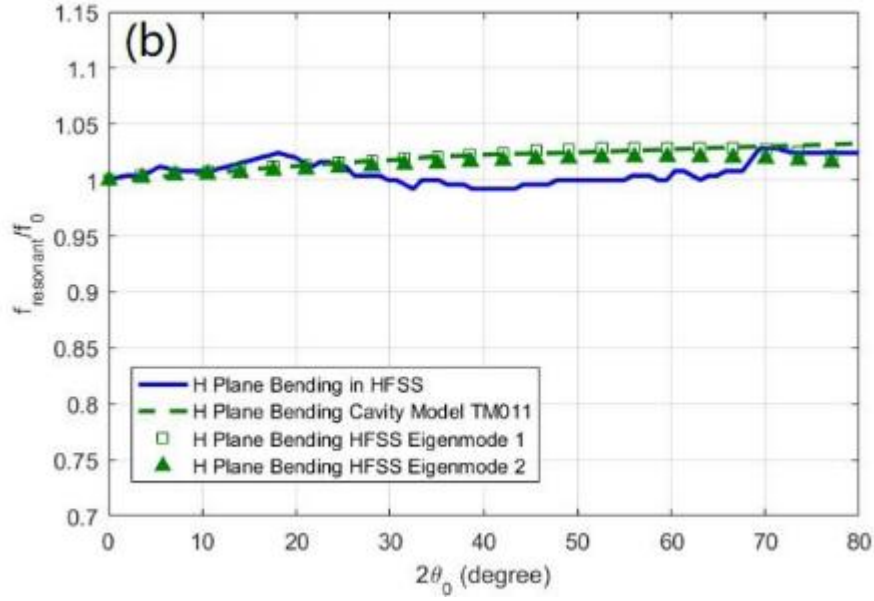


Figure 5.9: Normalized frequency shift plotted against antenna bending angle using the cylindrical-rectangular cavity model, HFSS port driven full wave simulation and HFSS eigenmode simulation for (a) E-plane bending and (b) H-plane bending [5].

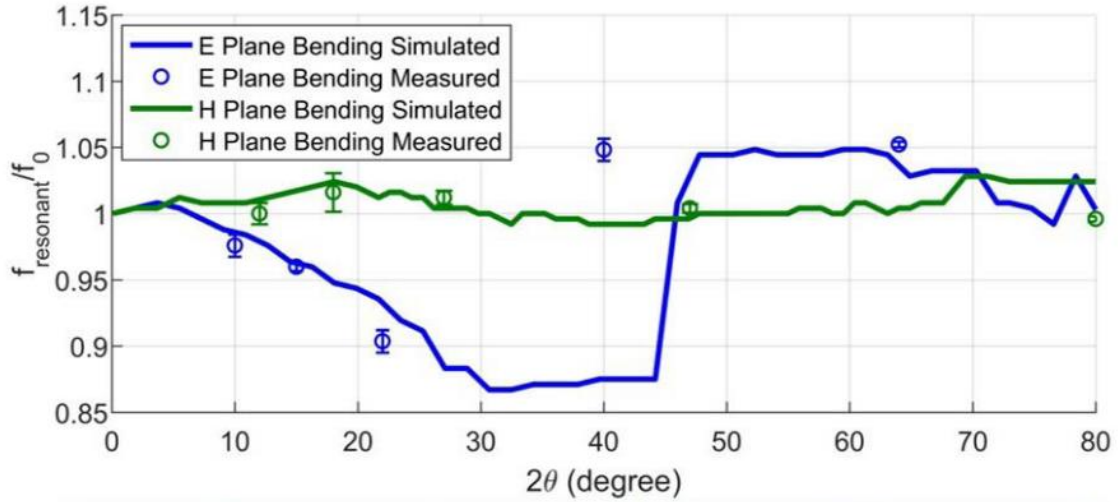


Figure 5.10: Measured normalized frequency shift of the patch antenna prototype compared with the simulated results for E and H-plane bending [5].

The patch antenna's bandwidth was also investigated. It was found that for E-plane bending the antenna's bandwidth would gradually decrease for increasing angles of bending, while for H-plane bending the bandwidth remained mostly unchanged.

R (mm)	2θ (E Plane)	2θ (H Plane)	Bandwidth(E Plane)	Bandwidth (H Plane)
35	64°	80°	1.03%	1.26%
60	40°	47°	1.01%	1.25%
105	22°	27°	1.21%	1.29%
160	15°	18°	1.31%	1.27%
250	10°	12°	1.28%	1.27%
flat	0°	0°	1.25%	1.25%

Table 5.1: Measured -10 dB bandwidth for the patch antenna bent over a cylinder of various radii [5].

In 2015, T. Bird presented the Cylindrical Aperture Array Model (CAAM) for a microstrip patch antenna placed on a cylindrical surface. The model for the analysis is shown in Figure 5.11, where the patch antenna is modelled as a dual-aperture array on a conducting cylinder. The radius of the cylinder R_0 is assumed to be much larger than the width w , height h , and length l of the patch. The electric field from each aperture is polarized in the radial ρ direction and the magnetic current at each aperture is given by $\mathbf{M} = -\hat{\phi} \times (\hat{\rho} E_0)$. In this analysis the height of the patch antenna is assumed to be small and thus the two apertures are modelled as two uniform magnetic line sources of length w separated by a distance l radiating in the presence of a large cylinder.

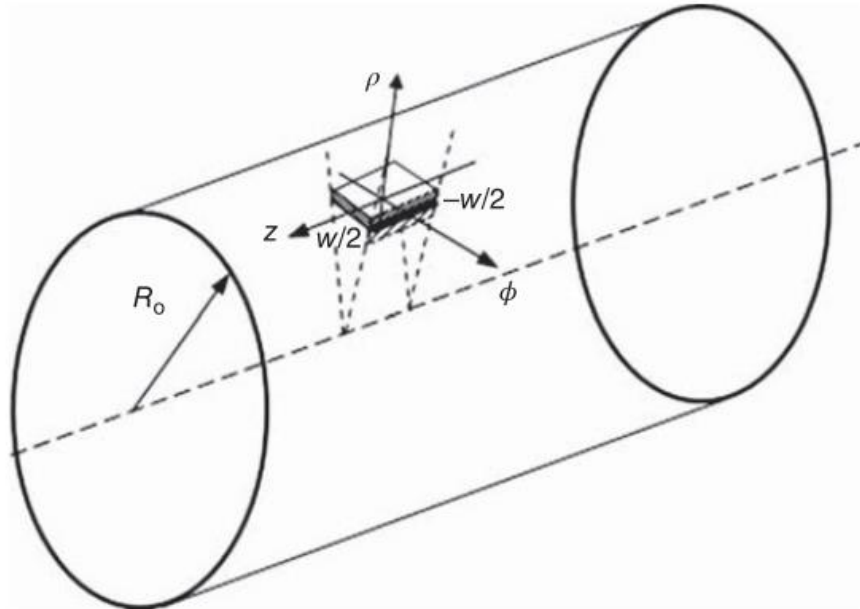


Figure 5.11: Microstrip patch antenna mounted on a cylinder to model antenna bending [8].

The electric field components radiating at the far-field is given by James R. Wait [9] as

$$dE_\phi(r, \theta, \phi) \sim -dM_z \frac{1}{(2\pi)^2 R_0} \frac{e^{-jkr}}{r} e^{jkz' \cos \theta} \times \sum_{n=-\infty}^{\infty} j^n \frac{e^{jn(\phi-\phi')}}{H_n^{(2)'}(\gamma)} \quad (5.10)$$

$$dE_\theta(t, \theta) = 0 = dE_r(t, \theta) \quad (5.11)$$

where $H_n^{(2)'}(\gamma)$ is the derivative of the Hankel function of the second kind of order n , $\gamma = kR_0 \sin \theta$ and z' corresponds to the source co-ordinate.

By calculating the field radiated by each aperture at $\phi' = +\Delta\phi/2$ and $\phi' = -\Delta\phi/2$, the total electric field is calculated as the superposition of the fields radiated by each aperture given as

$$E_\phi(r, \theta, \phi) = -\frac{hE_0}{2\pi^2 R_0} \frac{e^{-jkr}}{r} S\left(\frac{kW}{2} \cos(\theta)\right) \left(\frac{1}{H_n^{(2)'}(\gamma)} + 2 \sum_{n=1}^{\infty} \frac{j^n}{H_n^{(2)'}(\gamma)} \cos\left(\frac{n\sqrt{\epsilon}L}{2R}\right) \cos(n\phi) \right) \quad (5.12)$$

T. Bird plotted the radiation pattern for a patch antenna on a cylinder with a radius $R_0 = 50\lambda$ and shown in Figure 5.12. The dimensions of the patch antenna are given as $w = \lambda/4$, $l = \lambda/2\sqrt{\epsilon_r}$ and $h = 0.005\lambda$. The dielectric constant of the substrate is $\epsilon_r = 2.54$. Using (2.9) the normalized radiation pattern plot is computed for the E and H-plane. The E-plane corresponds to when $\theta = 90^\circ$, while the H-plane corresponds to when $\phi = 0^\circ$.

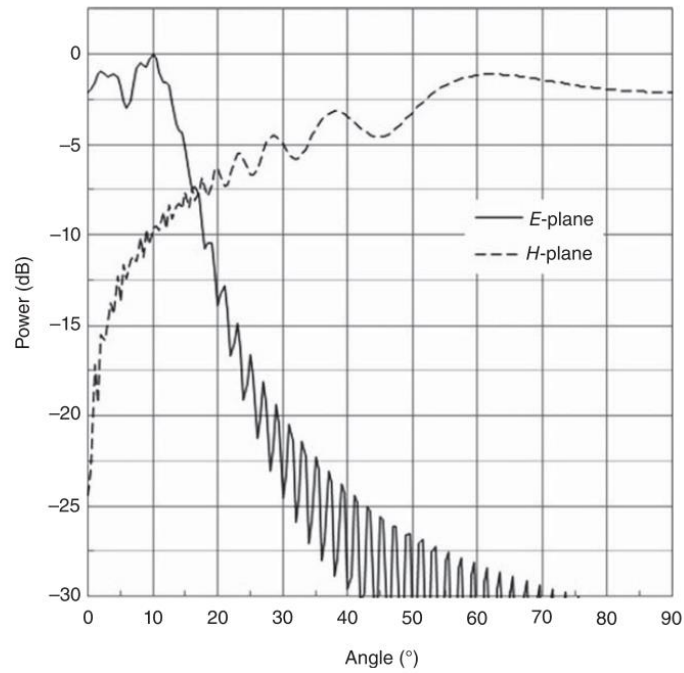


Figure 5.12: Radiation pattern displaying the E-plane and H-plane for a microstrip patch antenna on a cylinder.

A comparison of the radiation pattern is made to a planar patch antenna with the same dimensions in Figure 5.13 to highlight the key differences.

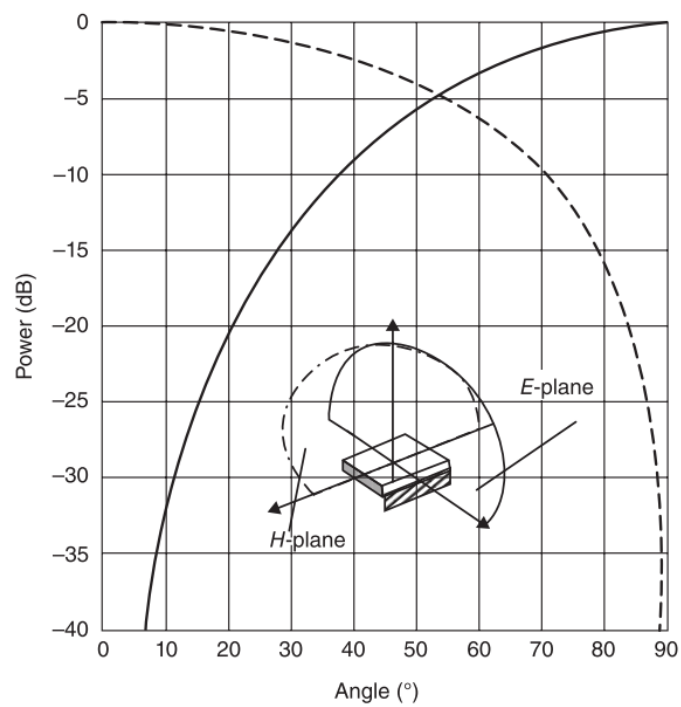


Figure 5.13: Radiation pattern displaying the E-plane and H-plane for a planar microstrip patch antenna.

The significant difference when comparing Figure 5.12 and Figure 5.13 is the ripples in the pattern. This is due to the interference between the space waves produced by the apertures and the creeping waves that are excited around the conductive cylinder [8]. A cylindrically bent microstrip patch antenna does not have a ground plane that wraps around the entire cylinder. This reduces the effects of creeping waves.

Overall, some conclusions can be made from the cylindrical-rectangular cavity model and the aperture array model on a conducting cylinder to model patch antenna bending.

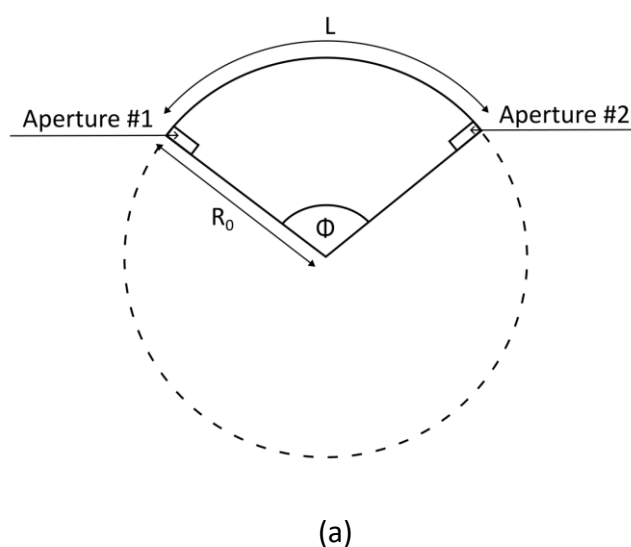
- Bending along the antenna's E-plane detunes the antenna's resonant frequency considerably when compared to bending along the antenna's H-plane.
- Increasing the bending angle alters the radiation pattern resulting in large back lobe radiation.
- Bandwidth is affected for both the TM_{01} and TM_{10} modes. As bending angle increases the bandwidth increases for TM_{01} modes and decreases for TM_{10} modes.
- Tensile properties of the substrate can affect the resonant behaviour during bending but can be ignored in the analysis if the materials designing the patch antenna have a low strain, which depends on the Young's modulus of both the substrate, radiating element and ground plane [5].
- The aperture array model can be used to approximate the far-field radiation for a bent microstrip patch antenna placed on a conductive cylinder.
- Creeping waves can affect the radiation pattern of the patch antenna when bent over a conductive cylinder.

5.2 Aperture Array Model for Patch Antenna Bending

The Aperture Array Model was used in Chapter 4 to approximate the far-field characteristics of the Trapezoidal Patch Antenna (TPA). In this section the aperture array model for a TPA is applied onto a conductive cylinder. By modelling the patch antenna's apertures onto a conductive cylinder, it is possible to form a simple yet insightful analysis for antenna bending.

In Chapter 4, radiating apertures are modelled around the edges of the TPA with the ground plane. For simplicity of the analysis, the fields across the apertures are modelled as uniform. Since the electric field only exists over the aperture and zero elsewhere, a perfect electric conductor (PEC) is introduced behind the aperture. The equivalent source on the aperture produces an image or an electric field which is twice the strength of the original source.

By modelling the separation distance between the two apertures as a constant of arc length L over the cylinder, antenna bending can be represented as the change in the angle Φ or radius R_0 that forms the sector between the two apertures. The geometry of the model is shown in Figure 5.14. The dotted lines represent the rest of the cylinder. As the length L remains fixed, an inverse relationship can be established between the radius R_0 of the cylinder and the bending angle Φ , that is if the size of the radius of the cylinder increases the bending angle decreases.



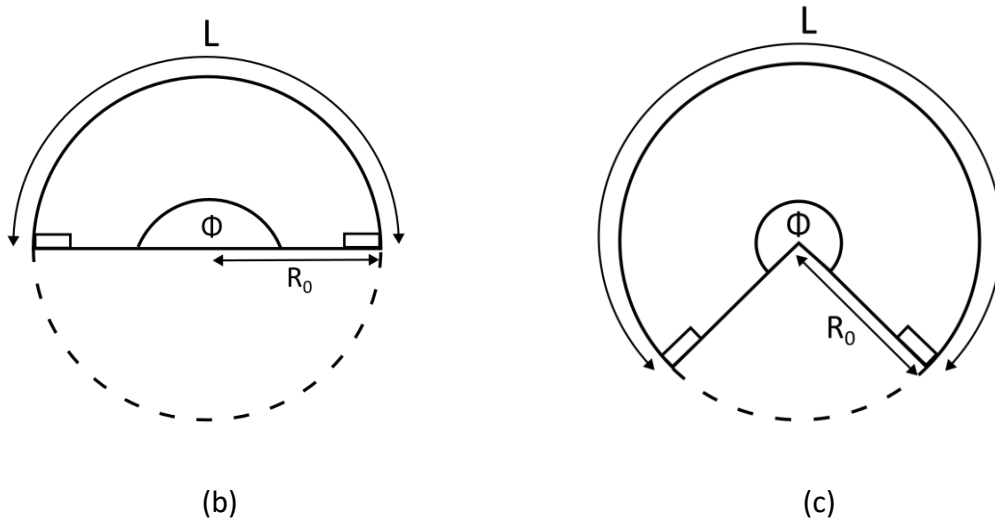


Figure 5.14: Model representing antenna bending as a sector of a cylinder where (a) represents a sector with a bending angle $\Phi < 180^\circ$, (b) $\Phi = 180^\circ$ and (c) $\Phi > 180^\circ$.

The sources for the apertures on the cylinder are considered to radiate in the presence of a large cylinder using (5.10) which is a high frequency approximation for the elemental field radiated by a line source on the axis of a cylinder [8].

The magnetic current at each aperture is given by

$$\mathbf{M}_s = -\hat{\phi} \times (\hat{\rho} E_0) \quad (5.13)$$

The main radiating apertures appear as two uniform magnetic line sources of length W_1 and W_2 which are separated by a distance L . Both apertures are parallel to the z axis of the cylinder. There also exists a uniform magnetic line source of length W_{ext} which corresponds to the horizontal component along the side apertures of the trapezoid of length L_d .

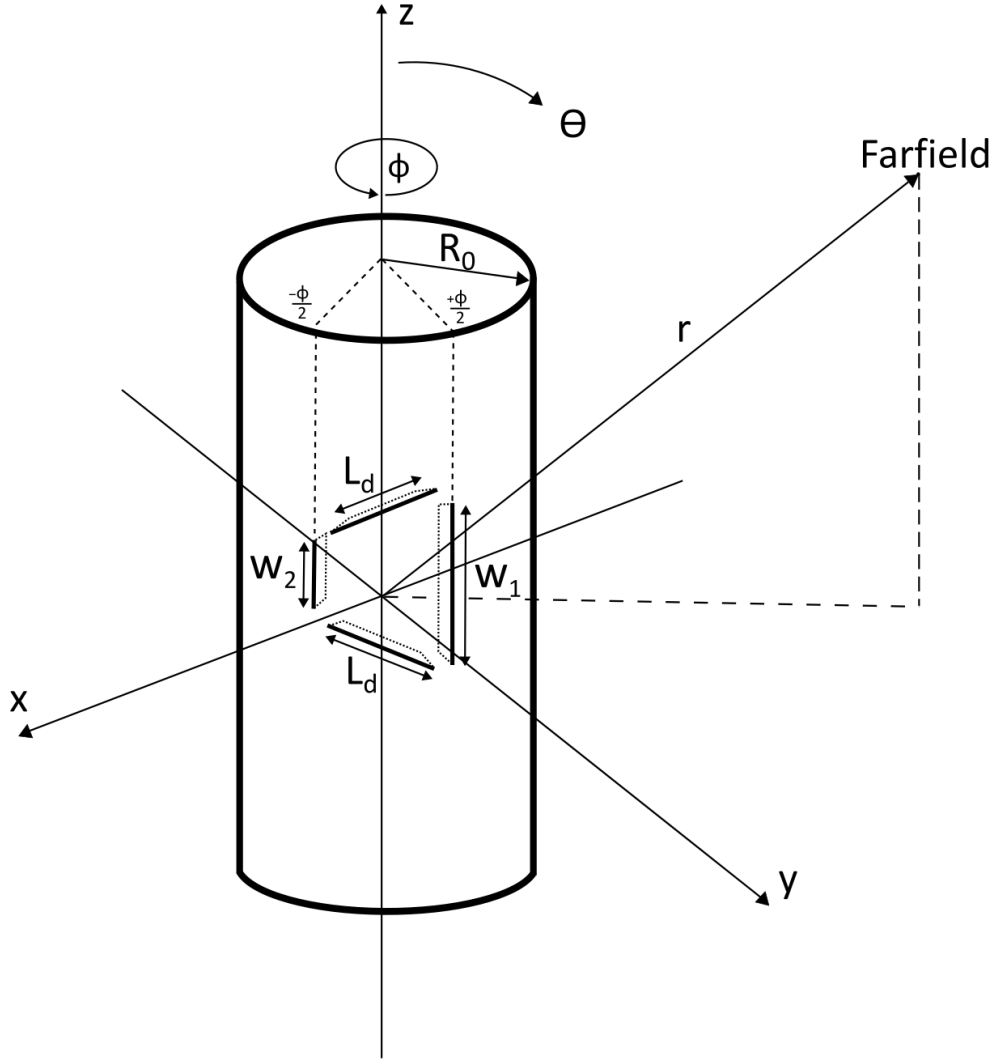


Figure 5.15: Aperture array model for the TPA on a cylinder.

For this analysis, it is assumed that there are two uniform magnetic line sources at each aperture placed at $\Delta\phi/2$ and $-\Delta\phi/2$. The sources on the cylinder are assumed to be separated by the angle $\Delta\phi = L/R_0$. An additional line source is assumed for the side apertures at $\phi = 0^\circ$. In Chapter 4 it was shown that the side apertures are regarded as ‘non-radiating apertures’, however since the side apertures for the TPA are angled with horizontal and vertical components they are regarded as partial radiators. Therefore, an additional line source is introduced to account for the horizontal component of the side apertures of length W_{ext} .

Combining the electric field contributions $E_\phi^{W_1}$, $E_\phi^{W_2}$ and $E_\phi^{W_{ext}}$ of each aperture of width W_1 , W_2 and W_{ext} using equation (5.10) results in the total electric field E_ϕ^T

$$E_\phi^T = E_\phi^{W_1} + E_\phi^{W_2} + 2E_\phi^{W_{ext}} \quad (5.14)$$

where $E_\phi^{W_1}$, $E_\phi^{W_2}$ and $E_\phi^{W_{ext}}$ are

$$E_\phi^{W_1} = -\frac{hE_0}{2\pi^2 R_0} \frac{e^{-jkr}}{r} \int_{-\frac{W_1}{2}}^{\frac{W_1}{2}} e^{jkz' \cos \theta} dz' \sum_{n=-\infty}^{\infty} \frac{j^n e^{jn(\phi - \frac{\Delta\phi}{2})}}{H_n^{(2)'}(\gamma)} \quad (5.15)$$

where $\gamma = kR_0 \sin \theta$

$$E_\phi^{W_2} = -\frac{hE_0}{2\pi^2 R_0} \frac{e^{-jkr}}{r} \int_{-\frac{W_2}{2}}^{\frac{W_2}{2}} e^{jkz' \cos \theta} dz' \sum_{n=-\infty}^{\infty} \frac{j^n e^{jn(\phi + \frac{\Delta\phi}{2})}}{H_n^{(2)'}(\gamma)} \quad (5.16)$$

$$E_\phi^{W_{ext}} = -\frac{hE_0}{2\pi^2 R_0} \frac{e^{-jkr}}{r} \int_{-\frac{W_{ext}}{2}}^{\frac{W_{ext}}{2}} e^{jkz' \cos \theta} dz' \sum_{n=-\infty}^{\infty} \frac{j^n e^{jn(\phi)}}{H_n^{(2)'}(\gamma)} \quad (5.17)$$

After solving the integrals, E_ϕ^T can be simplified to

$$E_\phi^T = -\frac{hE_0}{2\pi^2 R_0} \frac{e^{-jkr}}{r} \sum_{n=-\infty}^{\infty} \frac{j^n e^{jn\phi}}{H_n^{(2)'}(\gamma)} \left(W_1 S\left(\frac{kW_1}{2} \cos \theta\right) + W_2 S\left(\frac{kW_2}{2} \cos \theta\right) + 2W_{ext} S\left(\frac{kW_{ext}}{2} \cos \theta\right) \right) \left(e^{\left(\frac{jn\Delta\phi}{2}\right)} + e^{\left(-\frac{jn\Delta\phi}{2}\right)} + 1 \right) \quad (5.18)$$

$$E_\phi^T = -\frac{hE_0}{\pi^2 R_0} \frac{e^{-jkr}}{r} \left(W_1 S\left(\frac{kW_1}{2} \cos \theta\right) + W_2 S\left(\frac{kW_2}{2} \cos \theta\right) + 2W_{ext} S\left(\frac{kW_{ext}}{2} \cos \theta\right) \right) \sum_{n=-\infty}^{\infty} \frac{j^n e^{jn\phi}}{H_n^{(2)'}(\gamma)} \left(\cos\left(\frac{nL}{2R_0}\right) + 1 \right) \quad (5.19)$$

Using the Dirichlet kernel, the series term can be expanded as

$$\begin{aligned}
E_{\phi}^T = & -\frac{hE_0}{\pi^2 R_0} \frac{e^{-jkr}}{r} \left(W_1 S\left(\frac{kW_1}{2} \cos \theta\right) + W_2 S\left(\frac{kW_2}{2} \cos \theta\right) \right. \\
& \left. + 2W_{ext} S\left(\frac{kW_{ext}}{2} \cos \theta\right) \right) \\
& \left(\frac{1}{H_n^{(2)'}(\gamma)} + 2 \sum_{n=1}^{\infty} \frac{j^n}{H_n^{(2)'}(\gamma)} \cos\left(\frac{nL}{2R_0}\right) \cos(n\phi) \right) \left(\frac{1}{H_n^{(2)'}(\gamma)} \right. \\
& \left. + 2 \sum_{n=1}^{\infty} \frac{j^n}{H_n^{(2)'}(\gamma)} \cos(n\phi) \right) \quad (5.20)
\end{aligned}$$

The average power density, radiation intensity, radiated power P_{rad} and maximum directivity D_{max} is calculated using (2.7), (2.9), (2.8) and (2.10). Due to the complexity in solving the equation for the radiated power, a numerical series approximation method is used to solve the integral which is outlined by C. Balanis in [10].

The normalized radiation intensity is plotted for the antenna's E and H-plane while varying the size of the cylinder as shown in Figure 5.16 and Figure 5.17. To investigate how the antenna's far-field radiation varies as the bending angle or radius changes, the radiation intensity is plotted for a bending angle of 30° to 240° which is equivalent to the radius of the cylinder varying from 90.1 mm to 11.3 mm. The antenna's E-plane corresponds to $\theta = 90^\circ$, while the H-plane corresponds to $\phi = 0^\circ$.

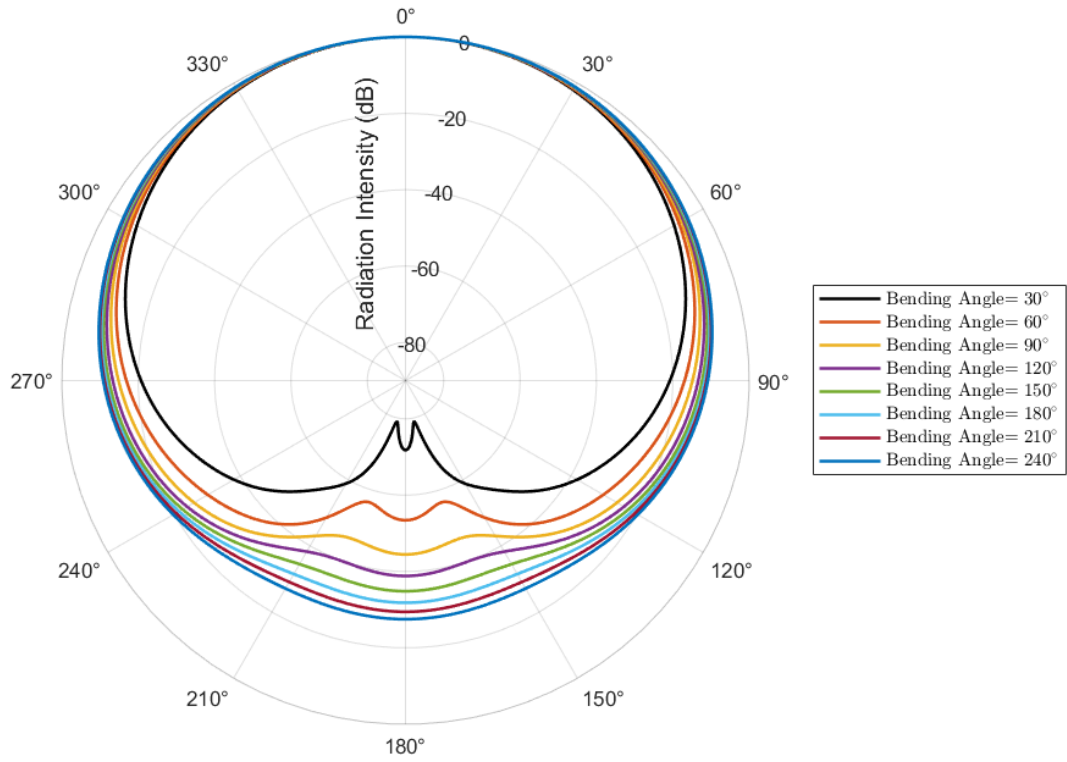


Figure 5.16: E-plane far-field radiation intensity pattern for the TPA bent over a conductive cylinder for $0^\circ \leq \phi \leq 360^\circ$.

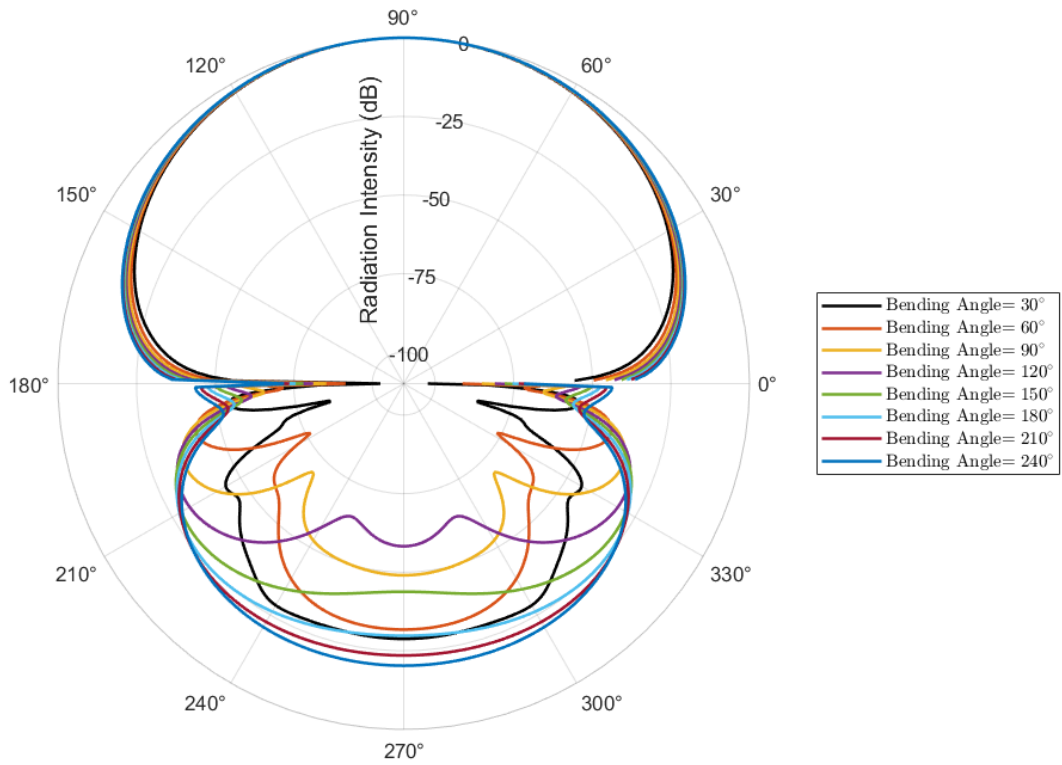


Figure 5.17: H-plane far-field radiation intensity pattern for the TPA bent over a conductive cylinder for $0^\circ \leq \theta \leq 360^\circ$

The antenna's E-plane is shown in Figure 5.16, where for bending angles between 30° and 240° the major and back lobe appear to get broader for larger bending angles. The H-plane plot in Figure 5.17 shows similar behaviour with the major and side lobe pattern broadening for higher angles of bending. However, the radiation intensity at the back lobe varies in magnitude from a bending angle of 30° by initially decreasing and then increasing for bending angles greater than 120° .

The computed directivity for bending angles between 1° and 240° by varying the radius of the cylinder is shown in Figure 5.18. For a small bending angle, the directivity starts to diverge exponentially to large values. This is because for very small bending angles, the CAAM models the conductive cylinder to be extremely large. A 1° bending angle corresponds to the antenna laying on a cylinder with a radius of approximately 3 meters. For a small bending angle of 30° , the analysis predicts a high directivity of 12.6 dBi which decreases to 10.9 dBi for a bending angle of 240° .

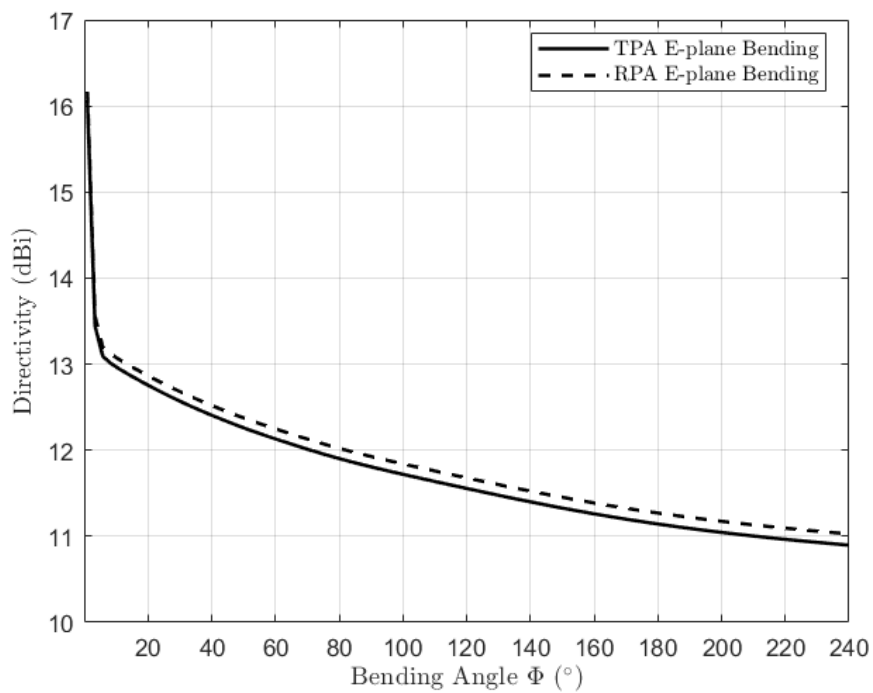


Figure 5.18: Computed directivity using the cylindrical aperture array model for various bending angles.

Comparing the directivity in Figure 5.18 to the equivalent calculated directivity for a flat patch antenna in Figure 4.22 there is a clear difference. In Chapter 4, the AAM for the TPA predicts a much lower directivity for the TPA with an aperture width $W_2 = 5$ mm. The Dual and Quad-AAM predicts a directivity of 5.8 dBi and 6.5 dBi. This is due to the model representing the

entire cylinder with an infinite conductivity. For a large cylinder the apertures can be assumed to be placed on a large cylindrical ground plane. Since the region surrounding the apertures is conductive and extends for a significant distance, the directivity of the antenna is enhanced.

For bending angles $\Phi \ll 30^\circ$ the cylinder becomes exceedingly large, and ripples will start to form throughout the pattern. An example of this phenomenon is shown in Figure 5.12. This is due to the interference between the main lobe radiation produced by the patch and the creeping waves excited around the conductive cylinder [8].

Figure 5.18 also compares the performance of the RPA to the TPA. It can be shown that the decrease in directivity for increasing bending angle is almost the same for both antennas with the RPA having a slightly higher directivity by 0.1 dBi across all bending angles greater than 8° .

Overall, the CAAM attempts to model bending to investigate the far-field characteristics for the TPA. It was found that increasing the bending angle of the antenna decreases its directivity, which means that the antenna radiates less in the direction of maximum radiation. For very small antenna bending angles on a large conductive cylinder, parameters such as the antenna's radiation intensity and directivity exhibit different radiation characteristics when compared to the planar aperture array models presented in Chapter 4.

For modelling the antenna for on-body applications such as a human arm, the size of the cylinder can be modelled to approximate the typical bending angles that would occur when the antenna is placed conformally on the arm. Typical values for a radius R_0 of a cylinder size which is consistent to a human arm can be calculated from the average arm circumference which is shown to be 33.6 ± 7.5 cm [11]. Therefore, the radius of the cylinder will range between 11.9 mm to 53.5 mm, which is equivalent to a bending angle between the range of 51° to 227° .

5.3 Modelling Resonant Frequency Detuning

Using the cylindrical-rectangular cavity model which was previously presented in [2], the change in the patch antenna's resonant frequency for bending along the H-plane and E-plane can be analytically investigated. By treating the patch antenna as an isolated sector with appropriate boundary conditions and solving the auxiliary vector wave equation in cylindrical co-ordinates, the rectangular patch antenna's resonant frequencies can be solved for various bending angles.

Antenna bending in this analysis is characterized by the angle Φ subtended that forms the bending arc of the radiating element of the patch antenna. To solve equations (5.1)-(5.4) for a RPA, values for the antenna's resonant frequency, length L , width W , height h and dielectric constant of the substate ϵ_r must be known. The simulated values given in Table 4.2 for a RPA resonating at 2.4 GHz serves as a template for the required values.

Solving (5.1) for a resonant frequency at 2.4 GHz for a 0° bending angle requires an adjustment of the length L and width W of the antenna to $L = 49.4$ mm and $W = 49.4$ mm. By plotting (5.1) for various antenna bending angles Φ as shown in Figure 5.19, it can be shown that the Cylindrical-Rectangular Cavity Model predicts that for the TE^z mode or E-plane bending the resonant frequency increases up to 2.73 GHz for a bending angle of 240° . TM^z mode or H-plane bending does not alter the antenna's resonant frequency.

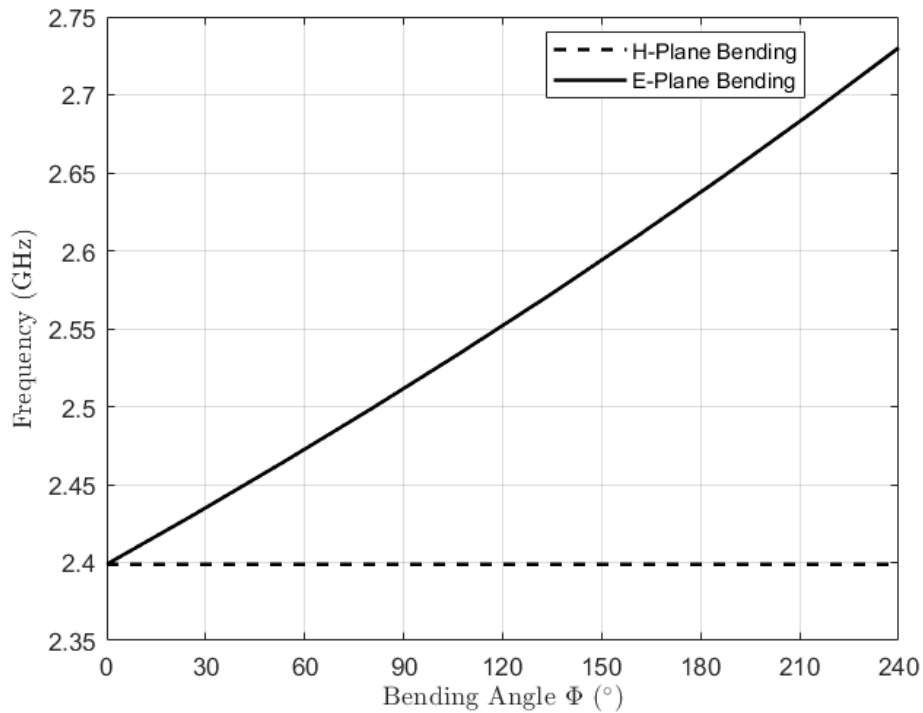


Figure 5.19: Resonant frequency shift for increasing bending angles using the cylindrical-rectangular cavity model for H-plane and E-plane bending.

This model is only valid for the RPA. Applying this model to the TPA is not possible since the cross-sectional width varies along the length of the trapezoidal radiating element. Nonetheless, the model provides insight to how bending along the different radiation planes for a patch antenna affects the resonant frequency.

Based on the insights from Chapter 4, for the TPA the non-radiating apertures along the diagonals of the patch become partial radiating apertures. Narrowing the aperture width W_2 caused the horizontal component of the diagonal apertures of length L_d to contribute more to the total electric field over a smaller surface area than in the RPA. Therefore, it is predicted that the TPA will have less resonant frequency detuning due to this redistribution of the electric field over the smaller area.

5.4 Full-Wave Simulation Analysis of Patch Antenna Bending

To further investigate patch antenna bending a full wave simulation comparison is presented in CST Microwave Studio between a Rectangular Patch Antenna (RPA) and Trapezoidal Patch Antenna (TPA). The performance parameters investigated are the shift in the antenna's resonant frequency, bandwidth, radiation efficiency, directivity and realized gain. The performance parameters while bending across the antennas H and E-plane are investigated. The dimensions used for the rectangular and trapezoidal patch antenna are the values given in Table 4.3, where the aperture width $W_2 = 52$ mm corresponds to the RPA while $W_2 = 5$ mm corresponds to a TPA. The frequency domain solver based on the Finite-Element Method (FEM) is used since the tetrahedral mesh enables efficient modelling of curved structures [12].

Antenna bending in this analysis is characterized by the angle subtended that forms the bending arc of the patch antenna. Previously in the CAAM, the bending arc spans the edges of the apertures, however for a patch antenna with a finite ground plane this can no longer be the reference. Therefore, the bending arc spans from the edges of the ground plane for both E and H-plane bending. Stretching and compression of the patch antenna while bending are assumed negligible for simplicity as this is said to be a reasonable approximation for thin substrates [5].

To begin the analysis a performance benchmark is set by simulating antenna bending for a rectangular patch antenna (RPA). This is used as a reference to compare antenna bending for a trapezoidal patch antenna. The dimensions used to investigate antenna bending for both the RPA and TPA is shown in Table 5.2.

Parameter	Symbol	Simulated Rectangular Patch Dimensions (mm)	Simulated Trapezoidal Patch Dimensions (mm)
Vertical Patch Length	L_v	49.0	52
Aperture #1 Width	W_1	52.0	52
Aperture #2 Width	W_2	52.0	5
Inset Recessed Distance	y_0	13.0	13.0
Feedline Width	W_0	10.5	10.5
Substrate Height	H_s	2.8	2.8
Inset Feed Gap Width	G_{tp}	2.53	2.23

Table 5.2: Simulated dimensions for a rectangular and trapezoidal patch antenna for operation at 2.4 GHz.

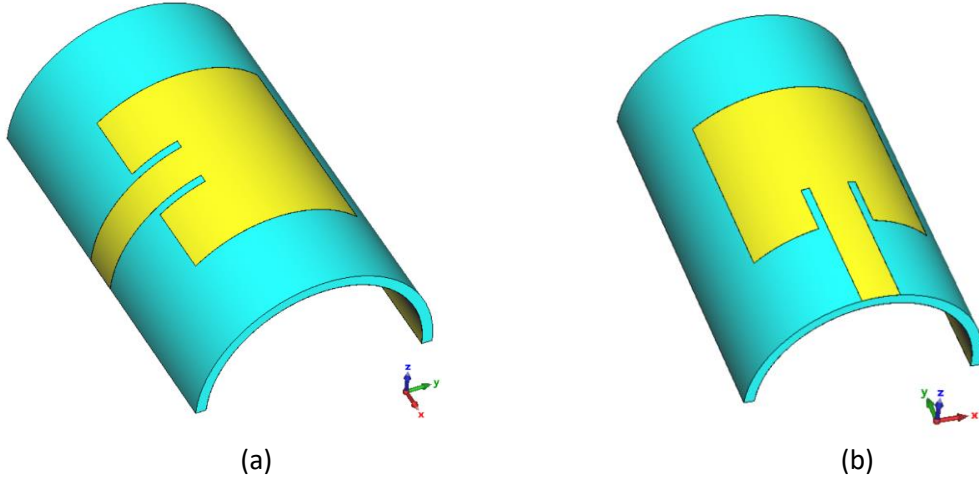


Figure 5.20: RPA cylindrically bent along its (a) width, YZ plane or E-plane and (b) length, XZ plane or H-plane.

Figure 5.20 shows the RPA bent along the H and E-plane. The bending angle for the antenna is increased from its flat condition and incremented in steps of 30° up to 180° .

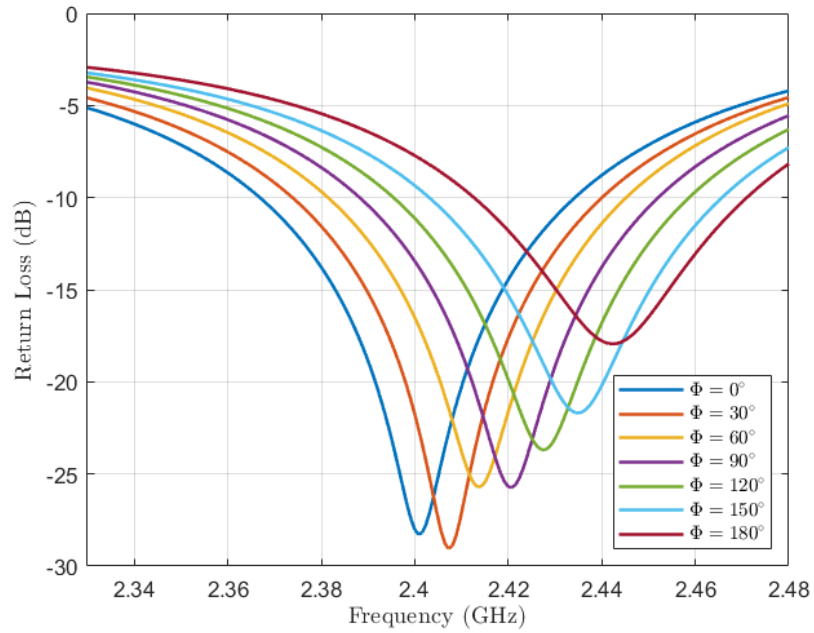


Figure 5.21: Simulated input reflection coefficient for various E-plane bending angles Φ of the RPA.

The simulated reflection coefficients for various E-plane bending angles from Figure 5.21 and Figure 5.22 shows a linear increase in the RPA's resonant frequency from 2.4 – 2.44 GHz. As the bending angle increases the impedance mismatch increases. The gradual shift in frequency and increasing mismatch is due to the change in the antenna effective length between the radiating edges as bending increases. The fringe fields at both apertures are moving closer together thus creating an electrically smaller structure. This changes both the antenna input impedance and resonant frequency.

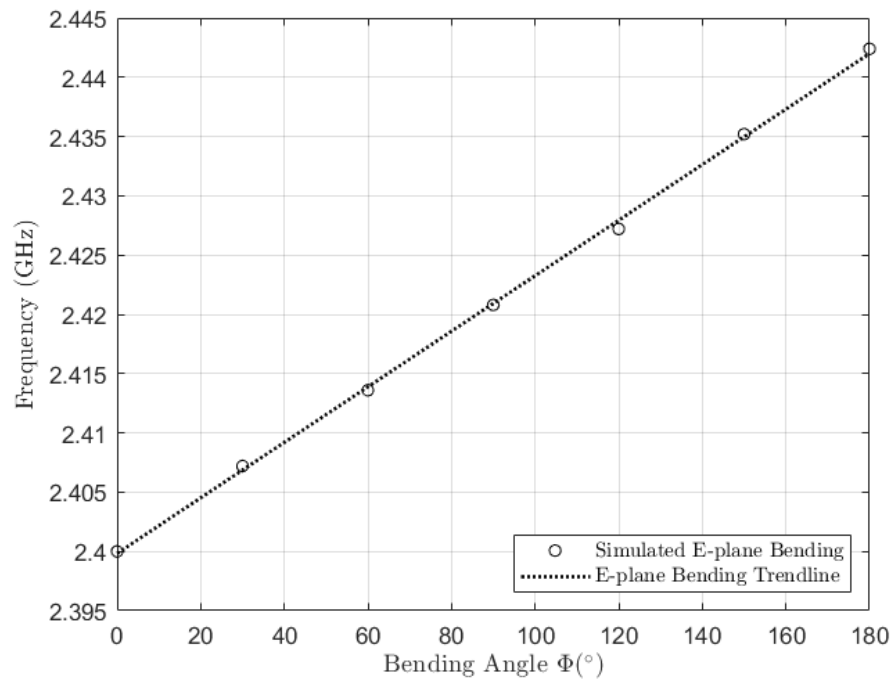


Figure 5.22: Change in the RPA's resonance frequency while bending along the YZ plane or E-plane.

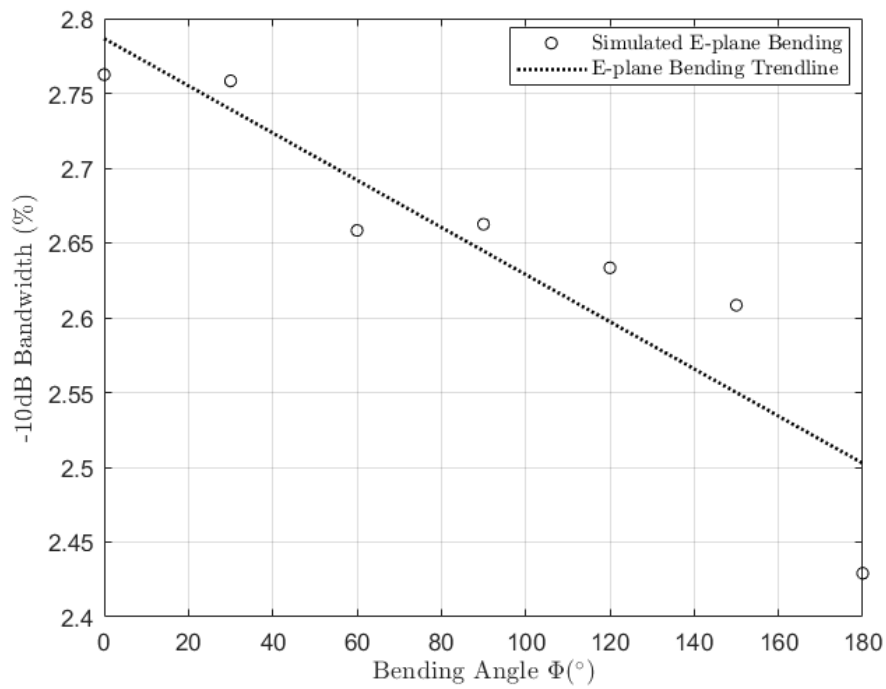


Figure 5.23: Change in the RPA's bandwidth while bending along the YZ-plane or E-plane.

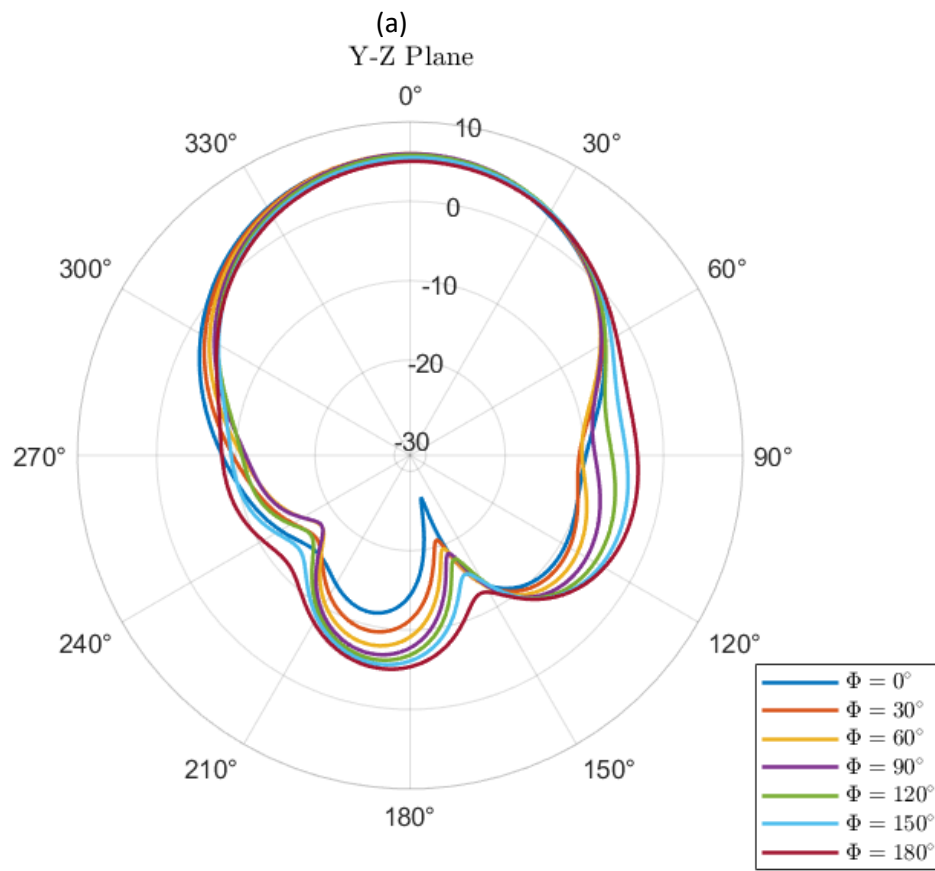
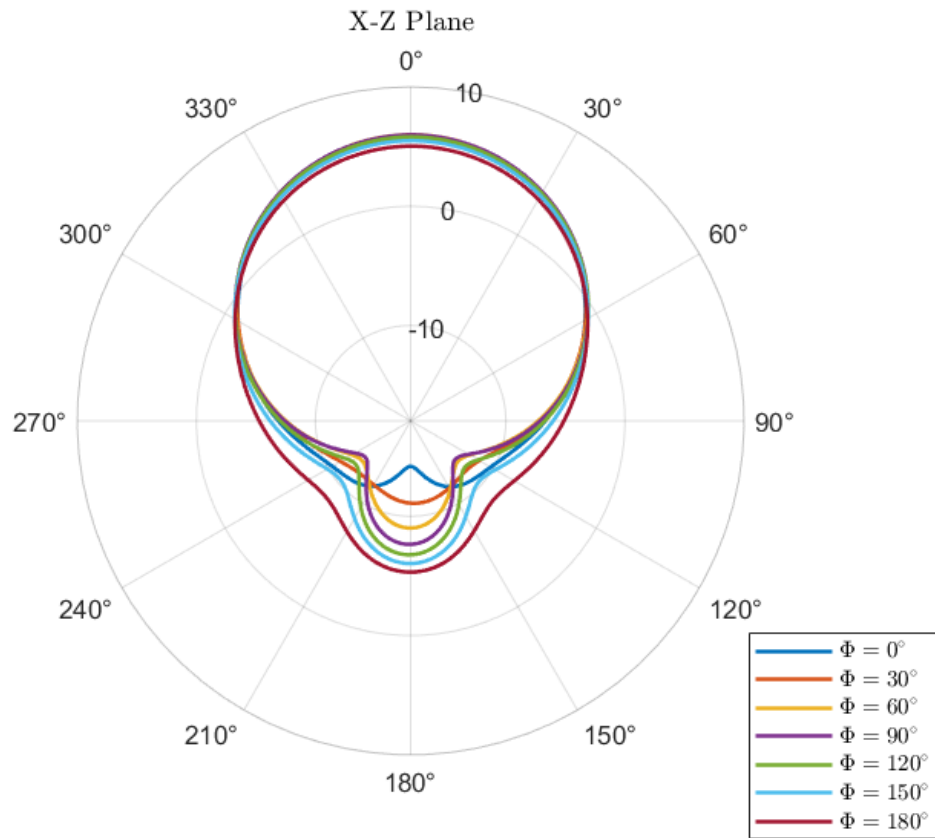


Figure 5.24: Simulated radiation pattern for (a) X-Z or H-plane and (b) Y-Z or E-plane while bending the antenna across its length or E-Plane.

Bending Angle Φ (°)	f_{res} (GHz)	–10 dB Bandwidth (MHz)	Total Radiation Efficiency (%)	Directivity (dBi)	Realized Gain (dBi)
0	2.400	2.76	57.7	8.14	5.73
30	2.407	2.76	56.7	8.38	5.91
60	2.414	2.66	56.4	8.49	6.00
90	2.421	2.66	56.6	8.45	5.98
120	2.427	2.63	57.2	8.25	5.82
150	2.435	2.61	57.9	7.88	5.50
180	2.442	2.43	58.4	7.38	5.04

Table 5.3: Performance Parameters Comparison for YZ-plane or E-Plane Bending Angles.

The antenna's far-field radiation patterns are plotted in terms of realized gain in Figure 5.24. Both the radiation patterns for the H and E-plane for E-plane bending show the RPA reducing in realized gain for increasing bending angles. The major lobe for both planes reduce in magnitude while the back lobe becomes larger and broader. The performance parameters are summarised in Table 5.3.

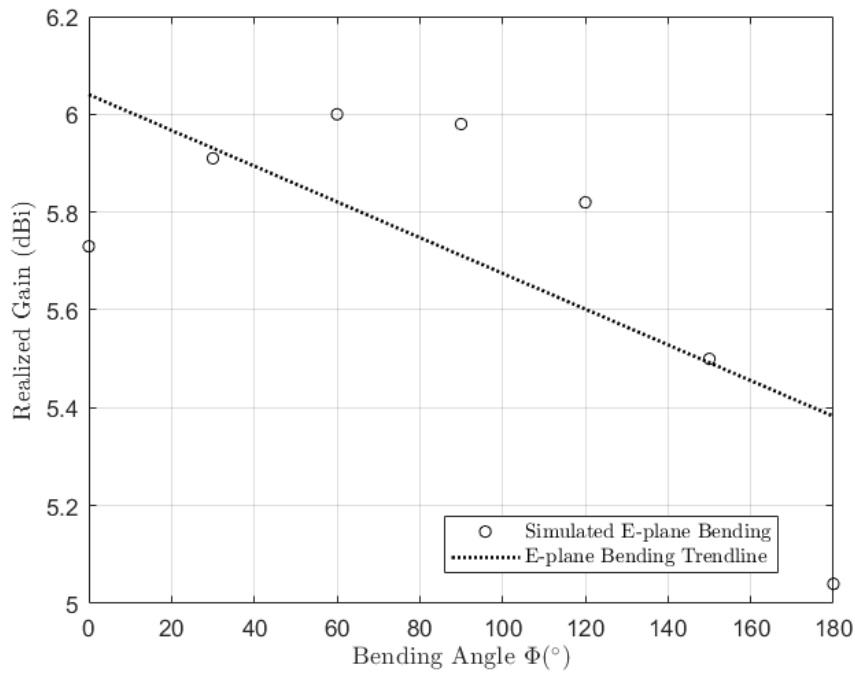


Figure 5.25: Change in realized gain while bending along the YZ plane or E-plane.

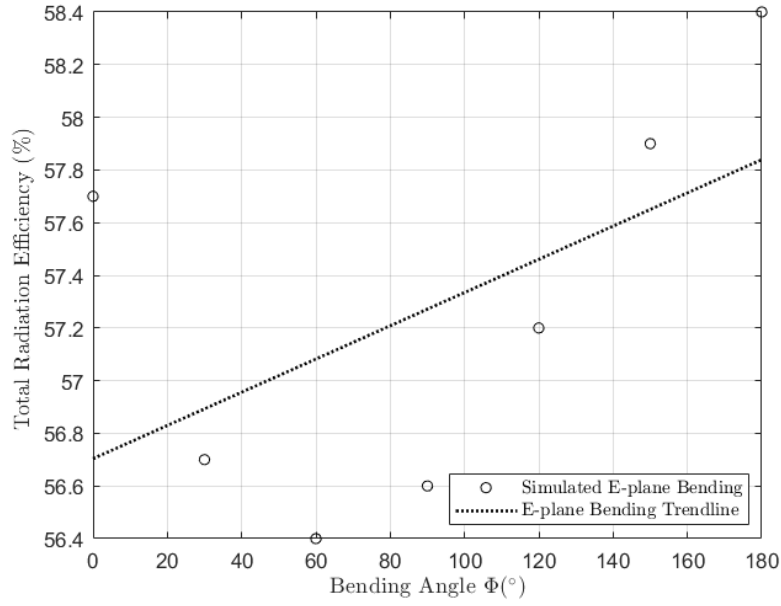


Figure 5.26: Change in total radiation efficiency while bending along the YZ plane or E-plane.

The realized gain for the flat RPA is shown to be 5.73 dBi. For small bending angles up to 60° the realized gain increases to 6 dBi. For bending angles greater than 60° the realized gain reduces to 5.04 dBi for a bending angle of 180° . The total radiation efficiency varies inversely to the realized gain as shown in Figure 5.26. The shift in total radiation efficiency from a flat RPA to an RPA bent 180° is marginal ranging from 56.4% – 58.4%.

The reflection coefficient of the RPA while bending along the antenna's XZ-plane or H-plane is shown in Figure 5.27. Bending along the H-plane of the RPA shows little changes in the resonant frequency shift when compared to E-plane bending. The resonant frequency shifts from 2.4 GHz from its unbent condition to 2.408 GHz for an 180° bend. The effective length between the radiating apertures is not changed as significantly compared to E-plane bending. It can also be shown that the magnitude of the reflection coefficient decreases as the bending angle increases due to the impedance mismatch occurring by bending across the inset feed line. Bending the inset feedline increases the input impedance from 50Ω .

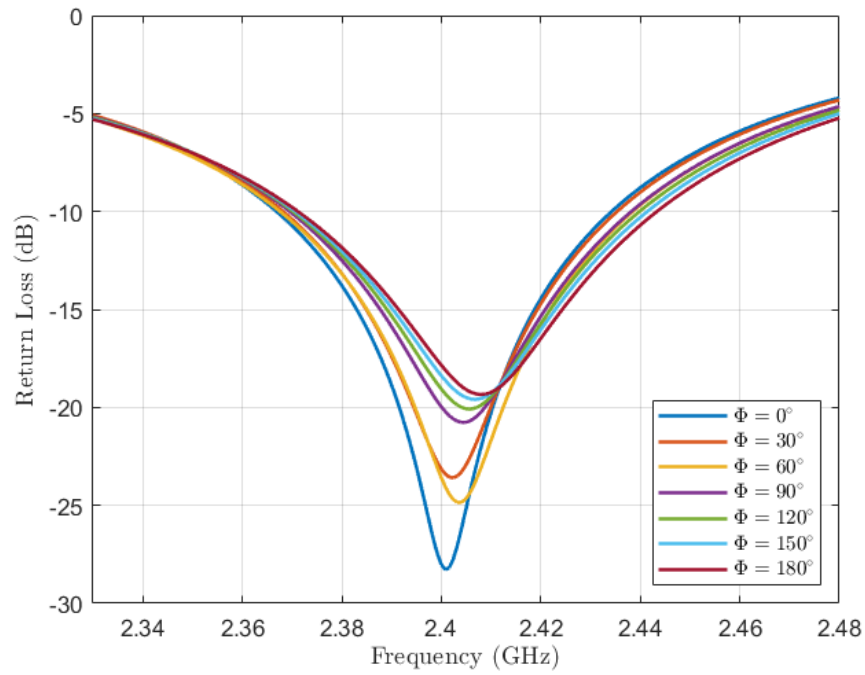


Figure 5.27: Simulated input reflection coefficient for various XZ Plane Bending angles Φ of the RPA.

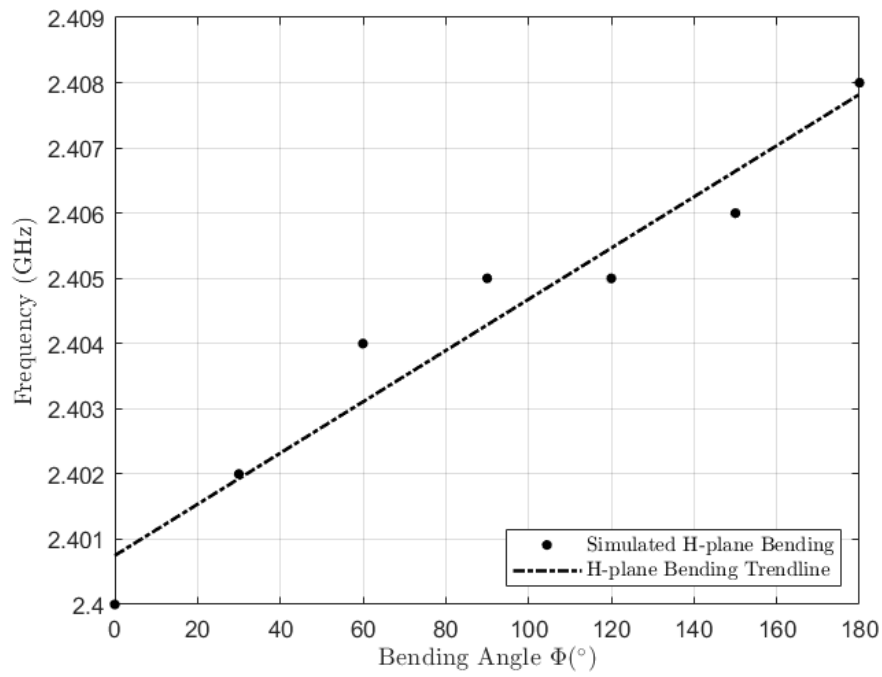


Figure 5.28: Change in resonance frequency while bending the RPA along the XZ plane or H-plane.

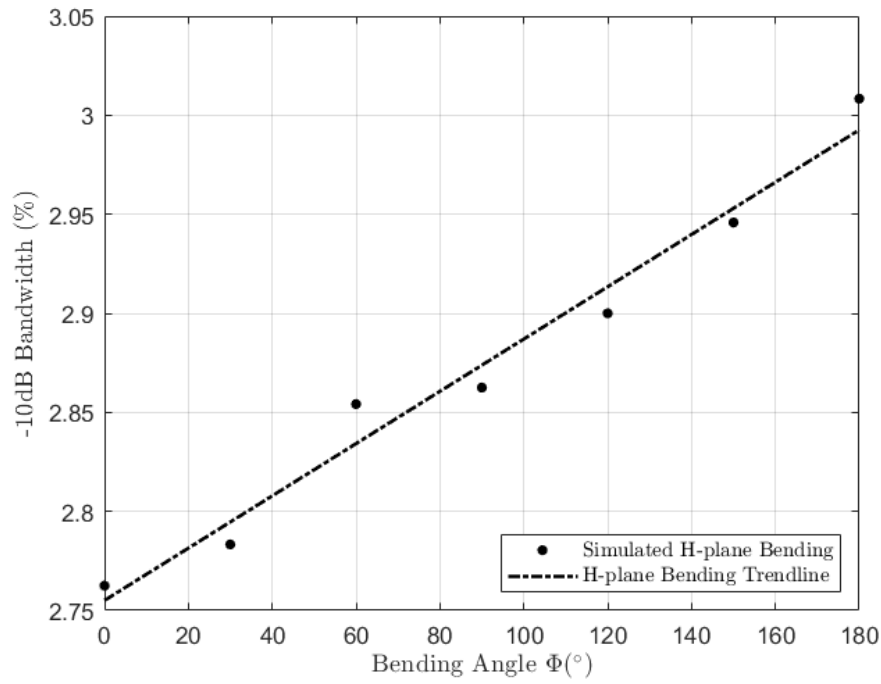
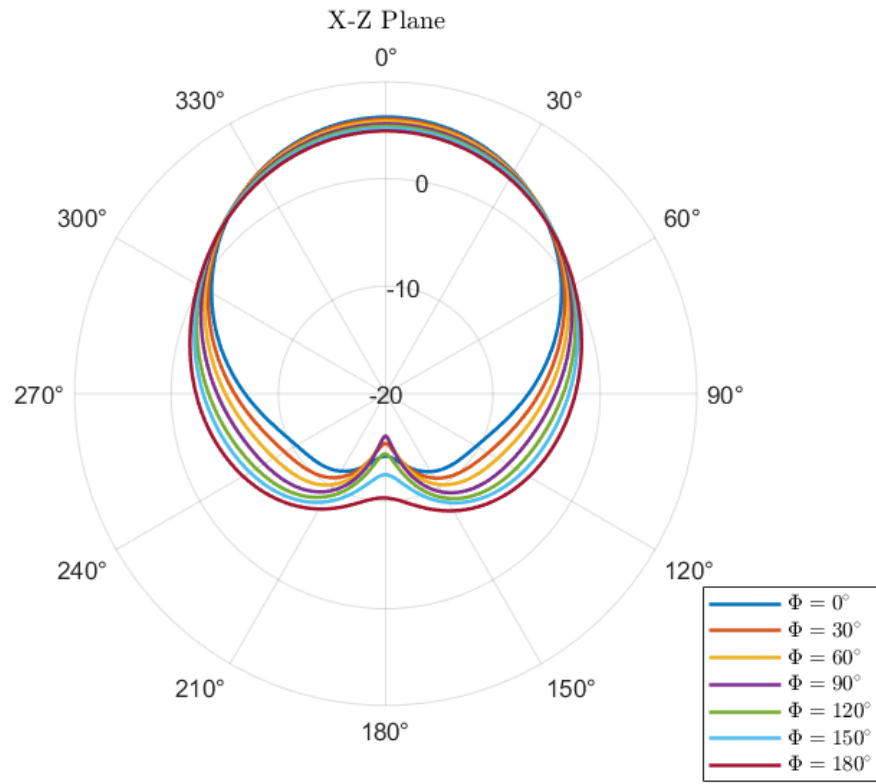


Figure 5.29: Change in bandwidth while bending the RPA along the XZ plane or H-plane.

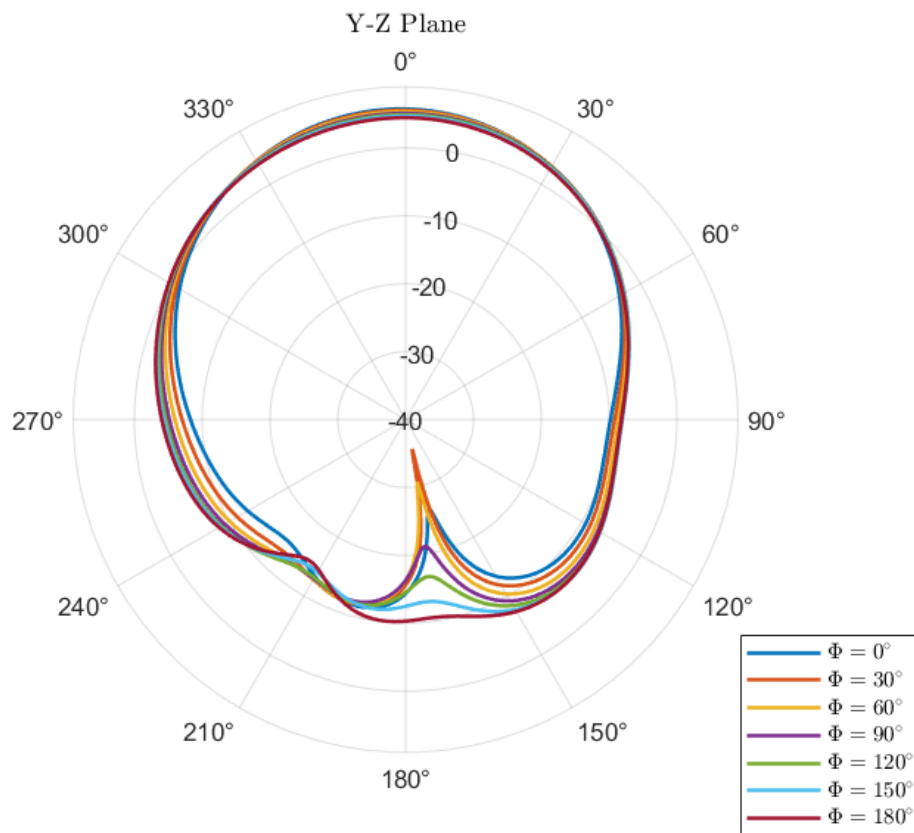
From Figure 5.29, increasing the H-plane bending angles is shown to linearly increase the RPA's bandwidth. The impedance mismatch due to bending over the inset feedline is shown to gradually increase the antenna's -10 dB bandwidth for increasing bending angles.

Bending Angle $\Phi (^{\circ})$	f_{res} (GHz)	-10 dB Bandwidth (MHz)	Total Radiation Efficiency (%)	Directivity (dBi)	Realized Gain (dBi)
0	2.40	2.76	57.7	8.14	5.73
30	2.402	2.78	59.0	7.86	5.54
60	2.404	2.85	60.5	7.53	5.34
90	2.405	2.86	61.5	7.20	5.08
120	2.405	2.90	62.7	6.91	4.88
150	2.406	2.95	63.5	6.66	4.68
180	2.408	3.01	64.2	6.36	4.43

Table 5.4: Performance parameter comparison for H-Plane bending angles.



(a)



(b)

Figure 5.30: Simulated radiation pattern in terms of realized gain for the (a) X-Z plane and (b) Y-Z plane while bending the antenna across its width or H-Plane.

The antenna's far-field is plotted in Figure 5.30 which shows a similar reduction in the realized gain at the major lobe in both planes when compared to E-plane bending. However, it can be seen in Figure 5.31 that the realized gain reduction is more significant for this type of bending, reducing to 4.43 dBi for a 180° bending angle.

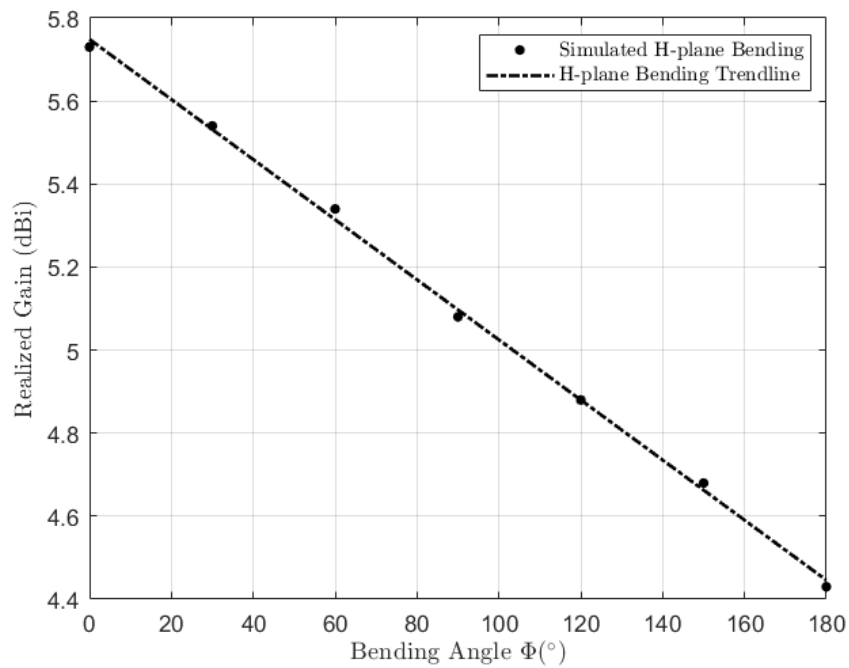


Figure 5.31: Change in realized gain while bending the RPA along the XZ plane or H-plane.

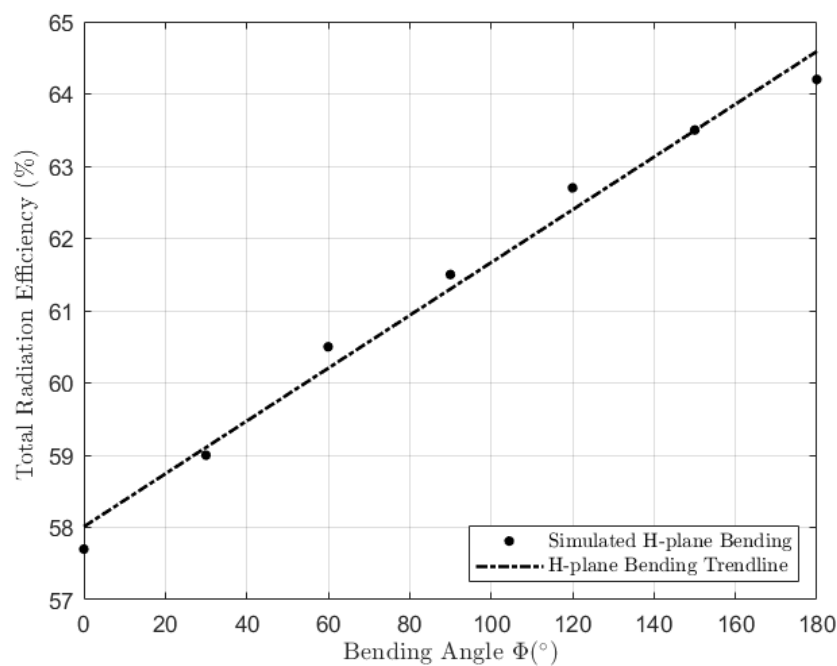


Figure 5.32: Change in total radiation efficiency while bending the RPA along the XZ plane or H-plane.

While the RPA's gain reduces for increasing bending angles, Figure 5.31 shows that the total radiation efficiency increases linearly from 57.7% to 64.2%.

The same bending analysis is performed for the Trapezoidal Patch Antenna (TPA). The bending configurations for E and H-plane bending is shown in Figure 5.33.

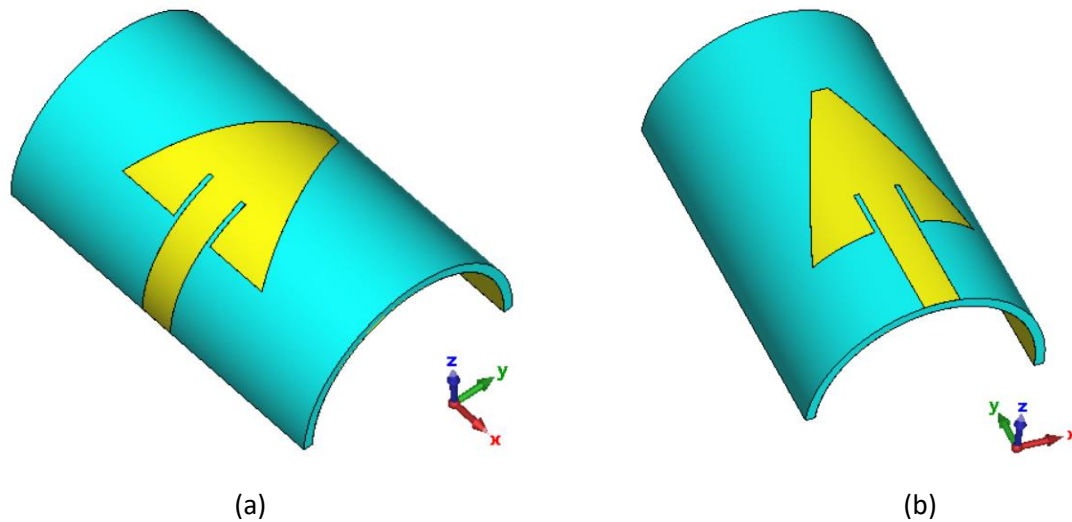


Figure 5.33: Trapezoidal Patch Antenna Cylindrically Bent Along Its (a) Width, YZ plane or E-plane and (b) Length, XZ plane or H-plane.

The TPA's reflection coefficient for bending along the E-plane is shown in Figure 5.34. The unbent TPA has a resonant frequency of 2.4 GHz which shifts to 2.43 GHz for a bending angle of 180° . Compared to the RPA for the same bending angle of 180° , the TPA resonant shift is 10 MHz less than for the RPA.

By increasing the bending angle of the antenna, a linear relationship for the increase in the TPA's resonant frequency can be established similarly to the RPA as shown in Figure 5.35. For a bending angle of 180° the magnitude of the reflection coefficient increases due to the impedance mismatch created by changing the effect length between the apertures. This results in the sudden decrease for the -10 dB bandwidth as shown in Figure 5.36.

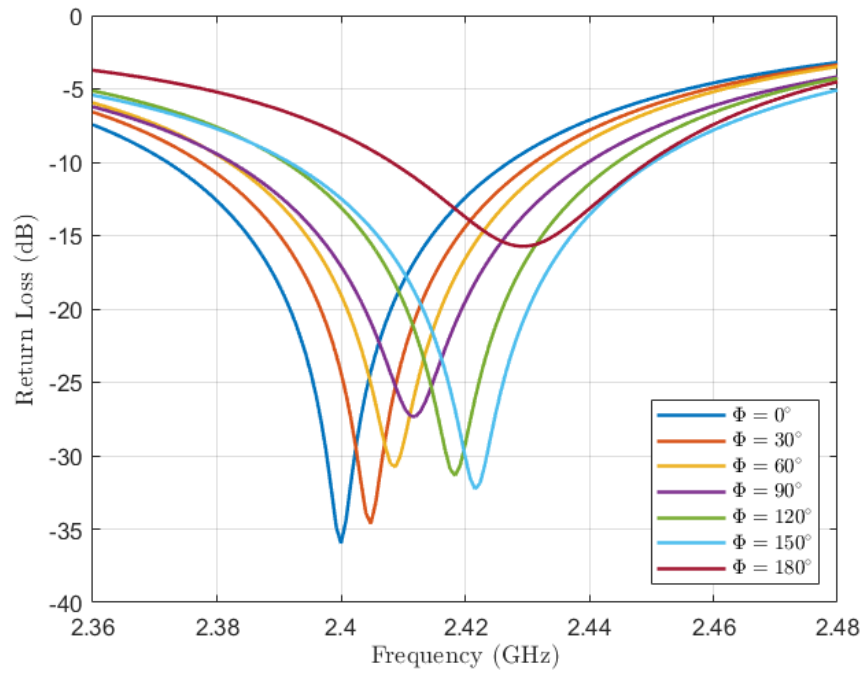


Figure 5.34: Simulated input reflection coefficient for E-plane bending of the TPA.

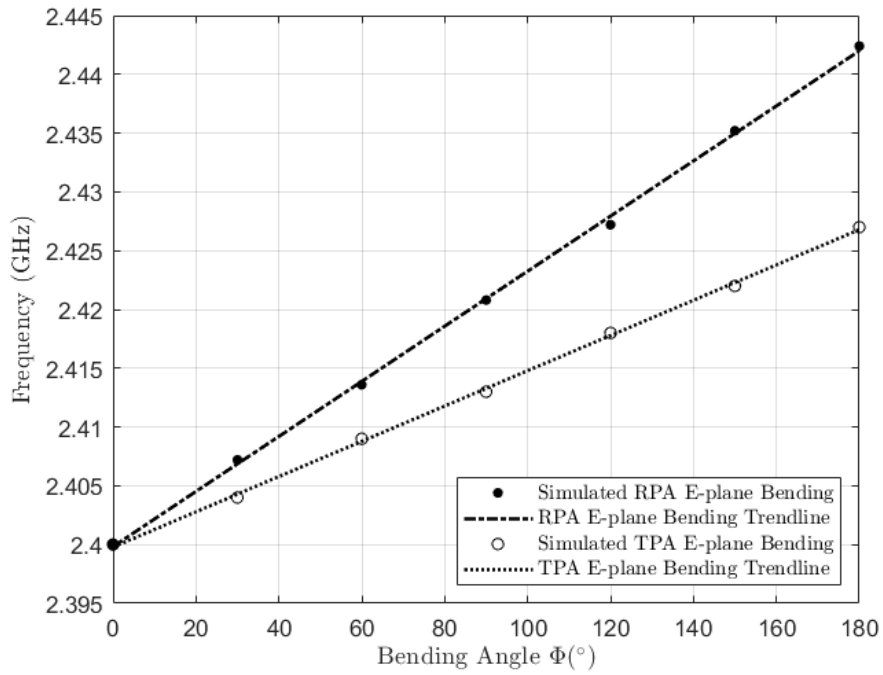


Figure 5.35: Change in the resonance frequency while bending the TPA along the YZ-plane or E-plane.

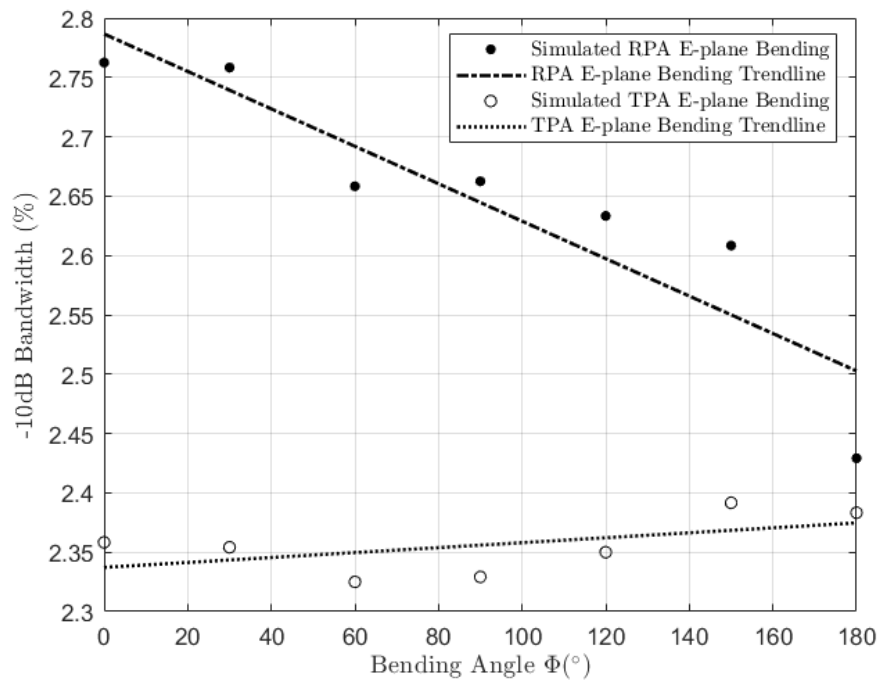
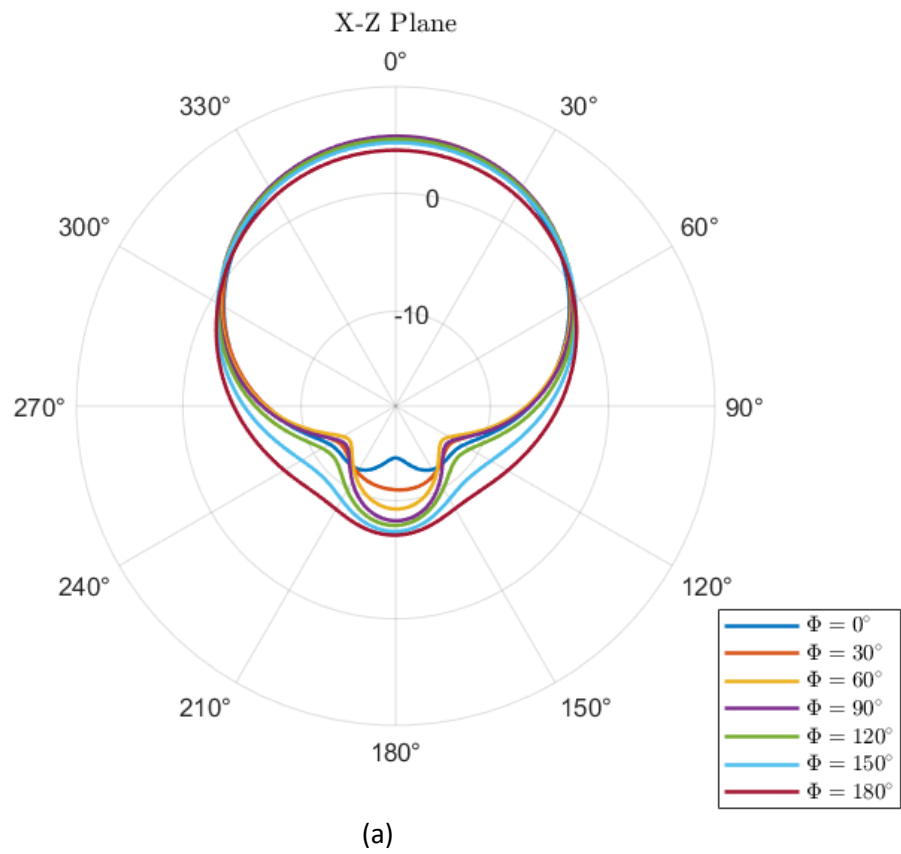


Figure 5.36: Change in the bandwidth while bending the TPA along the YZ-plane or E-plane.



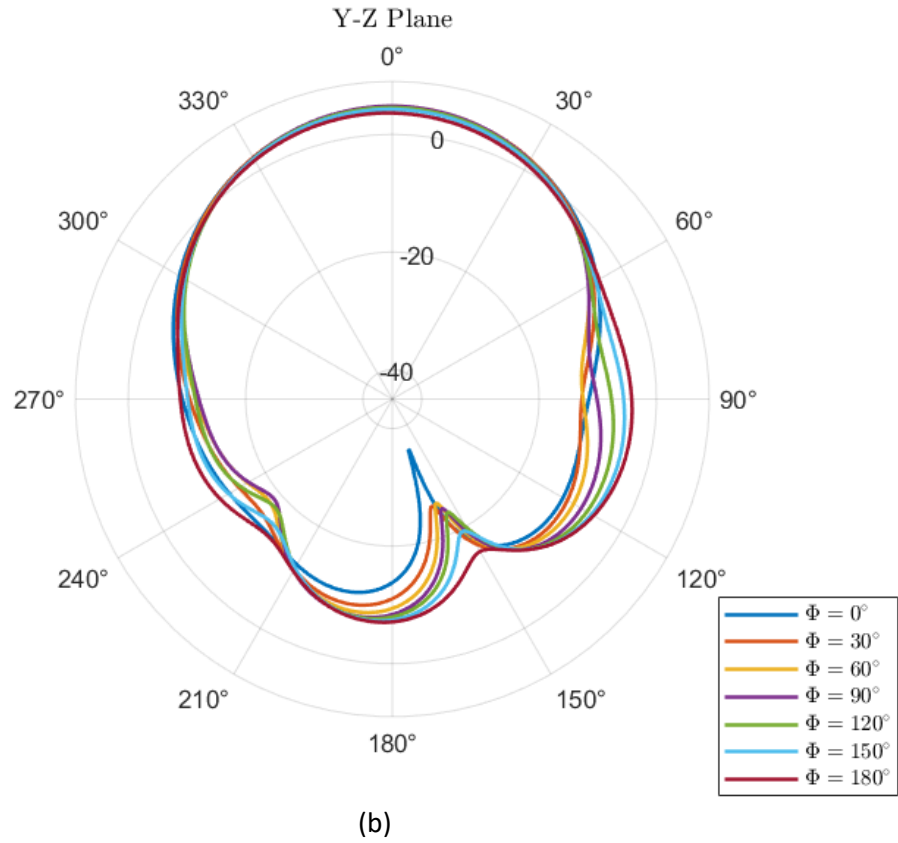


Figure 5.37: Simulated radiation pattern for (a) X-Z plane, (b) Y-Z plane and (c) X-Y plane while bending the antenna across its length or E-Plane.

Bending Angle Φ (°)	f_{res} (GHz)	-10 dB Bandwidth (MHz)	Total Radiation Efficiency (%)	Directivity (dBi)	Realized Gain (dBi)
0	2.400	2.36	47.1	7.74	4.47
30	2.404	2.35	46.3	8.01	4.67
60	2.409	2.33	46.1	8.16	4.79
90	2.413	2.33	46.4	8.08	4.75
120	2.418	2.35	47.0	7.81	4.54
150	2.422	2.39	48.4	7.38	4.23
180	2.427	2.38	49.7	6.79	3.76

Table 5.5: Performance parameters comparison for E-plane bending angles.

The far-field is plotted in terms of realized gain in Figure 5.37 for bending the TPA along its E-plane. The far-field behaviour shows the same behaviour as bending the RPA, whereby for increasing bending angles the main lobe reduces in magnitude, while the back lobe and side lobes increase in magnitude.

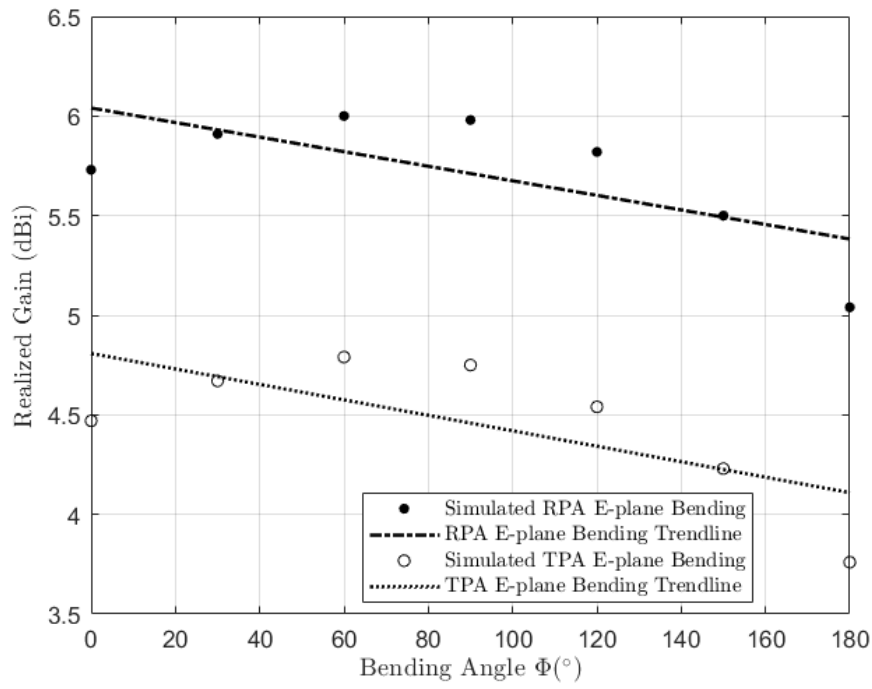


Figure 5.38: Change in realized gain while bending the TPA along the YZ plane or E-plane.

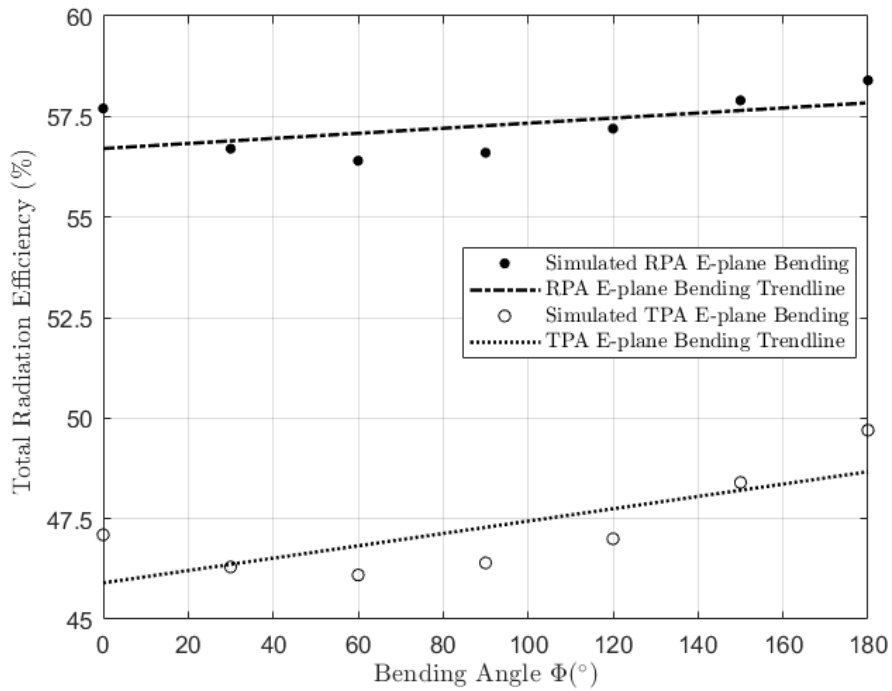


Figure 5.39: Change in radiation efficiency while bending the TPA along the YZ plane or E-plane.

Figure 5.38 shows that the TPA has a smaller realized gain due to the size reduction of a radiating aperture when compared to the RPA. The realized gain for the unbent TPA is 4.47 dBi which increases to 4.79 dBi for a 60° bending angle then decreases to 3.76 dBi for a bending angle of 180°.

Finally, the bending analysis is performed on the TPA for H-plane bending. The simulated reflection coefficient results in Figure 5.40 show that bending has a small impact on the resonant frequency shift, but rather a significant impact on the magnitude of the return loss. Just like for bending along the H-plane of the RPA, the gradual increase in the magnitude of the TPA's return loss for increasing bending angles is due to the impedance mismatch created by the bending across the inset feedline. The trends for the antenna's resonant frequency shift and bandwidth are shown in Figure 5.41 and Figure 5.42.

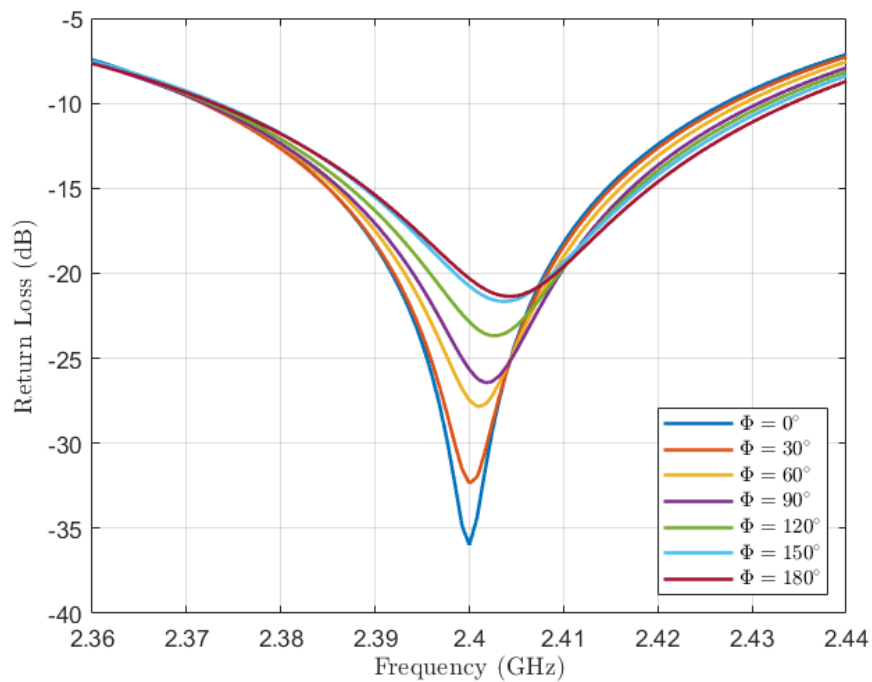


Figure 5.40: Simulated input reflection coefficient for H-plane bending of the TPA.

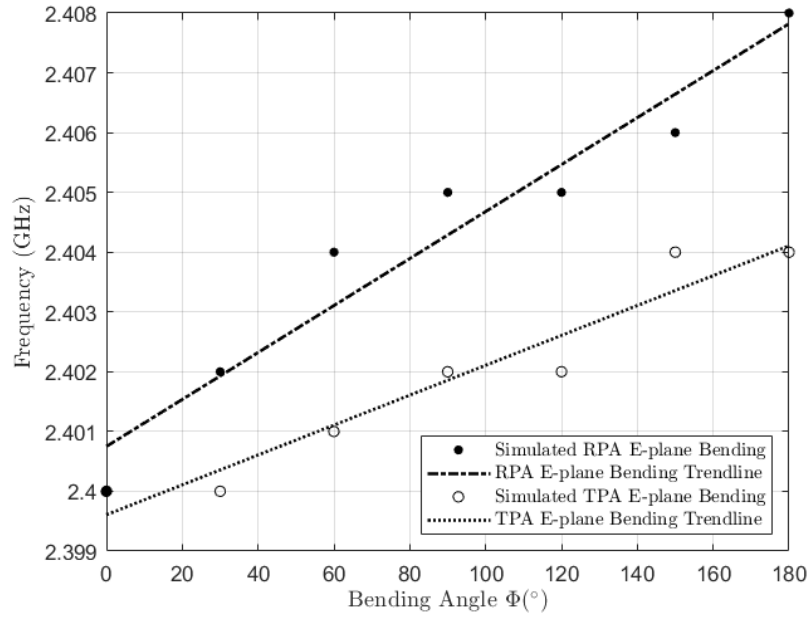


Figure 5.41: Change in resonance frequency of the TPA while bending along the XZ plane or H-plane.

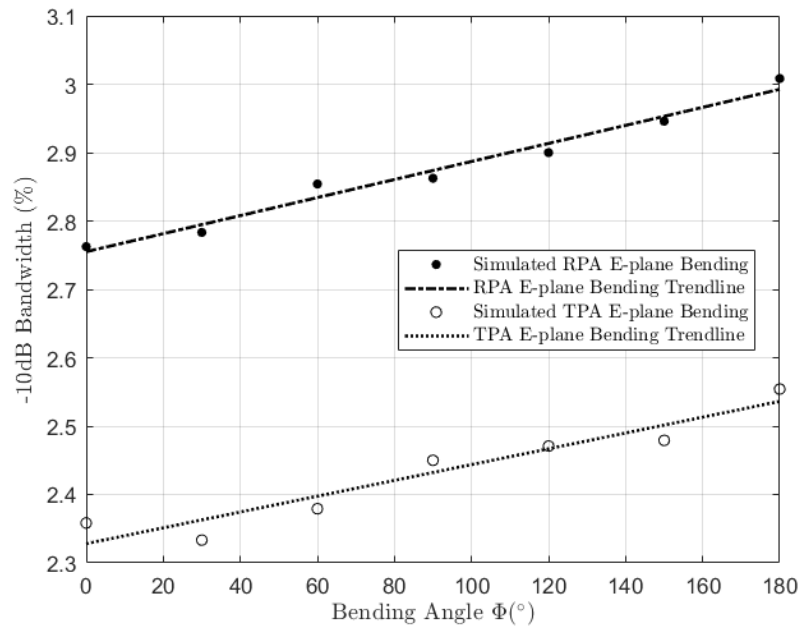
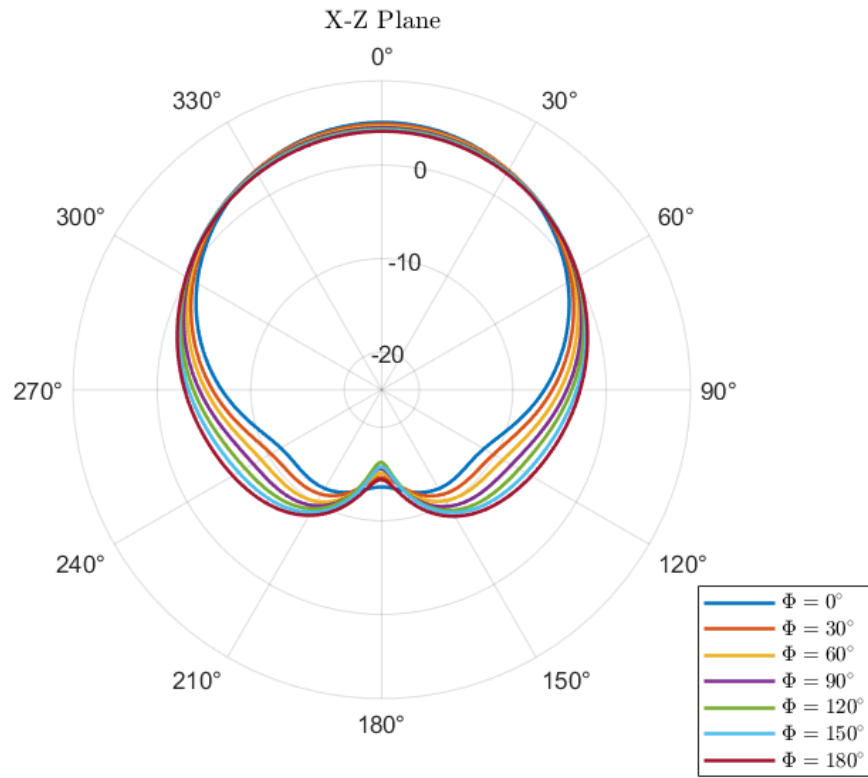
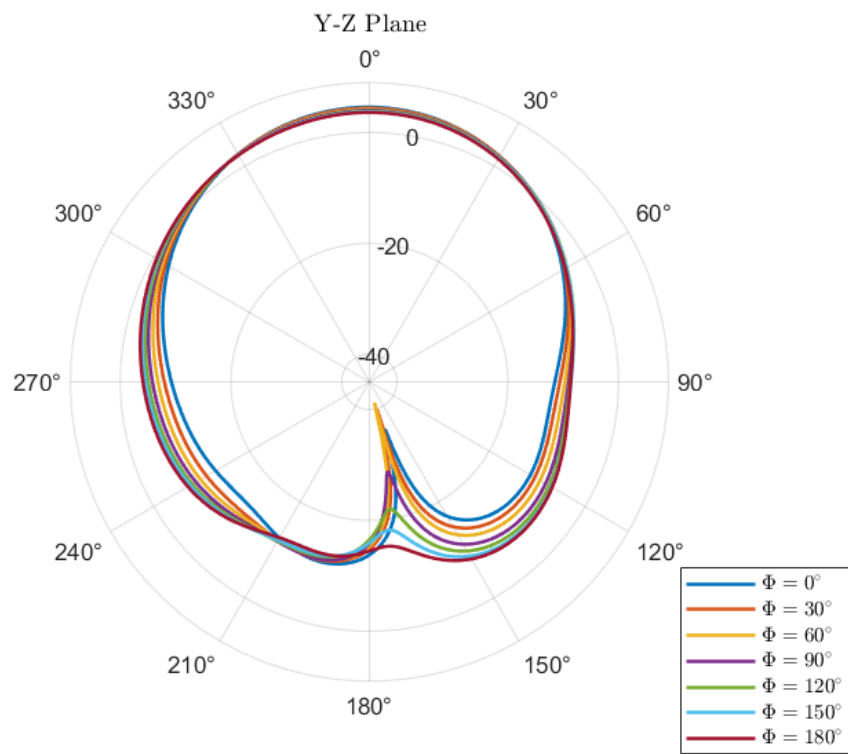


Figure 5.42: Change in the bandwidth of the TPA while bending along the XZ plane or H-plane.



(a)



(b)

Figure 5.43: Simulated radiation pattern for (a) X-Z plane and (b) Y-Z plane while bending the antenna across its width or H-Plane.

Bending Angle Φ ($^{\circ}$)	f_{res} (GHz)	–10 dB Bandwidth (MHz)	Total Radiation Efficiency (%)	Directivity (dBi)	Realized Gain (dBi)
0	2.400	2.36	47.1	7.74	4.47
30	2.400	2.33	49.3	7.36	4.29
60	2.401	2.38	51.2	6.98	4.07
90	2.402	2.45	53.3	6.63	3.90
120	2.402	2.47	55.0	6.35	3.75
150	2.404	2.48	56.0	6.07	3.55
180	2.404	2.55	58.0	5.86	3.49

Table 5.6: Performance parameters comparison of the TPA for H-plane bending angles.

The far-field behaviour for TPA H-plane bending in Figure 5.43 shows a reduction in the antenna's realized gain for increasing bending angles at the main lobe. Also, it can be shown that the side lobe broadens, and the back lobe increases in magnitude for increasing bending angles for both plane cuts.

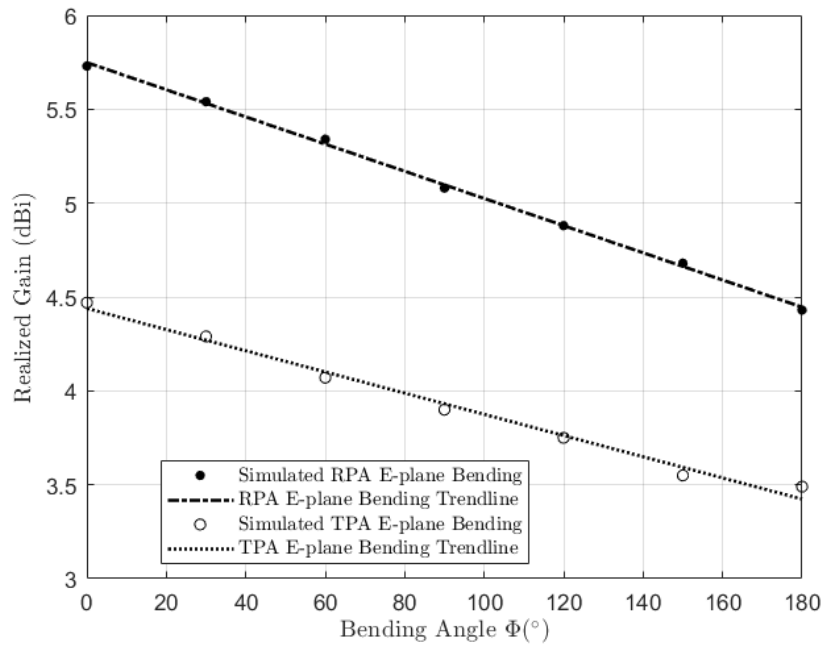


Figure 5.44: Change in realized gain while bending the TPA along the XZ-plane or H-plane.

The change in the TPA's gain and total radiation efficiency for increasing bending angles is plotted in Figure 5.44 and Figure 5.45. The realized gain reduces linearly for increasing bending angles while the total radiation efficiency increases.

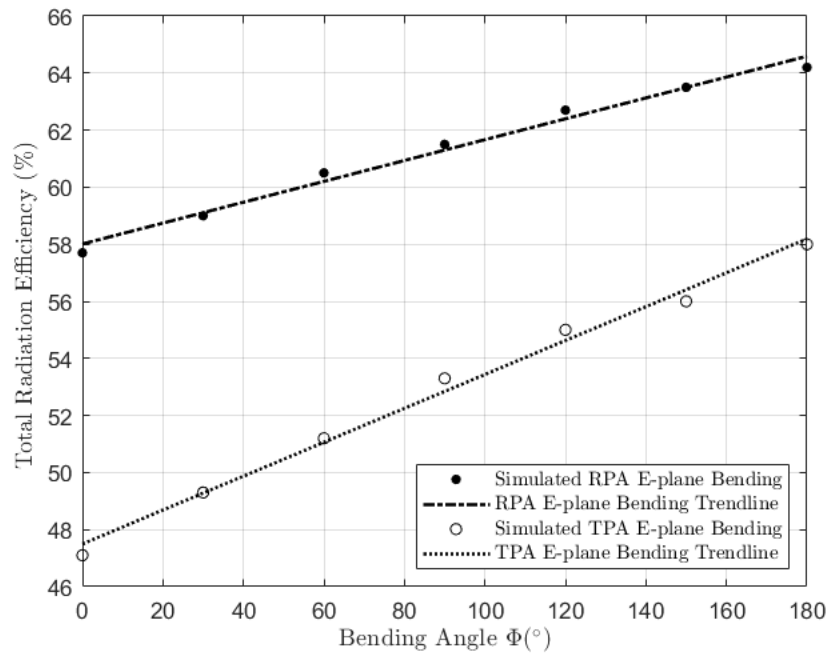


Figure 5.45: Change in radiation efficiency while bending the TPA along the XZ-plane or H-plane.

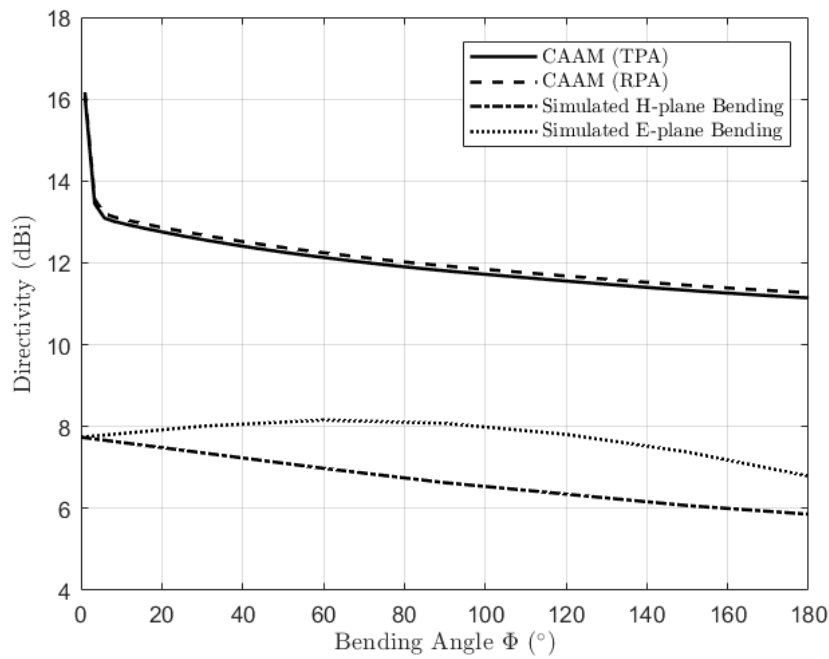


Figure 5.46: Comparison between the analytical and simulated directivity of the TPA during bending.

Lastly, a comparison is made between the computed directivity from the Cylindrical Aperture Array Model (CAAM) and the full wave simulations for bending across both the E and H-plane of the TPA. The CAAM and the simulated H-plane bending analysis have a common trend of decreasing directivity for increasing bending angles. The behaviour for the

simulated E-plane bending analysis is quite different, with an initial increase in directivity peaking for a bending angle of 60° and then decreasing. This type of bending is also where the biggest resonant detuning occurs for increasing bending angles.

The CAAM attempts to model E-plane bending whereby the radiating apertures are assumed to be uniform and rectangular for a given bending angle. As shown in Figure 5.46, while the CAAM predicts a much higher directivity, the result for directivity over various bending angles resembles the same trend as for the simulated H-plane bending analysis. The simulated analysis for E-plane bending shows more complex behaviour whereby the directivity increases for bending angles $\Phi < 60^\circ$ and then starts to deteriorate at higher bending angles.

While the CAAM provides insight into the far-field behaviour from an analytical perspective by computing the normalized power plot, it fails to accurately represent the directivities for E-plane bending when compared to the full-wave simulation results.

Overall, the CAAM predicts a much higher directivity when compared to the full wave simulations for E-plane bending. Previously this was explained due to the entire cylinder being conductive rather than being only conductive to where the ground plane extends to, therefore the directivity is enhanced. Also, the directivity in the CAAM might be overshooting due to not implementing the effects of the dielectric substrate. Currently in this analysis the dielectric properties are only implemented in calculating the length L between the apertures of the antenna as shown in Chapter 4. It may be that modifying the CAAM by adding a dielectric coating may lead to a better result. However, this adds further complexity to the model.

To conclude, the full wave simulation performance of the RPA and TPA for both types of bending exhibit similar properties. Bending the TPA along both the E and H-plane results in marginal frequency detuning. Bending along the E-plane results in a frequency detune which is 15 MHz less than the detuning for the RPA, while H-plane bending is 4 MHz less. The realized gain for the TPA suffers a lot more than the RPA. It was already shown in Chapter 4 that miniaturising the RPA to a TPA results in a reduction in the antenna's realized gain. Bending along the antenna's H and E-plane worsens the performance further. Some preliminary measurements for E-plane bending the RPA and TPA is reported in the appendix.

5.5 References

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6 Optimisation Techniques using an Artificial Magnetic Conductor (AMC) Metafilm

This chapter introduces the Artificial Magnetic Conductor (AMC) and the physics behind its zero-degree phase reflection properties. The application of the AMC can enhance the radiation performance of a patch antenna if loaded into the dielectric substrate in an array configuration. The specific array configuration is known as a metafilm. A characterization of the dielectric substrate with the AMC metafilm is performed through full wave simulation using the effective medium model and generalized sheet transition condition (GSTC) method. A full wave simulation analysis is performed on the TPA while embedding the AMC metafilm into the substrate. The Trapezoidal Patch Antenna loaded with an AMC MetaFilm (TPA-MF) displays enhanced radiation characteristics when compared to a Trapezoidal Patch Antenna (TPA). A bending analysis is performed on the antenna to investigate how it affects the antenna's resonant frequency and radiation characteristics.

6.1 Background of an Artificial Magnetic Conductor Metafilm

The ground plane on a microstrip patch antenna prevents radiation from propagating backwards by redirecting it in the opposite direction. The Interaction between the backpropagating waves from the patch radiator and the forward reflected waves from the ground plane are out of phase and form a standing wave. If the patch antenna consists of a very thin substrate which is less than $\lambda/4$, the phase of the backpropagating wave is reversed when reflected from the ground plane. This results in destructive interference with the waves that propagate outwards into free space from the top side of the radiating element. This is shown in Figure 6.1(a).

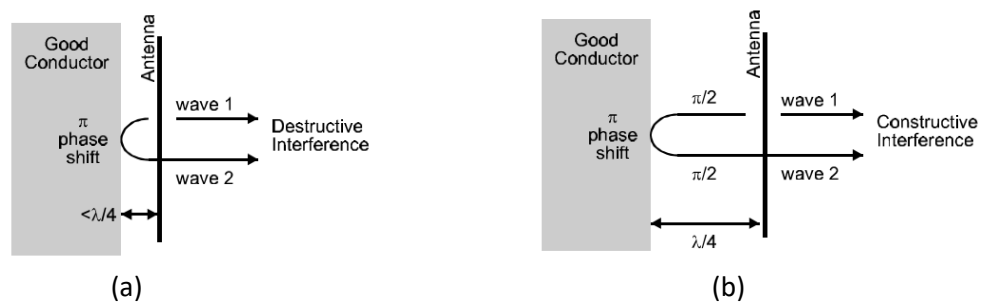


Figure 6.1: Radiating element close placed at a distance (a) less than $\lambda/4$, (b) approximate $\lambda/4$ from ground plane [1].

If the thickness of the substrate is approximately $\lambda/4$, as shown in Figure 6.1(b), the phase shift is cancelled leading to constructive interference. The main issue with this configuration is that it creates an extremely thick antenna if it were designed for operation at 2.4 GHz.

In 1999, Sivenpiper [2] found that the surface impedance of the ground can be altered by introducing an array of metallized patches above the ground plane. The introduced patches increase the surface impedance giving rise to the incident electric field having an antinode at the surface while the magnetic field has a node.

To understand this concept better a standing wave representation of the electric and magnetic as presented by Sivenpiper is shown in Figure 6.2.

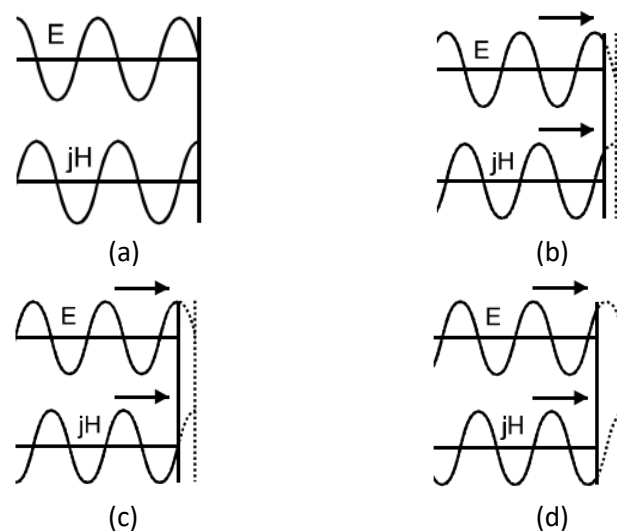


Figure 6.2: Standing wave fields at a frequency (a) far below resonance, (b) Near or below the resonance, (c) at the resonance and (d) above the resonance [1].

Away from the intended operational frequency of the structure, the surface impedance is low. The electric field has a node at the surface, and the magnetic field has an antinode, as shown in Figure 6.2(a). The behaviour away from the resonant frequency is the same to that of a ground plane or a metal sheet in which there is a phase reflection reversal. Near the resonant frequency or just below it, the formed standing wave behaves as if its reflection point has shifted which is represented by the dotted line in Figure 6.2. At resonance the surface is largely inductive, and the impedance is large which is represented by the antinode for the electric field at the surface in Figure 6.2(c) where it has now shifted enough where the reflection phase crosses through zero [1]. Above the resonant frequency the surface is now capacitive and the magnetic field is negative as shown in Figure 6.2(d), which behaves as if

the reflection phase has slipped into the surface by one-half wavelength as shown by the imaginary dotted line boundary [1].

Another issue with the presence of the ground plane is due to the generated surface currents that propagate along the ground plane. Due to the finite size of the ground, when these currents reach the edge, they radiate into free space. The fringe fields produced by each of the apertures at the radiating element have a resultant effect at the far-field which produces a space wave propagating normally from the patch. Spurious fringe fields propagate towards the edge of the ground plane and results in multipath interfere with the surface waves as shown in Figure 6.3.

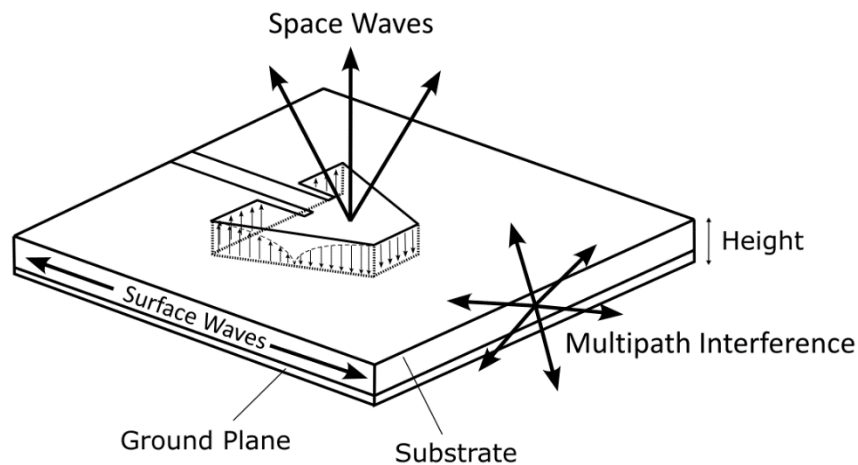


Figure 6.3: Interference of the space waves and surface waves at the edge of the ground plane.

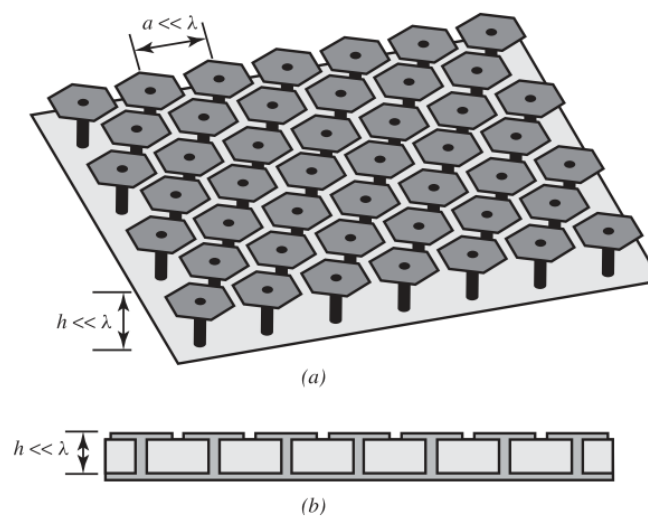


Figure 6.4: Hexagonal high impedance surface (a) perspective view and (b) cross sectional side view [3].

Sievenpiper [1] proposed a High Impedance Surface (HIS) ground plane that has an in-phase reflection characteristics and does not support surface waves. The proposed high impedance surface consists of an array of metallized patches of hexagonal geometry connected to the ground plane by a vertical shorting pin as shown in Figure 6.4. The addition of the vertical shorting pin makes sure that surface waves are not propagating below the metallized patch array. The vertical electric field current density that propagates through the shorting pin can be represented using the circuit as shown in Figure 6.5(b) having the impedance of a parallel resonant LC circuit. Where the capacitance represents the gaps between the hexagonal array elements and the total inductance is represented by the conductive elements from the hexagonal patch, shorting pin, and ground plane. The circuit is capacitive above the resonant frequency and thus the structure does not support the propagation of TM surface waves [2].

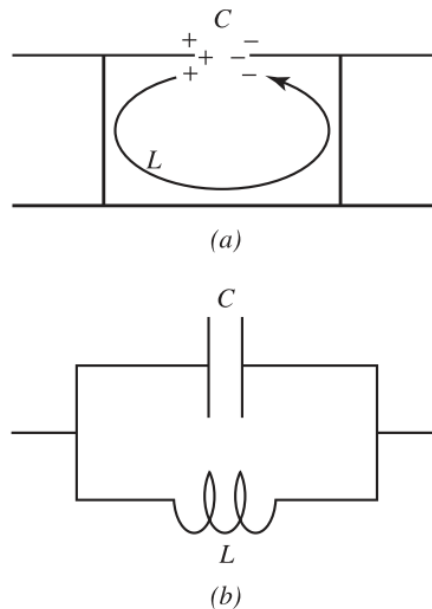


Figure 6.5: Unit cell of the hexagonal high impedance surface (a) cross-sectional view and equivalent circuit [3].

Sievenpiper compared the E-plane radiation performance of a centre fed patch antenna operating at 13.5 GHz, with the proposed high-impedance surface ground plane to that of a regular ground plane. The performance of the antenna with the regular ground plane showed to have significant back propagating radiation with ripples produced in the forward propagating main lobe.

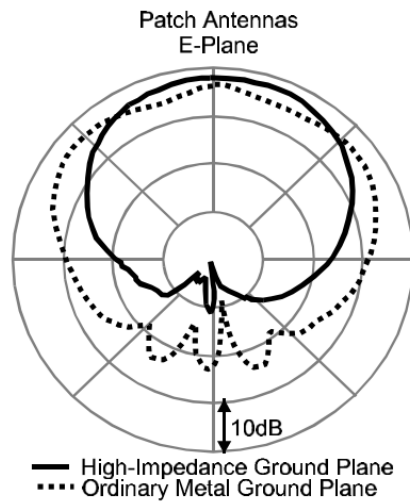


Figure 6.6: E-plane radiation pattern of a patch antenna consisting of an ordinary ground plane and a high-impedance surface as a ground plane [1].

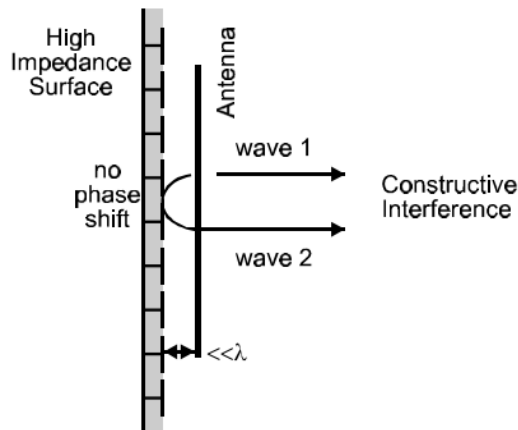
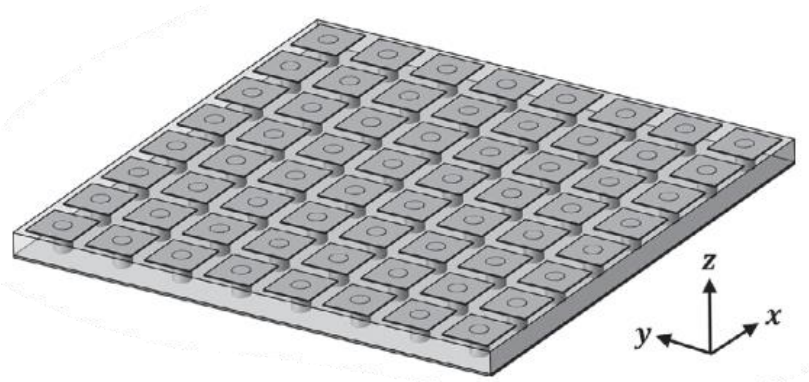
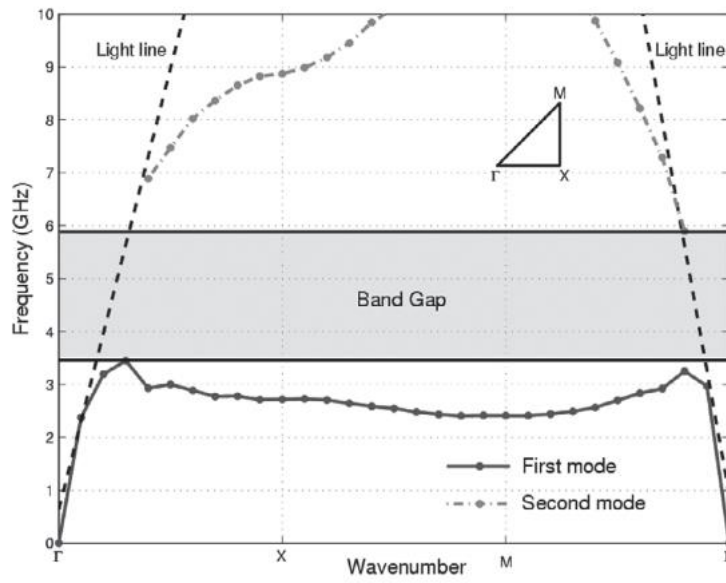


Figure 6.7: Dipole antenna mounted on a high impedance ground plane [1].

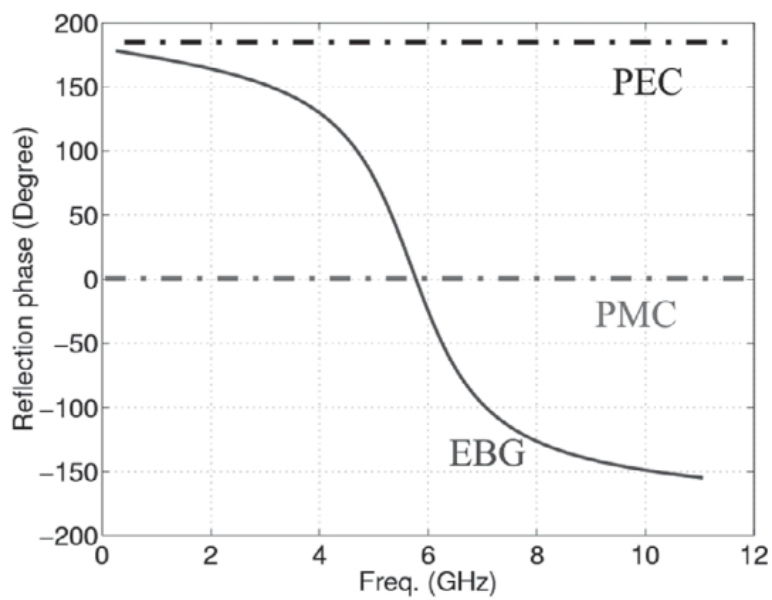
Figure 6.7 shows a dipole antenna mounted next to a high-impedance surface (HIS). The HIS allows the antenna to be near the ground plane without sacrificing performance from out of phase wave reflections and propagating surface waves. The high-impedance surface is also known as an electromagnetic band gap (EBG) surface [4] due to the region or band-gap produced in the dispersion plot for surface wave propagation. The EBG structure shown in Figure 6.8(a) was simulated to analyse the dispersion and reflection phase properties. Figure 6.8(b) shows distinct band where no surface waves can propagate regardless of its direction. The reflection phase properties of the EBG structure are shown in Figure 6.8(c) as well as the phase regions where the structure exhibits properties as a perfect electric conductor (PEC) and perfect magnetic conductor (PMC).



(a)



(b)



(c)

Figure 6.8: The (a) geometry of a high-impedance or electromagnetic band gap surface. The simulated (b) dispersion properties showing the band gap for surface waves and (c) reflection phase properties [4].

A high-impedance or EBG surface without the vertical vias is a surface known as an artificial magnetic conductor (AMC) [4]. The difference between a high-impedance or EBG surface and an artificial magnetic conductor (AMC) surface is that the AMC surface only allows the control of the phase reflection properties since there is no vertical vias to prevent surface mode propagation. The AMC structure can be thought of as a surface distribution of electrically small scatterers just like that of a frequency-selective surface (FSS). The term ‘metafilm’ is used to describe these structures as a two-dimensional pattern at a surface characterized by its electric and magnetic surface susceptibilities [5].

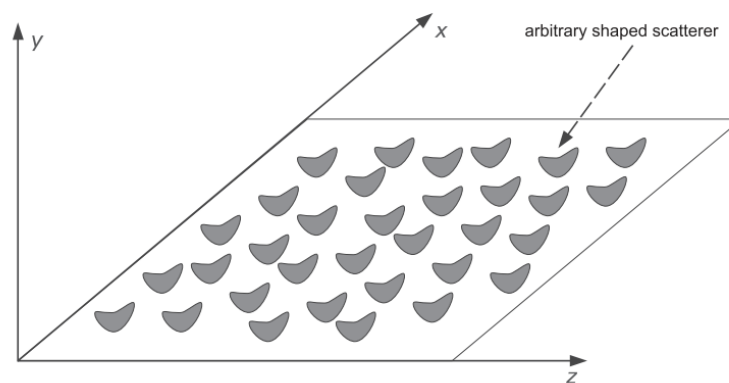


Figure 6.9: Illustration of a metafilm with a periodic arrangement of an arbitrarily shaped scatterer in the xy -plane [4].

A fundamental theorem of surface electromagnetics is the Generalized Sheet Transition Conditions (GSTC), which is argued as the most suitable way of characterizing a metafilm embedded on or in a dielectric substrate [6]–[9]. Surface electromagnetics focuses on the study of the boundary conditions where the surface equivalence theorem is the foundation. The surface equivalence theorem links the internal fields to the external fields to that of a closed surface by constructing surface currents on the boundary of the surface that satisfy the defined boundary conditions [4].

The generalized sheet transition conditions (GSTCs) are a set of equations that have coefficients that model the metafilms interaction with an electromagnetic field. These coefficients correspond to the uniquely defined electric χ_{ES} and magnetic χ_{MS} surface susceptibilities which have units of meters as they are dependent on the sheet thickness [10]. The surface susceptibilities describe the shape and size characteristics of the metafilm which are incorporated through the boundary conditions in equation (6.1) and (6.2) [11].

The GSTC equations for a metafilm are given by

$$\mathbf{a}_y \times [\mathbf{E}^A - \mathbf{E}^B]_{y=0} = -j\omega\mu_0(\boldsymbol{\chi}_{MS} \cdot \mathbf{H}_{av})_t - \mathbf{a}_y \times \nabla_t(\mathbf{a}_y \cdot \boldsymbol{\chi}_{ES} \cdot \mathbf{E}_{av}) \quad (6.1)$$

$$\mathbf{a}_y \times [\mathbf{H}^A - \mathbf{H}^B]_{y=0} = -j\omega\epsilon_0(\boldsymbol{\chi}_{ES} \cdot \mathbf{E}_{av})_t - \mathbf{a}_y \times \nabla_t(\mathbf{a}_y \cdot \boldsymbol{\chi}_{MS} \cdot \mathbf{H}_{av}) \quad (6.2)$$

Equations (6.1) and (6.2) represent the difference in the tangential components of the electric and magnetic fields across the metafilm [5] for an incident field as shown in Figure 6.10(b). The electric ($\mathbf{E}^A$ and $\mathbf{E}^B$) and magnetic fields ($\mathbf{H}^A$ and $\mathbf{H}^B$) above and below the reference plane are defined through the superscripts A and B respectively.

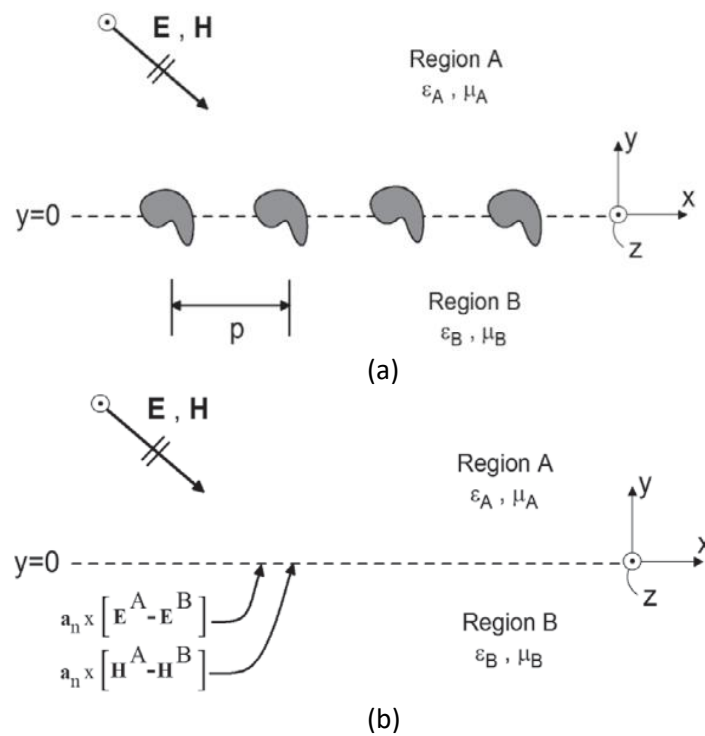


Figure 6.10: Metafilm with (a) a periodic arrangement of an arbitrarily shaped scatterer and the (b) equivalent model for a metafilm where the GSTCs are applied [4].

The subscript t refers to the x and z components which are tangential to the y axis, and the average fields are defined by

$$\mathbf{E}_{av} = \frac{1}{2}(\mathbf{E}^A + \mathbf{E}^B) \quad (6.3)$$

and

$$\mathbf{H}_{av} = \frac{1}{2}(\mathbf{H}^A + \mathbf{H}^B) \quad (6.4)$$

The actual and equivalent model for the metafilm can be shown in Figure 6.10(b) where the reference plane for the metafilm is centred at $y = 0$. It is also important to note that the fields

described by the GSTC equations are macroscopic, in that they exhibit variations only on a large scale such as the wavelength in the surrounding medium [11].

There are approximate analytical expressions based on dipole interactions for a square patch array to obtain the surface susceptibility of a metafilm [12]. For other metafilm geometries the surface susceptibilities can be determined using a full wave analysis or from experimental measurements. This is achieved by determining the reflection and transmission coefficients for an incident TE and TM polarized wave on the metafilm. For metafilm structures in which the scatterers have sufficient symmetry the surface susceptibility dyadic is diagonal. Therefore only χ^{xx} , χ^{yy} and χ^{zz} exist within the surface susceptibility dyadic. These are computed using the following equations [11].

For an incident TM polarized wave at normal incidence χ_{ES}^{xx} , χ_{MS}^{zz} and χ_{ES}^{yy} is determined by

$$\chi_{ES}^{xx} = \frac{2j R_{TM}(0) + T_{TM}(0) - 1}{k_0 R_{TM}(0) + T_{TM}(0) + 1} \quad (6.5)$$

$$\chi_{MS}^{zz} = \frac{2j R_{TM}(0) - T_{TM}(0) + 1}{k_0 R_{TM}(0) - T_{TM}(0) - 1} \quad (6.6)$$

$$\chi_{ES}^{yy} = -\frac{\chi_{MS}^{yy}}{\sin^2(\theta)} + \frac{2j \cos(\theta) T_{TM}(\theta) - 1 - R_{TM}(\theta)}{k_0 \sin^2(\theta) T_{TM}(\theta) + 1 - R_{TM}(\theta)} \quad (6.7)$$

and for an incident TE polarized wave at normal incidence χ_{ES}^{xx} , χ_{ES}^{zz} and χ_{MS}^{yy} is determined by

$$\chi_{MS}^{xx} = \frac{2j R_{TE}(0) - T_{TE}(0) + 1}{k_0 R_{TE}(0) - T_{TE}(0) - 1} \quad (6.8)$$

$$\chi_{ES}^{zz} = \frac{2j R_{TE}(0) + T_{TE}(0) - 1}{k_0 R_{TE}(0) + T_{TE}(0) + 1} \quad (6.9)$$

$$\chi_{MS}^{yy} = -\frac{\chi_{MS}^{yy}}{\sin^2(\theta)} + \frac{2j \cos(\theta) T_{TE}(\theta) - 1 - R_{TE}(\theta)}{k_0 \sin^2(\theta) T_{TE}(\theta) + 1 - R_{TE}(\theta)} \quad (6.10)$$

where $R(0)$ is the reflection coefficient at normal incidence, $T(0)$ is the transmission coefficient at normal incidence, the subscripts denote whether it is for a TE or TM polarized

wave. χ_{ES}^{yy} and χ_{MS}^{yy} is calculated using the transmittance $T(\theta)$ and reflectance $R(\theta)$ at some oblique incidence angle θ [10].

To summarize so far, a metafilms interaction with a field can be interpreted macroscopically using the GSTC equations. Also, characterizing a metafilm by its surface susceptibility has been shown to be an effective method of characterizing its properties over the traditional effective-medium layer model [8].

Through the GSTCs the derived reflection and transmission coefficients for the metafilm predict a resonator that overcomes the $\lambda/2$ size limit when placed between two parallel metal plates [12].

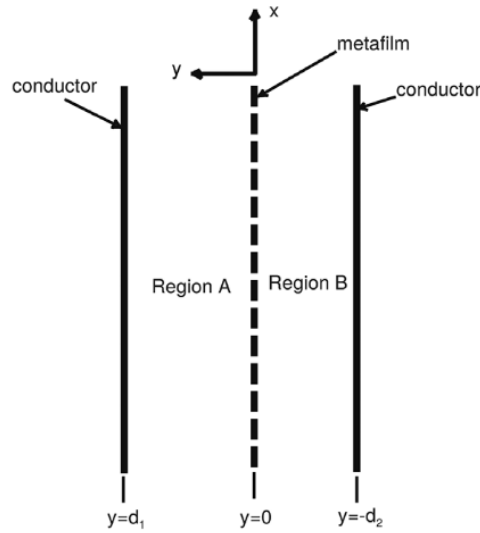


Figure 6.11: Metafilm placed between two parallel metal plates [4].

Figure 6.11 shows the metafilm placed at $y = 0$ modelled between two parallel plates located at $y = d_1$ and $y = -d_2$. For simplicity, region A and B are filled with the same homogenous medium. For a metafilm placed at the centre between the parallel plates the phase-matching conditions at resonance for a given separation distance $d_1 = d_2 = d$ can be expressed as

$$d = \frac{\lambda}{2} \left[1 + n\pi - \frac{2}{\pi} \tan^{-1} \left(\frac{\pi}{\lambda} \chi_{ES} \right) \right] \text{ for } n = 0, 1, 2, 3, \dots \quad (6.11)$$

$$d = \frac{\lambda}{2} \left[2n\pi - \frac{2}{\pi} \tan^{-1} \left(\frac{\pi}{\lambda} \chi_{MS} \right) \right] \text{ for } n = 0, 1, 2, 3, \dots \quad (6.12)$$

The equations show that designing a metafilm that exhibits a certain magnetic or electric susceptibility which makes it is possible to overcome the $\lambda/2$ size limit for the resonator or radiating element. The derivations for this result can be found in [12].

An example for a square patch array is shown in Figure 6.12.

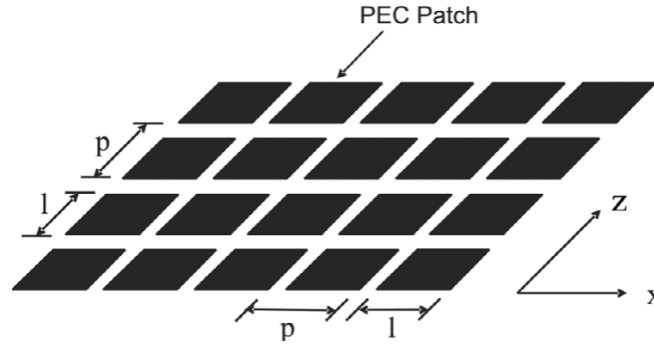


Figure 6.12: Surface array of metallic square patches with a period 'p' and a patch width 'l' [12].

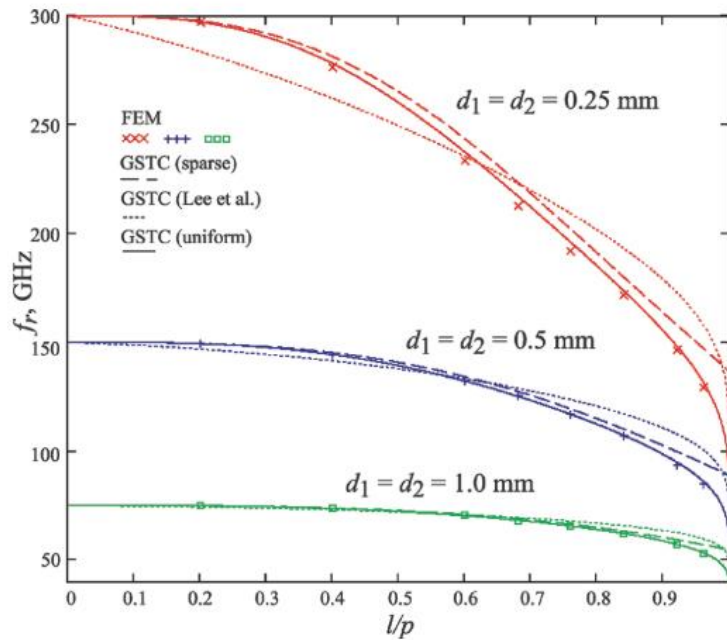


Figure 6.13: Resonant frequency of an air-filled parallel-plate resonator loaded with a square-patch metafilm [10].

The resonant frequencies for the parallel-plate resonator with a loaded square-patch metafilm are shown in Figure 6.13. Full wave result and various analytical results are compared with each other. The results show that for high values of l/p and small separation gaps d , a high degree of miniaturization can be achieved. The degree at miniaturization occurs

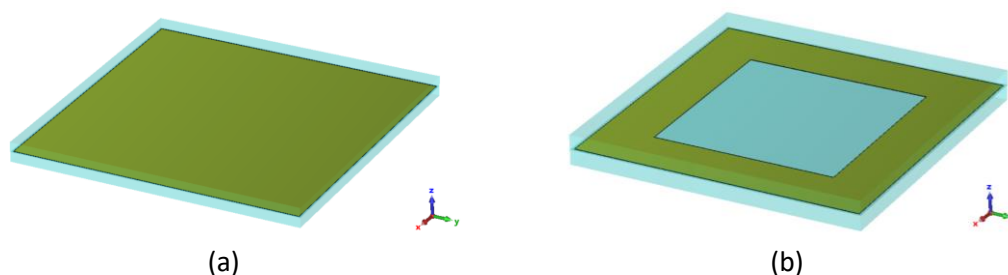
scales with the frequency. For low frequencies below 75 GHz, the metafilm must be designed with a high l/p value to achieve any significant degree of miniaturization.

The AMC metafilm is selected to be integrated into the TPA as a performance enhancement technique. Designing the AMC metafilm with in-phase reflection properties at 2.4 GHz is predicted to improve the antenna's performance and reduce the size of the resonator. The AMC metafilm surface is easier to implement into the patch antenna substrate as there are no vertical elements when compared to the high-impedance or EBG surface. The downside is that the AMC metafilm does not prevent the propagation of surface wave modes. Surface wave effects for microstrip structures are typically ignored when they operate at frequencies below 10 GHz and consist of substrates with low dielectric constants less than 5 [13], therefore surface wave effects is ignored in this research. The AMC also increases the antenna's effective aperture. Using more cells for the AMC ground plane enhances the antennas performance [14].

6.2 Design of Proposed AMC Metafilm

Researchers have developed and incorporated various AMC metafilm structures into the design of patch antennas and other planar antennas. They have found that an AMC square ring [15]–[17], as also shown in Figure 6.14(b) and (d), performs similarly to a solid square AMC or other more complex geometries described in [18]–[23].

Symmetrical shapes are investigated as they exhibit an isotropic wave response for an incident TE or TM plane wave for various angles of incidence. To begin, a square patch is studied for its properties to perform as an AMC. This result is compared to the proposed square ring AMC. The unit cell is designed in CST using the frequency domain solver. The geometry of the unit cell is shown in Figure 6.14 below. The proposed AMC conductor is placed at the centre of two cotton substrates.



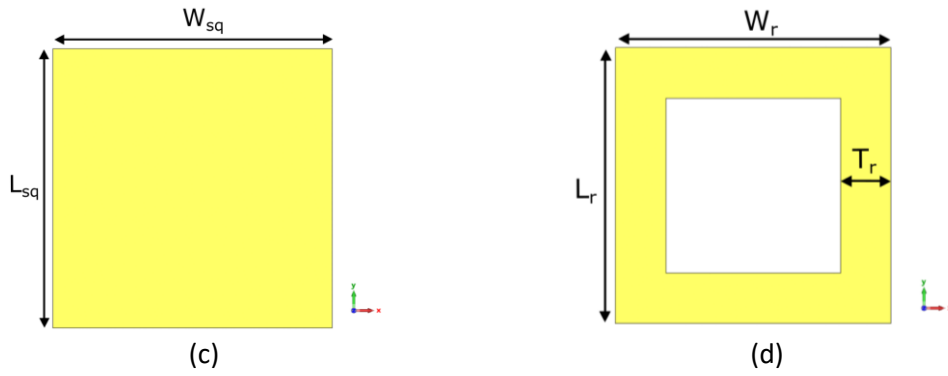
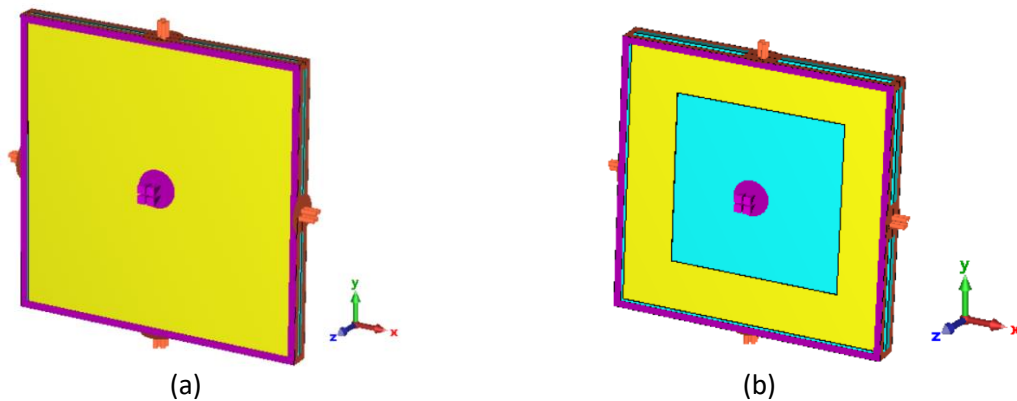


Figure 6.14: Overall unit cell geometry of (a) square AMC and (b) AMC square ring centred between substrates. Front view dimensions of (c) square AMC (d) square ring AMC.

To model the electromagnetic response of the AMC's periodicity, Floquet's theorem can be used which provides the underlying theory of wave propagation in periodic structures [24]. Unit cell analysis in CST is carried out using an excitation method for periodic structures called a 'Floquet port' which allows planes of periodicity modelling. The fundamental Floquet modes that are excited using this port are TM and TE linearly polarized plane waves. The boundary condition defined around the edges of the structures is called a 'Unit Cell' boundary condition, which repeats the modelled structure periodically in two directions up to infinity. In the example given in Figure 6.15(c) the directions of infinite periodicity are along the x and y axis. The open boundary condition on the front of the unit cell across the xy plane as shown in Figure 6.15(a) and (b) is where the Floquet port modes interact with the front face of the AMC structure. The incident fields across the unit cell is the same at every point. A perfect electrical conductor boundary condition is defined at the backside of the AMC unit cell as shown in Figure 6.15(d), which represents a ground plane.

For this analysis, the phase of the reflection coefficient S_{11} is the parameter of interest.



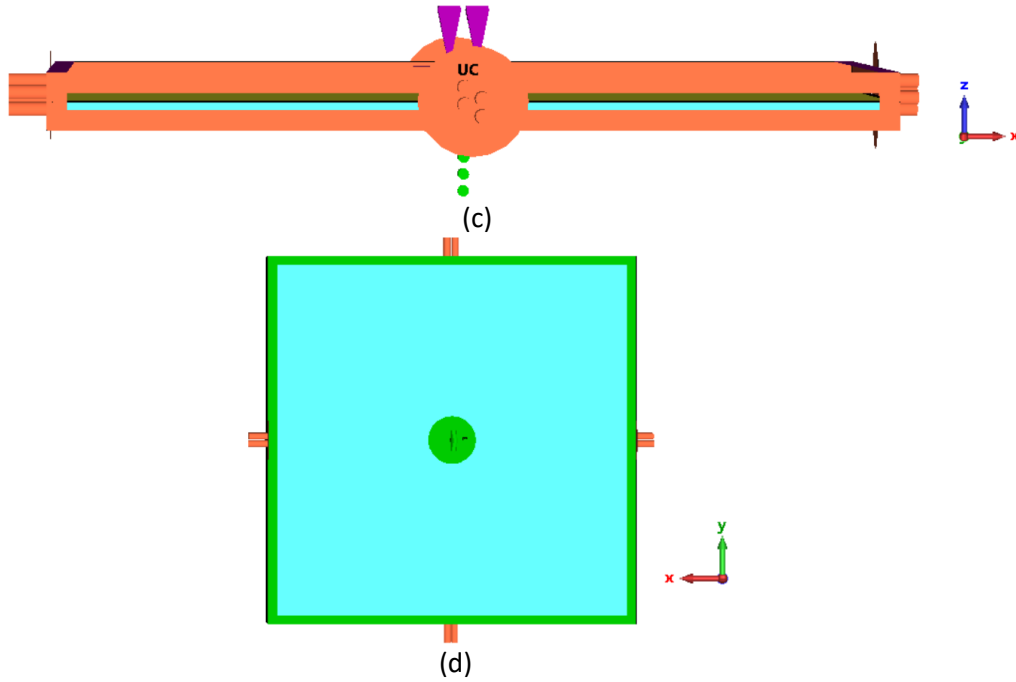


Figure 6.15: Floquet port boundary conditions for a square and square ring AMC. (a) and (b) shows the defined open boundary conditions, (c) is the defined periodic boundary condition around all four edges and (d) is the perfect electrical conductor boundary condition at the base of the structure in place of a ground plane.

The dimensions of the AMC unit cell are optimized so that the reflection phase is 0° at the antenna operating frequency of 2.4GHz.

Dimensions of Square AMC (mm)	Parameter	Value (mm)
Length	L_{sq}	42.1
Width	W_{sq}	42.1
Height of Conductor	H_c	0.035
Height of Top Dielectric	H_{d1}	1.4
Height of Bottom Dielectric	H_{d2}	1.4
Spacing between each unit cell element	G_a	0.3

Table 6.1: Dimensions of Solid Square AMC unit cell.

Dimensions of Square Ring AMC	Parameter	Value (mm)
Length	L_r	27.4
Width	W_r	27.4
Ring Thickness	T_R	5
Height of Conductor	H_c	0.035
Height of Top Dielectric	H_{d1}	1.4
Height of Bottom Dielectric	H_{d2}	1.4
Spacing between each unit cell element	G_a	0.3

Table 6.2: Dimensions of the Square Ring AMC unit cell.

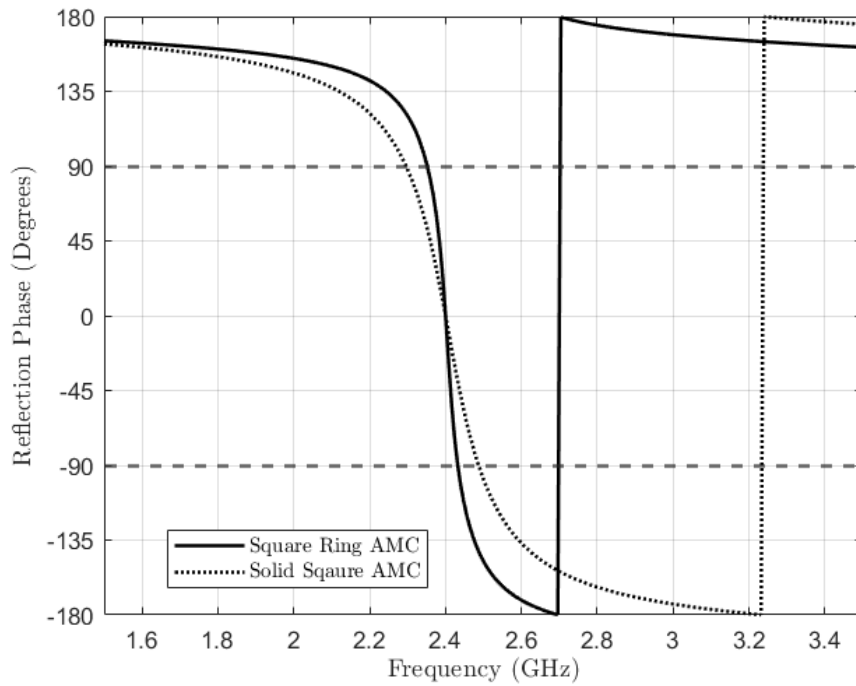


Figure 6.16: Reflection phase comparison of a solid square and square ring AMC metafilm.

AMC Configuration	Bandwidth (%)	Surface Area (mm ²)
Solid Square	7.96	1772.41
Square Ring	3.35	447.6

Table 6.3: AMCs bandwidth performance in comparison to its occupied surface area at a zero-degree reflection phase at 2.4 GHz.

The bandwidth for the AMCs in Figure 6.16 is defined as the region of frequencies around the zero-degree phase reflection between $+90^\circ$ to -90° . The extracted bandwidth and surface area profile of the AMC unit cell is shown in Table 6.3. The bandwidth of the solid square AMC unit cell is 7.96%, while the square ring AMC is 3.35%. The square ring AMC offers a size reduction of 11.1 mm for both the length and width and a conductive surface area size reduction of 1324.81 mm², making the conductive surface area profile 74.7% less than that of the solid square. However, the size reduction comes at a cost in which the zero-degree phase reflection bandwidth of the AMC is reduced by 4.61%.

A circuit analysis of the AMC structure can give insight into why the miniaturization occurs for the AMC cells.

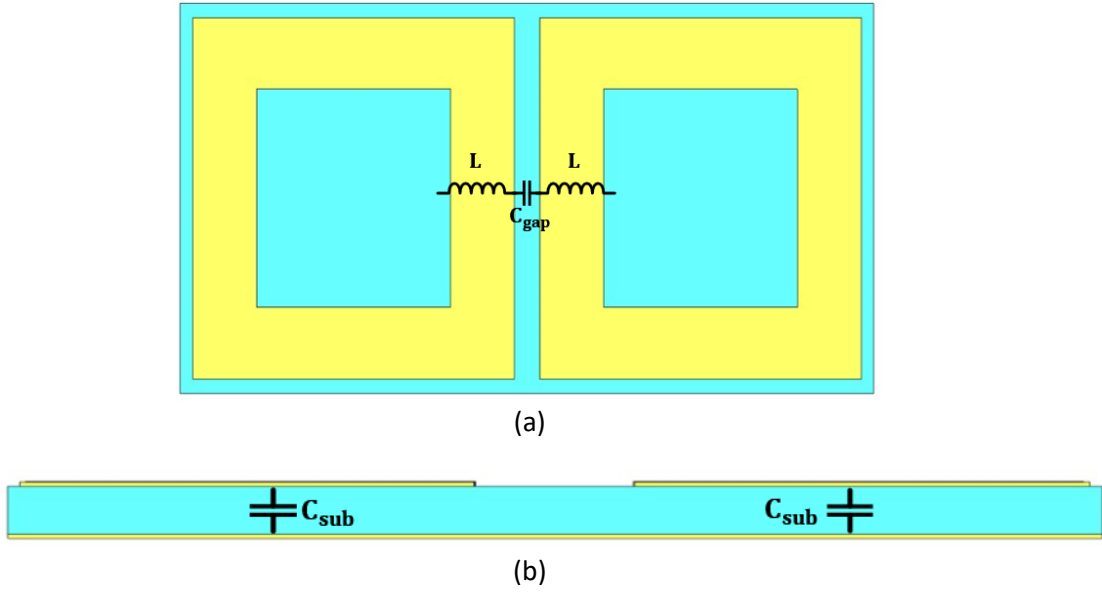


Figure 6.17: Adjacent AMC unit cells with equivalent circuit elements (a) front view and (b) cross-sectional view.

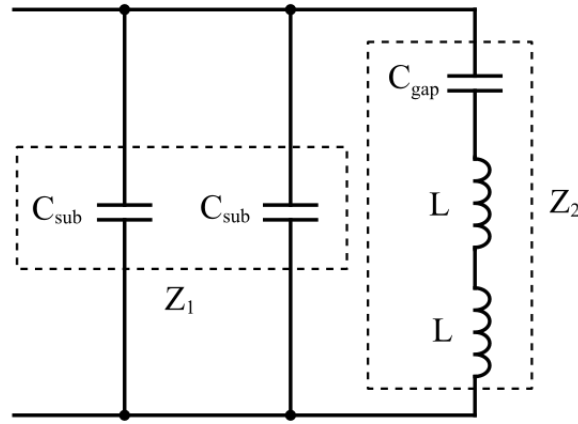


Figure 6.18: Equivalent circuit model for two-unit cells of the proposed AMC.

Using the equivalent circuit model in Figure 6.18, the total surface impedance can be calculated by combining the impedances Z_1 and Z_2 in parallel.

From Figure 6.18, the total surface impedance Z_s can be calculated as

$$Z_s = \frac{Z_1 Z_2}{Z_1 + Z_2} = \frac{-\frac{2j}{\omega C_{\text{sub}}} \left(-\frac{j}{\omega C_{\text{gap}}} + 2j\omega L \right)}{-\frac{2j}{\omega C_{\text{sub}}} + \left(-\frac{j}{\omega C_{\text{gap}}} + 2j\omega L \right)} \quad (6.13)$$

And simplified to

$$Z_s = \frac{-2j(2C_{\text{gap}}L\omega^2 - 1)}{\omega(2C_{\text{gap}}C_{\text{sub}}L\omega^2 - 2C_{\text{sub}} - C_{\text{gap}})} \quad (6.14)$$

The resonant frequency can be expressed when the sum of the reactive impedances equal zero.

$$Z_1 + Z_2 = -\frac{2j}{\omega C_{\text{sub}}} + \left(-\frac{j}{\omega C_{\text{gap}}} + 2j\omega L \right) = 0 \quad (6.15)$$

$$f_{\text{res}} = \frac{1}{2\pi} \sqrt{\frac{(C_{\text{gap}} + 2C_{\text{sub}})}{2C_{\text{gap}}C_{\text{sub}}L}} \quad (6.16)$$

Thus, a relationship can be established between the resonant frequency and the effective inductance L of the square ring and square AMC. For the square patch AMC, the inductance L will be small since the induced currents can spread further over the conductive surface decreasing the effective inductance. Tuning the size of the gap in the ring increases the effective inductance around the rectangular ring and reduces the frequency at which the 0° reflection phase occurs for the same dimensions.

6.3 Effective Parameters of the Proposed AMC Metafilm

In this section the proposed AMC structure is analysed using the effective medium theory [25] to extract the material properties such as the refractive index, permittivity, and electric susceptibility through a full wave simulation analysis. This analysis provides a full characterization of the dielectric properties of the AMC metafilm that is loaded into the dielectric substrate of the patch antenna.

CST offers a result template to extract the material properties for some inhomogeneous medium simulated between two ports through the S-parameters. The result template is based on the effective medium theory presented by X. Chen et al. [25]. Once the S-parameters are known through either simulation or measurements, the dielectric medium's permittivity and permeability are extracted using equations (6.17)-(6.20).

$$Z = \pm \sqrt{\frac{(1 + S_{11})^2 - S_{21}^2}{(1 - S_{11})^2 - S_{21}^2}} \quad (6.17)$$

$$e^{jnk_0d} = \frac{1}{2S_{21}(1 - S_{11}^2 + S_{21}^2)} \pm j \sqrt{\left(1 - \frac{1}{2S_{21}(1 - S_{11}^2 + S_{21}^2)}\right)^2} \quad (6.18)$$

$$\varepsilon = \frac{n}{Z} \quad (6.19)$$

$$\mu = nZ \quad (6.20)$$

where Z is the wave impedance, S_{11} is the reflectance, S_{21} is the transmittance, n is the refractive index, k_0 is the free space wavenumber, d is the thickness of the medium, ε is the permittivity and μ is the permeability of the medium.

The simulated unit cell analysis for the AMC is the same as that of Figure 6.15(c). The only difference is that the defined electric boundary condition at $z = 0$ across the xy plane is replaced by another Floquet port creating a two-port system. This allows the computation of the reflectance S_{11} and transmittance S_{21} through the unit cell as shown in Figure 6.19.

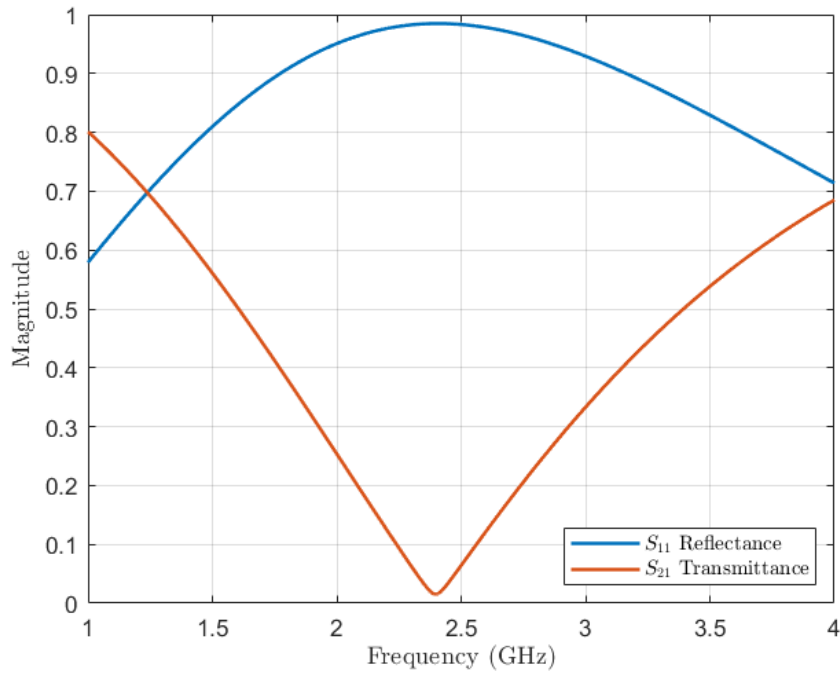


Figure 6.19: Reflectance and Transmittance across the unit cell metafilm.

Figure 6.19 shows that the maximum reflectance and minimum transmittance of the incident waveform occurs at 2.4 GHz. Figure 6.20 shows that the transmitted wave S_{21} has a zero-degree phase reflection at 2.4 GHz.

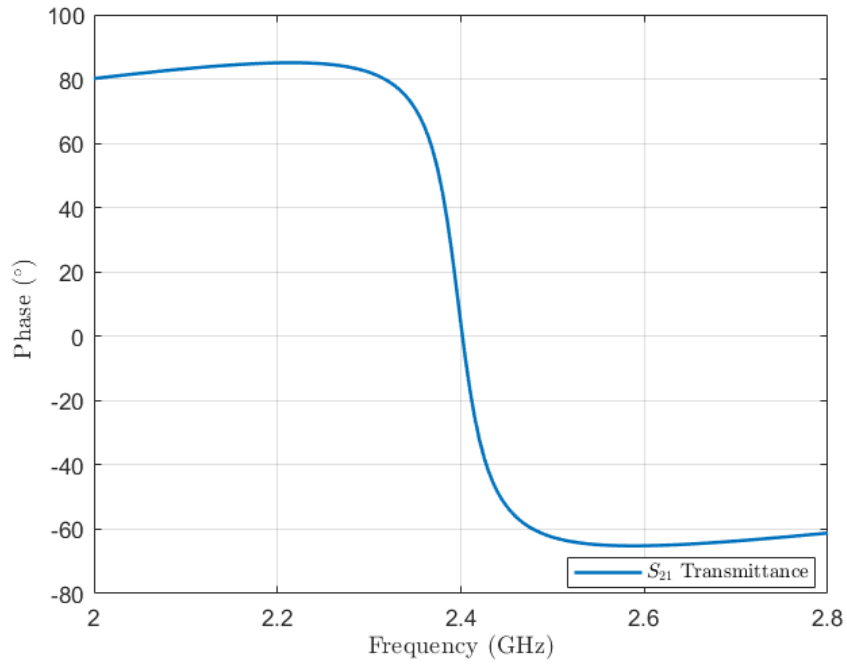


Figure 6.20: Phase of the transmittance of the unit cell metafilm

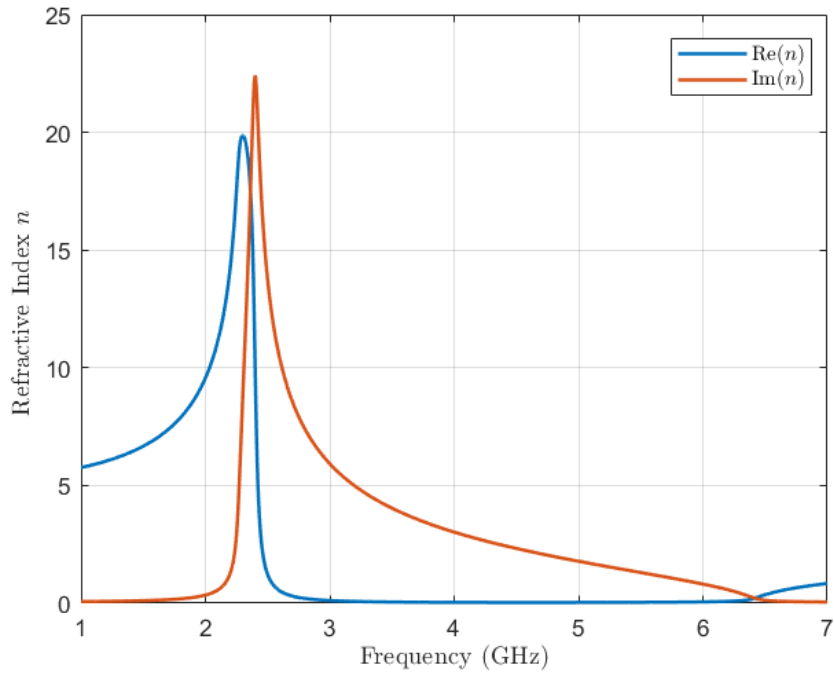


Figure 6.21: Complex Refractive Index of the unit cell metafilm.

The complex refractive index n can be defined as

$$n = n_0 + j\kappa = \pm\sqrt{\mu_r\epsilon_r} \quad (6.21)$$

Where n_0 is the real part of the refractive index which is also known as the ordinary refractive index. The ordinary refractive index describes how much a wave slows down inside the material relative to the speed of light. The term κ is the extinction coefficient which is the imaginary part which describes the attenuation of the wave due to losses [26]. From Figure 6.21 the peak of the ordinary refractive index n_0 occurs just below resonance at 2.3 GHz with a value of 19.9. At resonance the ordinary refractive index n_0 is 10, while the extinction coefficient κ is at a maximum of 22.4. This implies that at resonance the structure is exhibiting its maximal reflectance properties which agrees well with Figure 6.19. Above resonance the ordinary refractive index n_0 approaches near to zero with a value of 0.03, which begins to slowly rise at approximately 6.42 GHz. The extinction coefficient decreases exponentially after resonance approaching near zero above 6.42 GHz.

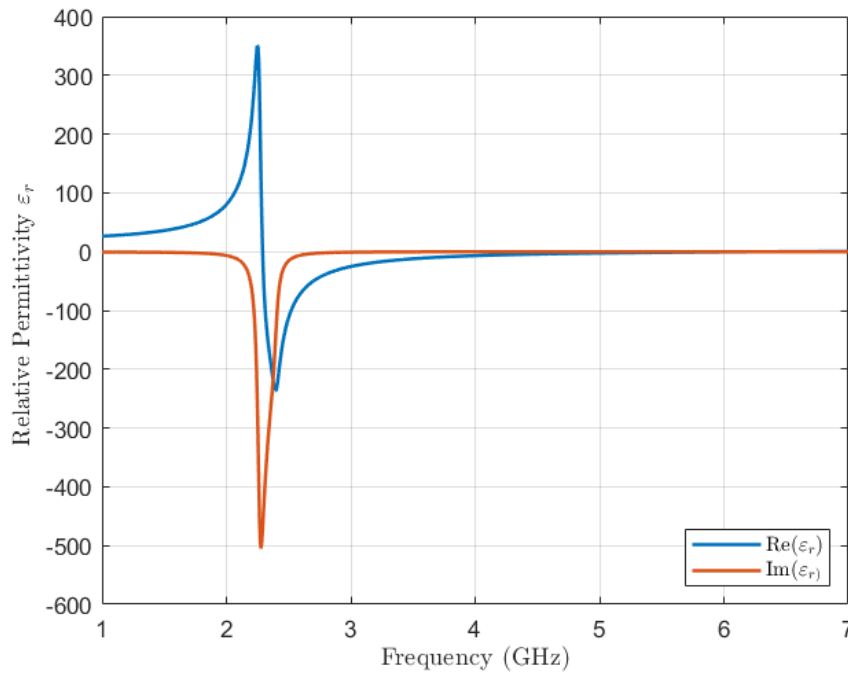


Figure 6.22: Relative permittivity of the unit cell metafilm.

The permittivity $\tilde{\epsilon}_r$ as a function of frequency can be defined as

$$\tilde{\epsilon}_r = \epsilon_r + j\epsilon_r'' \quad (6.22)$$

where ϵ_r is the real part and ϵ_r'' is the imaginary part of permittivity. This version of permittivity $\tilde{\epsilon}_r$ is a complex function and considers the losses of the material through the imaginary part. In radio frequency analysis it is common to specify a dielectric material through the real part of permittivity and its loss tangent $\tan \delta$ using [27]

$$\tilde{\epsilon}_r = \epsilon_r(1 - j \tan \delta) \quad (6.23)$$

Figure 6.22 shows that below resonance the imaginary part is zero, while the real part starts to decrease exponentially from $\epsilon_r = 22$ to a point below resonance at 2.25 GHz, peaking at $\epsilon_r = 330$. Just below resonance at 2.28 GHz the imaginary part decreases to a minimum of $\epsilon_r'' = -504$. At resonance the real part ϵ_r exhibits a negative permittivity decreasing to a value of $\epsilon_r = -236$. The imaginary part ϵ_r'' at resonance is approximately $\epsilon_r'' = -132$, which indicates high loss which is consistent with extinction coefficient κ in Figure 6.21. Above resonance the permittivity tends to zero where the metafilm unit cell starts to behave like a vacuum, reaching a zero-crossing point at approximately 6.42 GHz, which slowly keeps increasing until the second mode of resonance where it will exhibit similar properties again.

The surface susceptibilities are said to be the most appropriate way of characterizing a metafilm. The electric susceptibility can be calculated from the effective medium approach [25] using the simulated permittivity in Figure 6.22 and computing $H_d(\epsilon_r - 1)$. The electric susceptibility can also be calculated using (6.5) or (6.9) derived from the generalized sheet transition conditions (GSTCs) using the simulated values for reflectance S_{11} and transmittance S_{21} of the unit cell analysis.

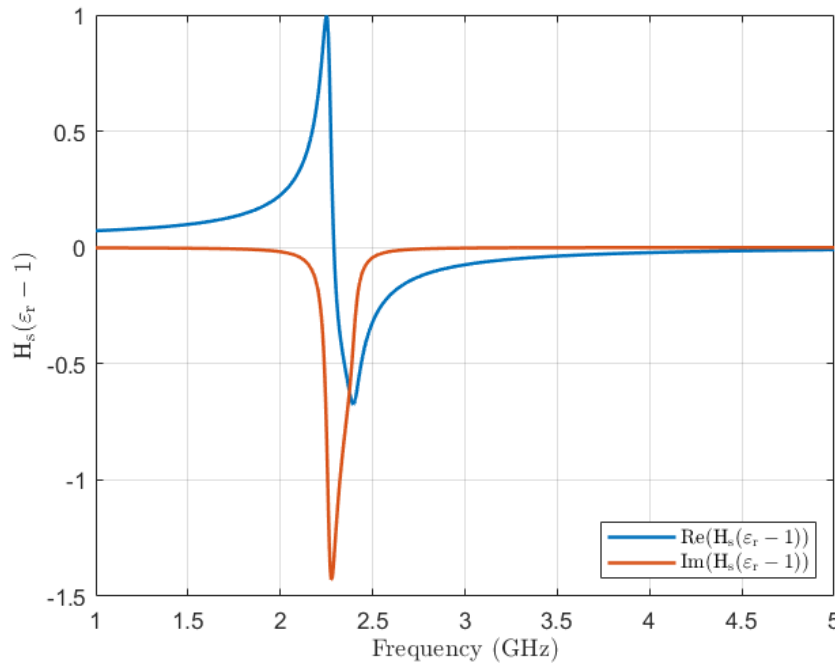


Figure 6.23: Electric susceptibility $H_d(\epsilon_r - 1)$ extracted using the effective medium approach of the AMC metafilm unit cell.

The electric susceptibility using the effective medium approach and using (6.5) based on the GSTCs is plotted in Figure 6.23 and Figure 6.24. The computed electric susceptibility in Figure 6.23 shows the same trend as the permittivity in Figure 6.22 which is expected. This form of susceptibility is a function of the substrate's height H_d and thus its units are in meters.

Figure 6.24 shows the electric susceptibility which is calculated using (6.5) or (6.9) for an incident TM or TE polarized wave. Since the geometry of the proposed AMC metafilm unit cell is symmetrical the susceptibility dyadic along the x and z component is equal ($\chi_{ES}^{xx} = \chi_{ES}^{zz}$). The simulated reflectance and transmittance are the two parameters required to characterize the substrate's susceptibility.

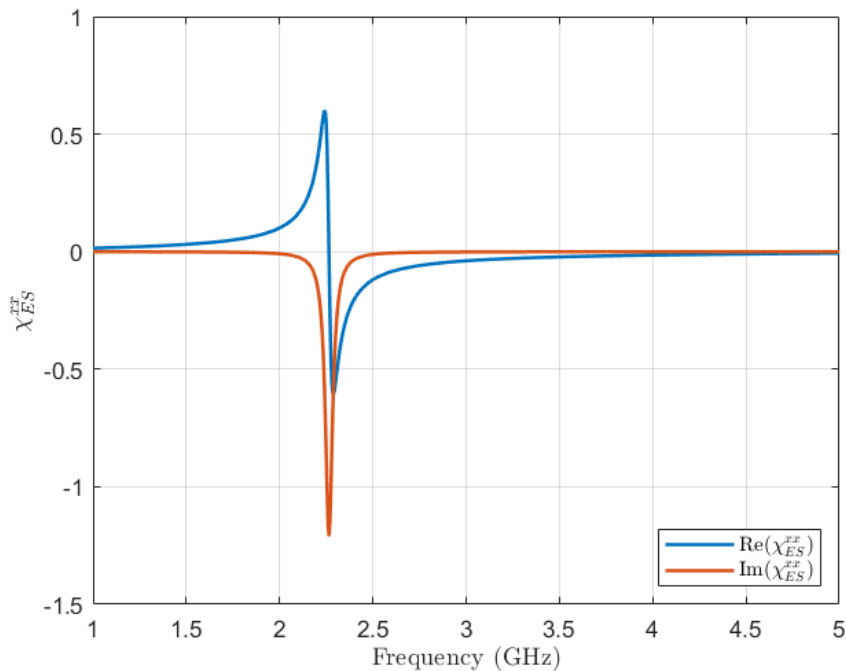


Figure 6.24: Electric susceptibility χ_{ES}^{xx} using the equations derived from the generalized sheet transition conditions (GSTCs).

A closer comparison between two methods of extracting the electric surface susceptibility is shown in Figure 6.25 and Figure 6.26, where the methods are compared with each other for the real and imaginary part respectively. While both results are similar there are differences when comparing the real and imaginary parts of the susceptibility in Figure 6.25 and Figure 6.26. A comparison must be made with measured results.

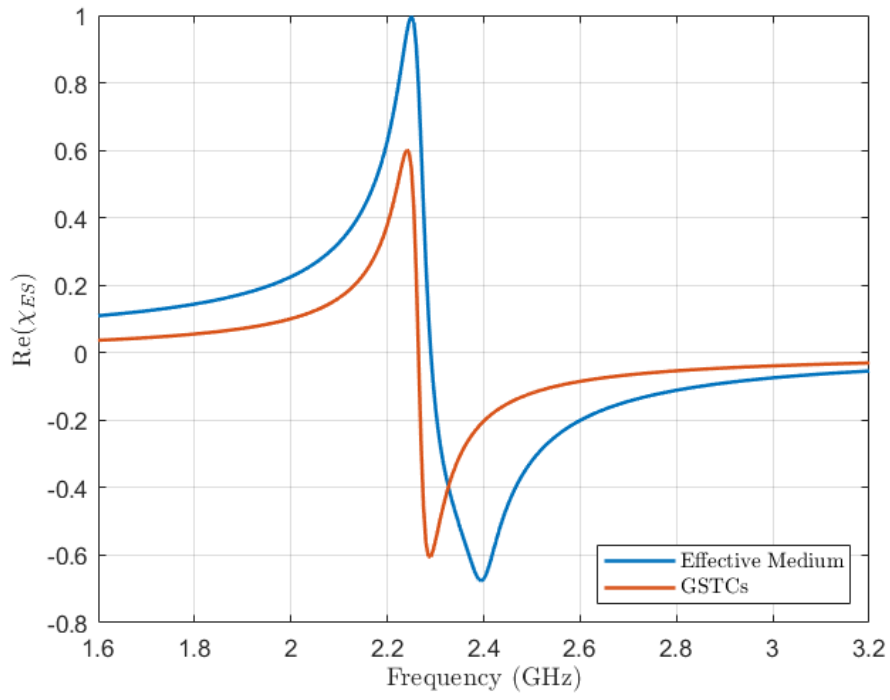


Figure 6.25: Comparison of the real electric susceptibilities from computed using the effective medium approach and the generalized sheet transition conditions (GSTCs) approach.

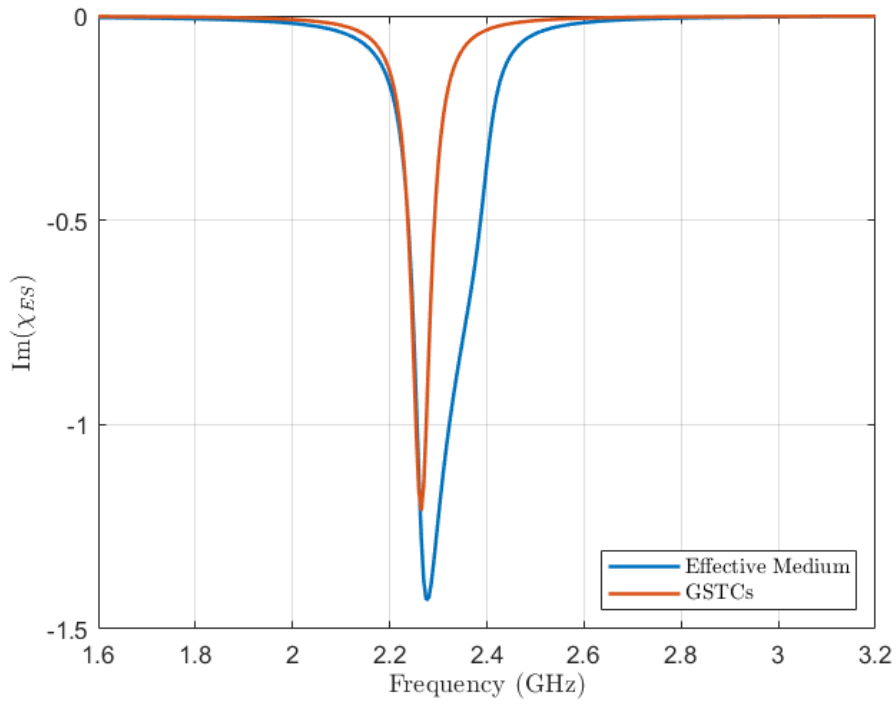


Figure 6.26: Comparison of the imaginary electric susceptibilities from computed using the effective medium approach and the generalized sheet transition conditions (GSTCs) approach.

Overall, using the GSTCs to extract the susceptibility of any given metasurface is argued as a more natural method to model the macroscopic interaction with an electromagnetic field since the method is independent from the medium's thickness [4]. The susceptibility coefficients in (6.1) and (6.2) are uniquely defined for a given metasurface. Embedding the AMC metafilm into the substrate showed that the dielectric properties vary as a function of frequency, exhibiting a very high dielectric constant at the operating frequency.

6.4 Integrating the AMC Metafilm in The Trapezoidal Patch Antenna

In this section the AMC metafilm is integrated into the trapezoidal patch antenna and simulated in CST Microwave Studio to investigate its radiation performance.

Implementing the AMC structure within the edge fed trapezoidal patch antenna is not trivial. Simply removing the ground plane and replacing it with a periodic array cannot be done since it disrupts the quasi-TEM line that is formed between free space, the feedline, and the ground plane. One method of implementing the AMC metafilm is by converting the ground plane to a partial ground plane up to the radiating element, in turn creating a monopole antenna [28]. Another method is the use of a coplanar waveguide [29]–[32] with the AMC placed underneath. For a centre fed patch antenna the placement of the AMC metafilm had a significant impact on the impedance matching [16]. Thus, the AMC was placed on the same layer as the patch antenna, with a partially reflective surface placed on top of the antenna. Another method of integrating the AMC metafilm is presented by S. Yan et al. [15] where a centre fed planar antenna consists of a an AMC metafilm which is loaded into the substrate. The antenna consists of three layers with the radiating element at the top, an AMC metafilm loaded between the substrate and a ground plane below. A via is placed from the ground plane to the radiating element.

All the designs mentioned add multiple layers and complexity to integrating the AMC metafilm. Sievenpiper [1] incorporated a high impedance surface into the dielectric substrate, placed within the substrate at a distance ' t ' away from the ground plane. As shown in Figure 6.27 the substrate and the high impedance surface combined are considered as an effective dielectric slab. The gap of height ' t ' between the high impedance surface and the metal ground consists of a substrate with its own dielectric properties.

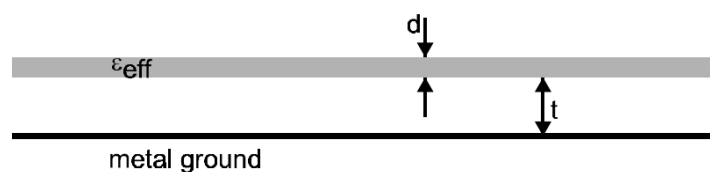


Figure 6.27: The equivalent effective medium model for a metafilm loaded into a dielectric substrate [1].

The high impedance surface without a ground plane are poor reflectors. With an added ground plane behind the structure becomes extremely reflective as well as allowing the control of the phase properties [1].

The proposed configuration is to place the AMC metafilm in the middle of the substrate, below the radiating element and the radiating edges. The proposed configuration is an alternative method to that proposed by S. Yan et al. [15] which is designed for an edge fed antenna configuration. An AMC metafilm array configuration is placed in the middle of the substrate where each element is spaced 0.3 mm apart. A single AMC cell is removed as to not disrupt the TEM line that is formed along the microstrip feed line towards the radiating element as shown in Figure 6.28.

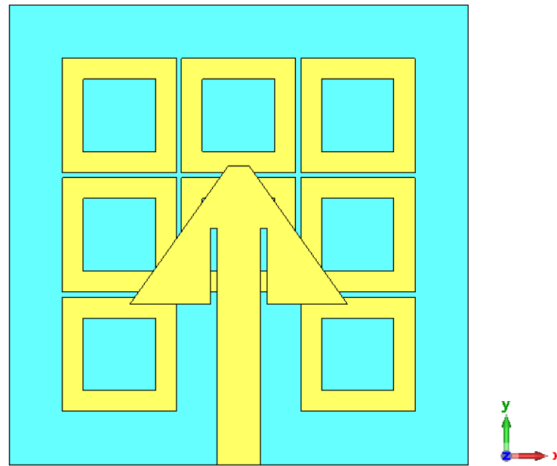


Figure 6.28: Front view of the proposed AMC metafilm with the trapezoidal patch superimposed.

The TPA-MF is designed and simulated in CST Microwave Studio. The geometry of the radiating element and feedline are shown in Figure 6.29. The optimised dimensions are shown in Table 6.4 with a comparison of the dimensions of the inset-fed rectangular and trapezoidal patch antenna.

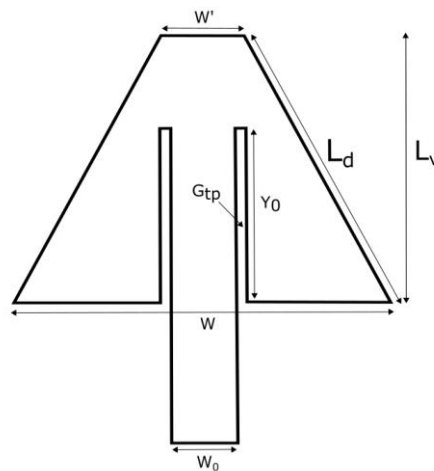


Figure 6.29: Front View Geometry of Trapezoidal Patch with a loaded AMC metafilm.

Parameter	Symbol	RPA (mm)	TPA (mm)	TPA-MF (mm)
Vertical Patch Length	L_v	48.5	52.0	34.0
Diagonal Patch Length	L_d	-	51.5	40.5
Aperture #1 Width	W	52.0	52.0	52.0
Aperture #2 Width	W'	52.0	5.0	5.0
Inset Recessed Distance	y_0	13.5	13.8	19.9
Feedline Width	W_0	10.5	10.5	10.5
Total Substrate Height	H_s	2.8	2.8	2.835
Inset Feed Gap Width	G_{tp}	2.6	1.5	1.5

Table 6.4: Dimensions of Rectangular patch, Trapezoidal patch, and Trapezoidal patch with an AMC metafilm.

The dimensions of the radiating element of the trapezoidal patch antenna in Chapter 4 are used as the template for the initial design for the TPA-MF. The addition of the AMC metafilm results in a resonance much lower than 2.4 GHz and a poor impedance match, thus optimisation is required on the dimensions of the radiator. During the optimisation process, the vertical length L_v is required to be smaller to maintain a resonant frequency of 2.4 GHz. It was also found that the inset feed depth y_0 had to be increased to maintain an input impedance match at 50 Ω .

The simulated performance parameters are shown in Table 6.5 where a comparison is made to the performance parameters of a RPA and TPA. The TPA-MF shows an increase in the antenna's bandwidth, realized gain and directivity. The simulated TPA-MF has a directivity of 8.7 dBi which is an increase by 0.65 dBi compared to the TPA. The realized gain is 5.92 dBi which is an increase by 1.37 dBi. The TPA-MF's radiating element is shown to be 55 % smaller than the RPA and 26.3 % smaller than the TPA.

Patch Antenna Type	f_{res} (GHz)	Bandwidth (%)	e_0 (%)	G_{re} (dBi)	D (dBi)	Radiating Element Surface area (mm^2)	Area Size Reduction (%)
RPA	2.40	2.8	57.7	5.7	8.11	2860.9	0
TPA	2.40	2.1	47.5	4.55	8.05	1748.6	38.9
TPA-MF	2.40	2.9	52.7	5.92	8.7	1288.1	55.0

Table 6.5: Performance parameter comparison of a RPA, TPA and TPA-MF.

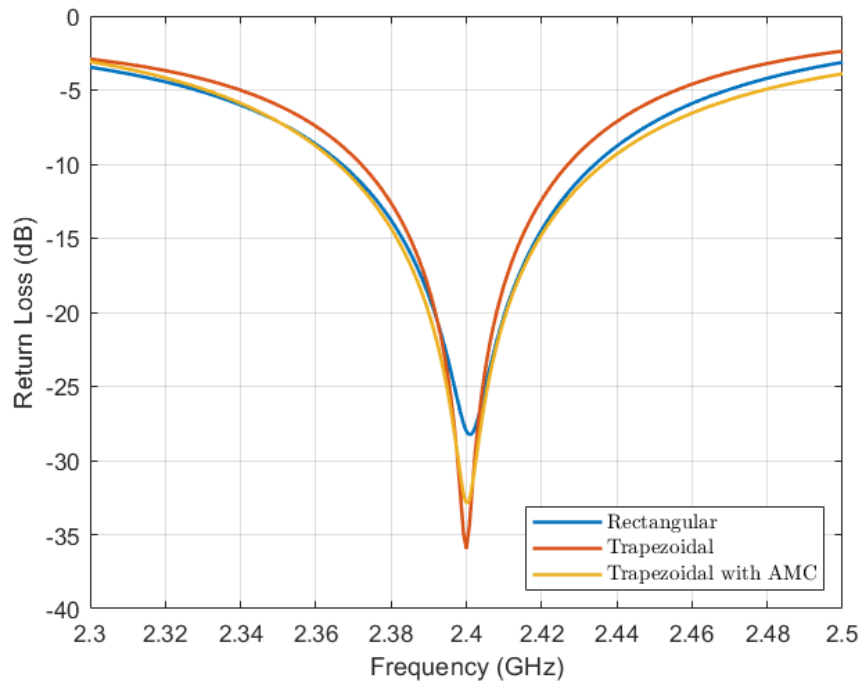
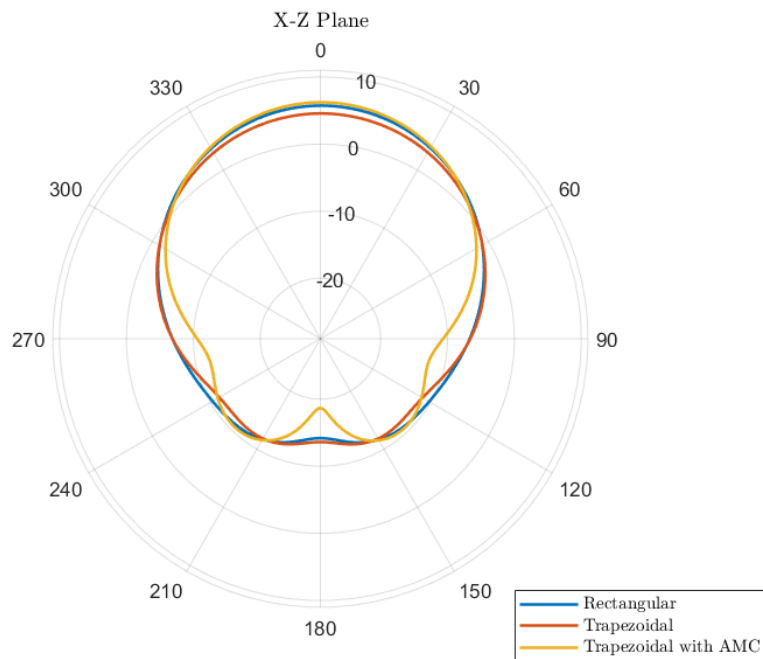


Figure 6.30: Simulated input reflection coefficient comparison between a rectangular patch antenna, trapezoidal patch antenna and trapezoidal patch antenna with an AMC metafilm loaded into the dielectric substrate.

The simulated reflection coefficient for the RPA, TPA and TPA-MF are compared to each other in Figure 6.30. The extracted bandwidth of the TPA-MF is shown to be 2.9% which is greater than the TPA by 0.8%.



(a)

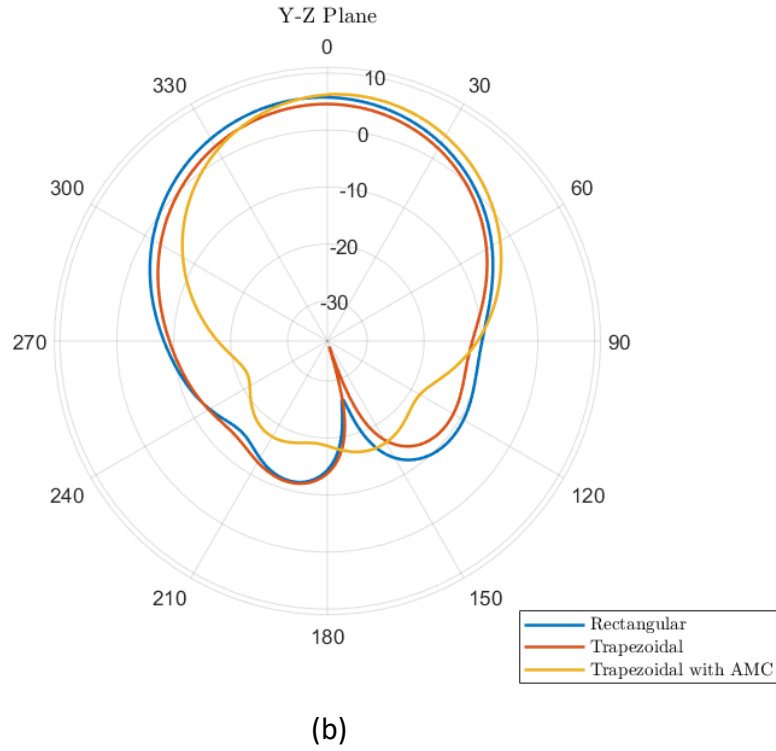


Figure 6.31: Simulated radiation patterns in terms of realized gain for the antennas (a) XZ plane and (b) YZ plane.

The antenna's far-field pattern is shown in Figure 6.31 which is compared to an RPA and a TPA. The X-Z or H-plane in Figure 6.31(a) shows that the TPA-MF has smaller side lobes than the RPA and TPA while also being the most directive at $\theta = 0$. The Y-Z or E-plane in Figure 6.31(b) shows that while the antenna remains most directive at $\theta = 0$ when compared to an RPA and TPA, the antenna also exhibits directive properties from 0° to 30° .

6.5 Antenna Bending Analysis

A simulated bending analysis is performed for the TPA-MF and compared to the performance of a TPA. Figure 6.33 shows the bending configurations to be analysed.

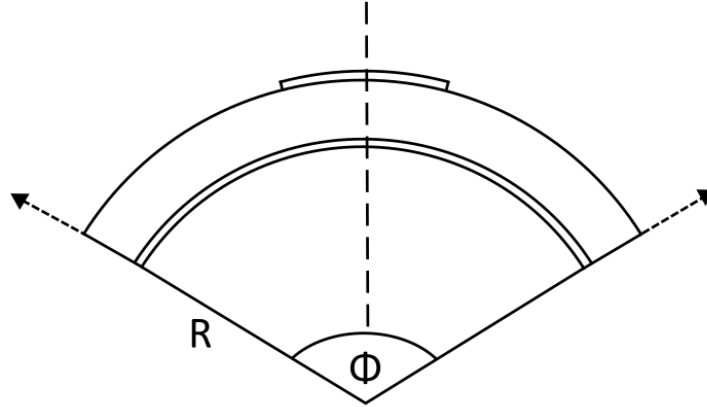


Figure 6.32: Cross sectional view for antenna bending modelled as an annulus bent over an invisible cylinder with a sectoral bending angle Φ and radius R .

It is worth noting that the antenna bending angle is defined by the radial distance to the edge of the antennas ground plane rather than the radial distance to edge of the antennas radiating element as used by C. Krowne [33] and L. Song et al. [34], since this definition encompasses the bending of the antenna as a whole than just the bending of the radiating element.

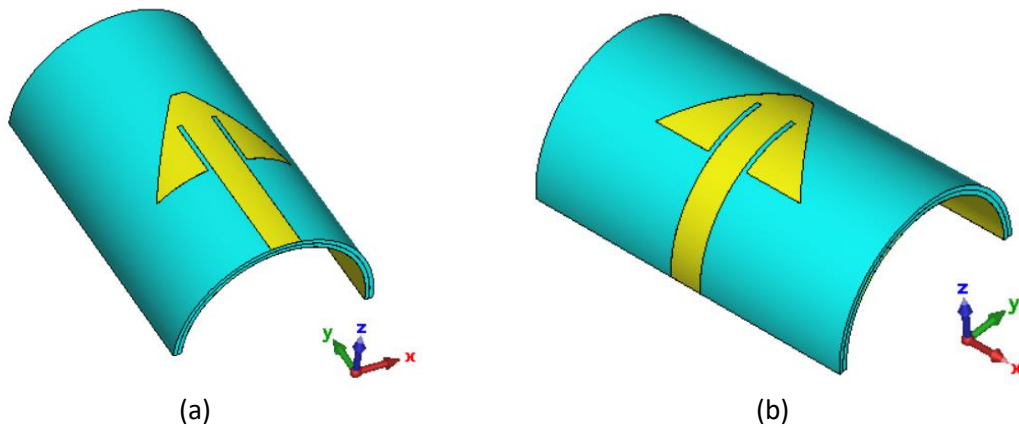


Figure 6.33: TPA-MF cylindrically bent along its (a) H-plane and (b) E-plane.

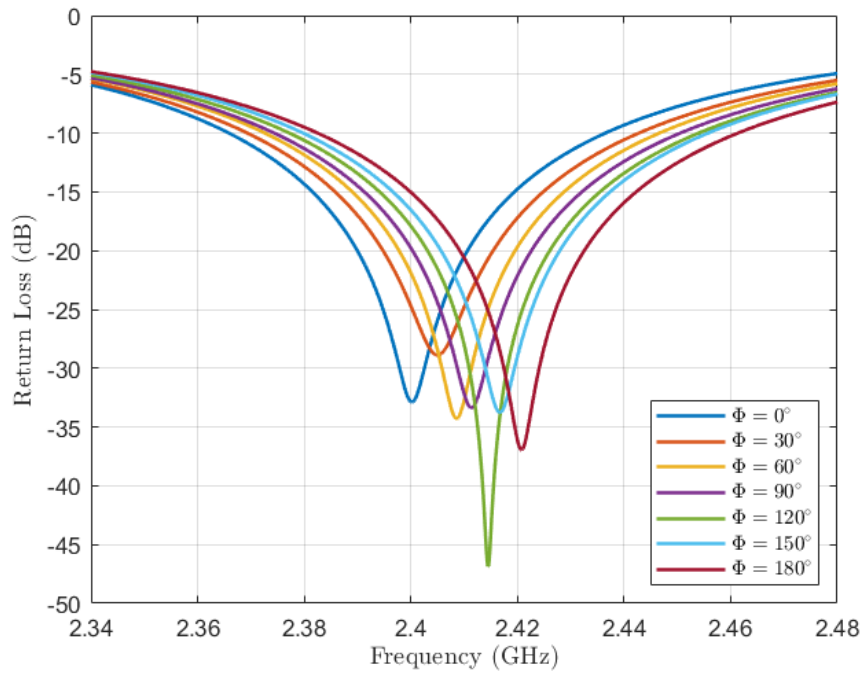
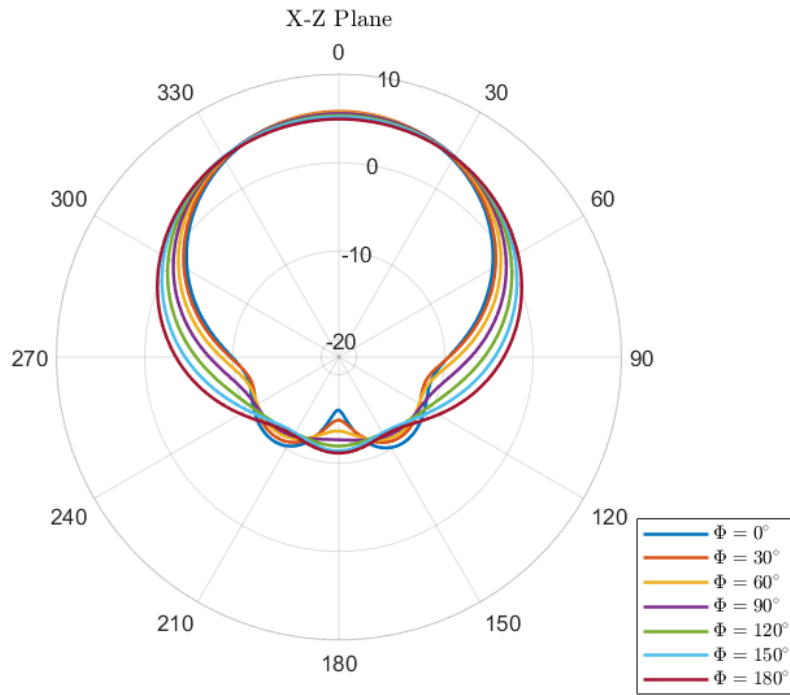


Figure 6.34: Simulated input reflection coefficient for H-plane bending of the TPA-MF.

Bending along the XZ or H-plane results in a small shift in the resonant frequency. A total shift of 8.2 MHz occurs from its flat condition to an extreme bend of 180°. Although the operating frequency of 2.4 GHz changes as the bending angle increases, all bending angles include 2.4 GHz as part of its bandwidth, so the operation of the antenna is still satisfactory. Figure 6.34 shows the gradual detune of the resonant frequency for the various bending angles.

Bending Angle θ (°)	f_{res} (GHz)	BW (%)	e_0 (%)	D (dBi)	G_{re} (dBi)
0	2.400	2.94	52.7	8.70	5.92
30	2.405	3.04	53.7	8.61	5.91
60	2.409	3.07	54.7	8.40	5.78
90	2.411	3.16	55.9	8.15	5.63
120	2.414	3.18	56.8	7.85	5.38
150	2.417	3.17	58.1	7.56	5.21
180	2.421	3.25	59.3	7.22	4.96

Table 6.6: Performance parameters comparison for XZ plane bending angles.



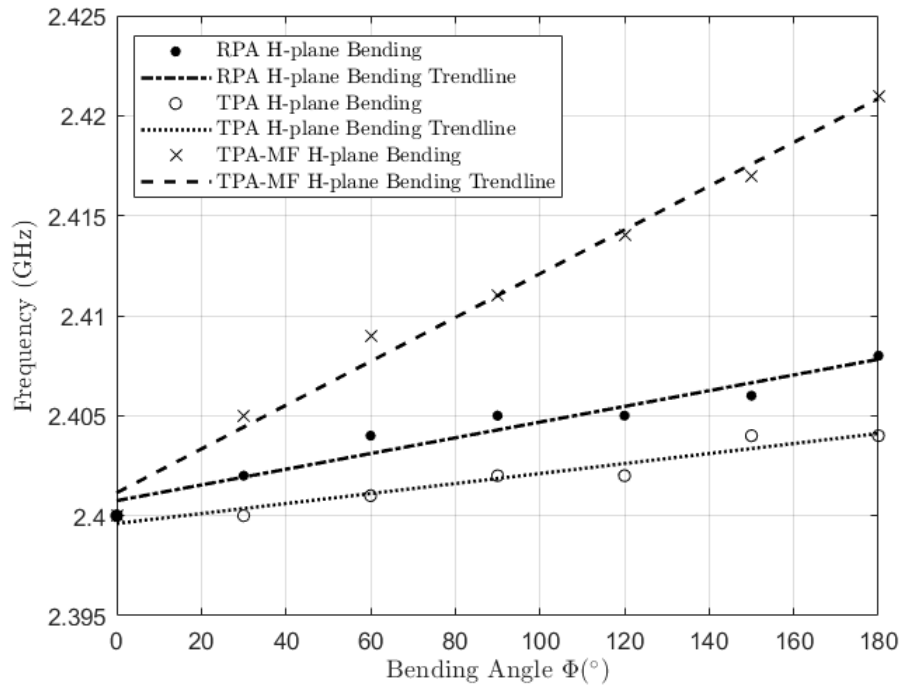


Figure 6.36: Resonant frequency shift comparison between the antennas with respect to increasing bending angles across the H-plane.

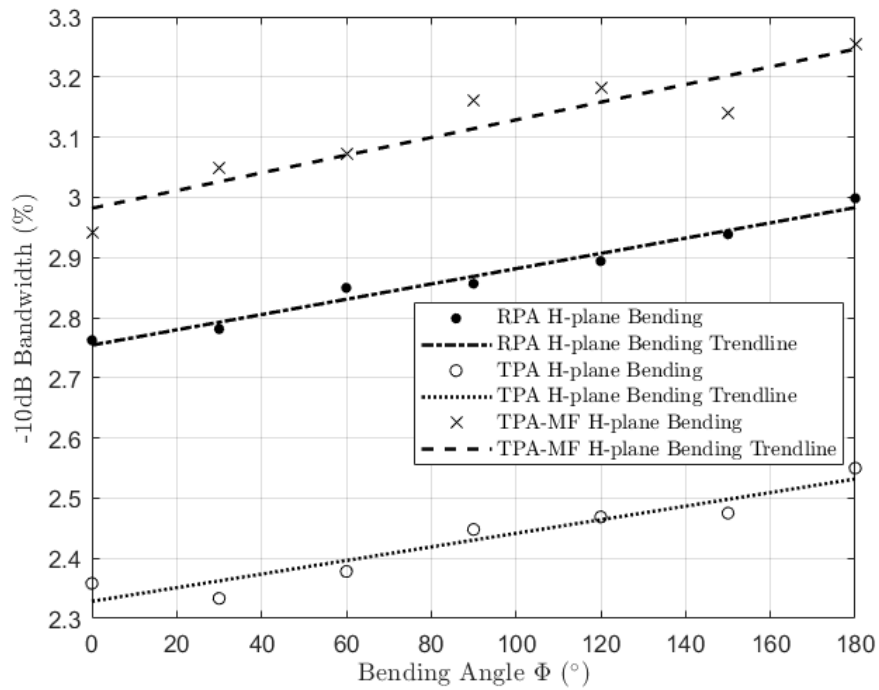


Figure 6.37: Bandwidth comparison between the RPA, TPA and TPA-MF while bending along the H-plane.

H-plane bending the TPA-MF in comparison to the RPA and TPA is shown in Figure 6.36. The TPA-MF is shown to exhibit a higher shift in the resonant frequency compared to the RPA and TPA. For a bending angle of 180° the TPA-MF detunes by almost 20 MHz compared to the TPA. Even with the resonant frequency detuning being much higher, the bandwidth contains the intended resonant frequency of 2.4 GHz. Comparing the bandwidth of the TPA-MF to the RPA and TPA, the TPA-MF has the highest bandwidth for all bending angles investigated. The bandwidth is on average 0.7% higher for all bending angles compared to the TPA.

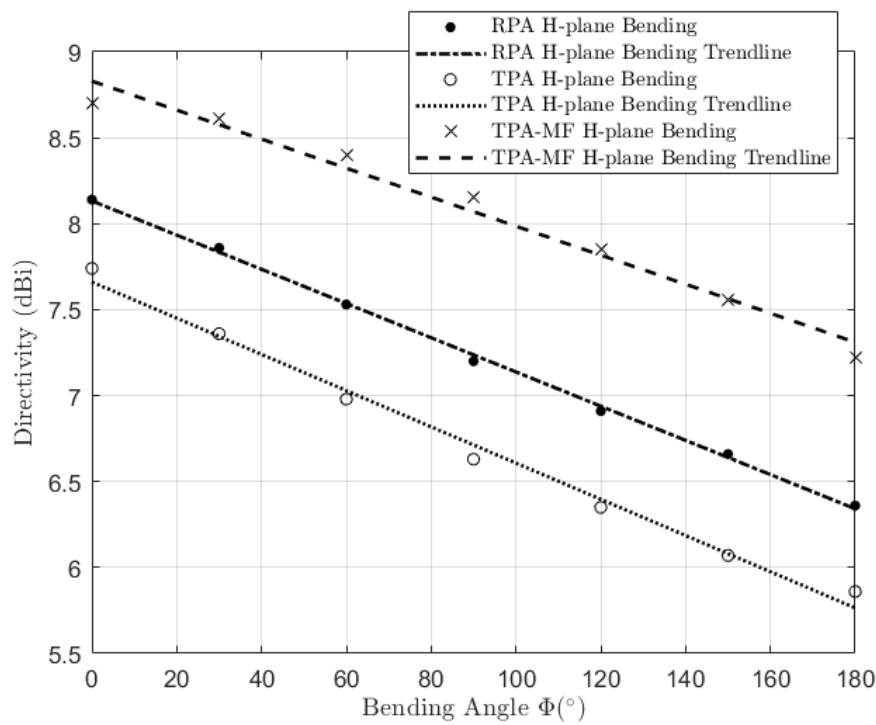


Figure 6.38: Directivity comparison between the RPA, TPA and TPA-MF while bending along the H-plane.

Increasing the bending angle to 180-degrees across the H-plane from its flat condition is shown to decrease the directivity and realized gain for the TPA-MF. The directivity reduces from 8.7 dBi to 7.2 dBi, while the realized gain reduces from 5.92 dBi to 4.96 dBi. A comparison is made between the RPA and TPA as shown in Figure 6.38 and Figure 6.39. The TPA-MF outperforms the RPA and TPA in terms of directivity and realized gain for all the investigated bending angles.

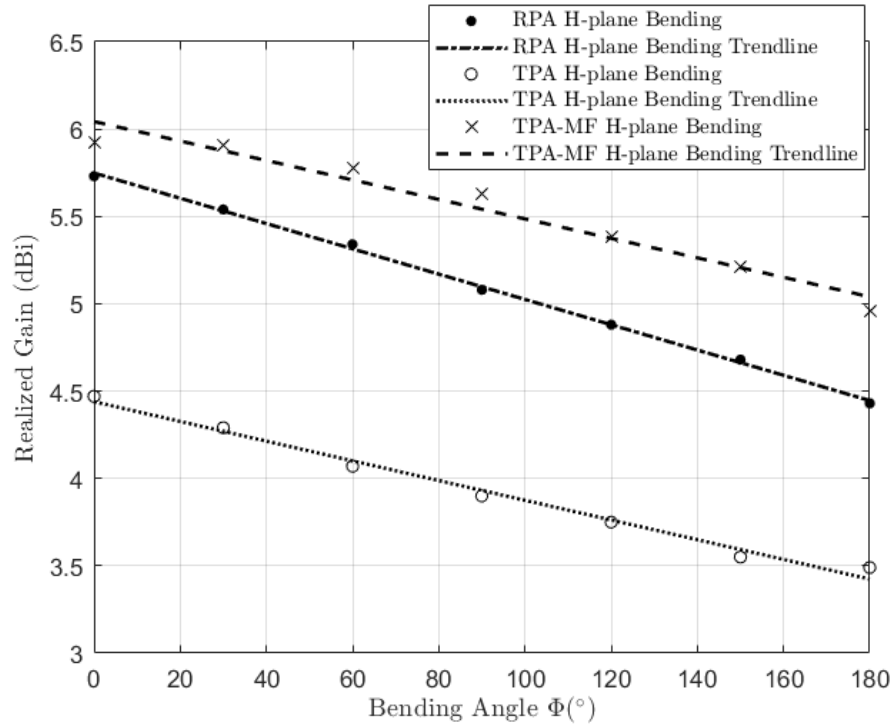


Figure 6.39: Realized Gain comparison between the RPA, TPA and TPA-MF while bending along the H-plane.

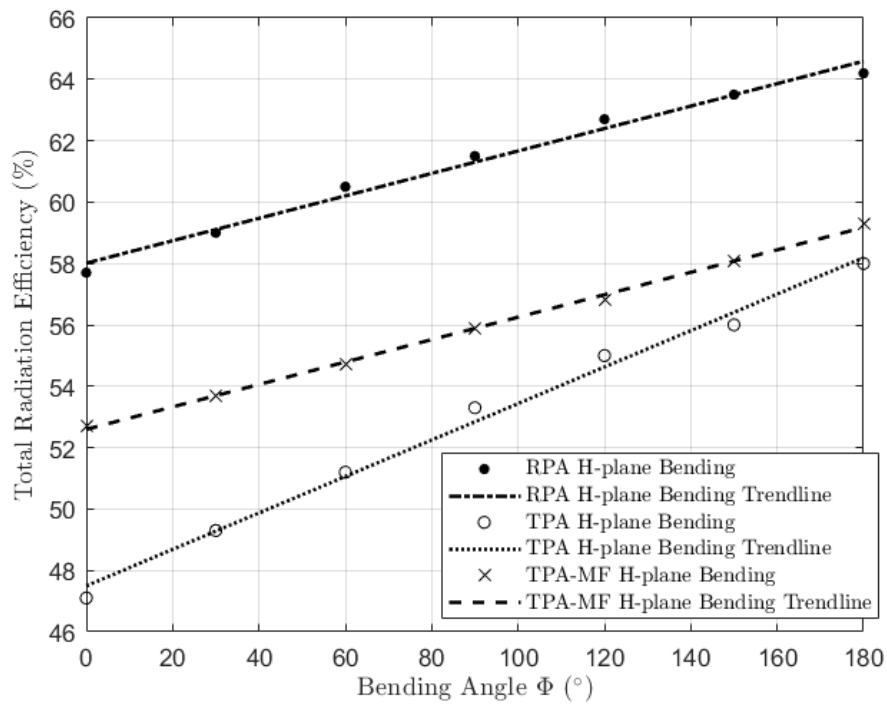


Figure 6.40: Total Radiation Efficiency comparison between the RPA, TPA and TPA-MF while bending along the H-plane.

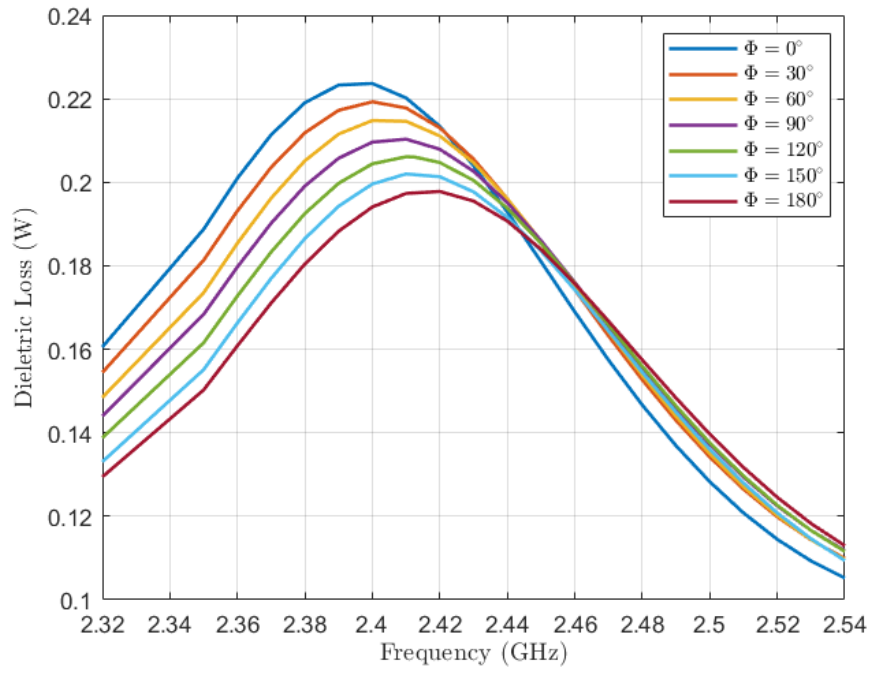


Figure 6.41: Dielectric losses for the various TPA-MF H-plane bending angles Φ .

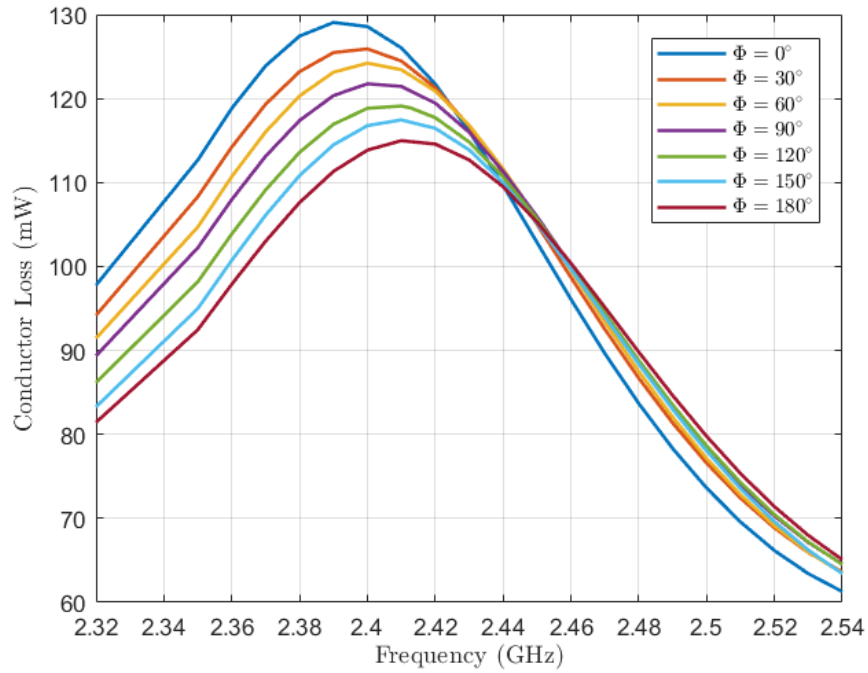


Figure 6.42: Conductor losses for the various TPA-MF bending angles Φ .

Furthermore, increasing bending decreases both the conductor and dielectric losses, which in turn increases the total radiation efficiency as shown in Figure 6.40. Typically, the total radiation efficiency would be expected to decrease caused by increasing mismatch losses due to bending over the microstrip line which changes the input impedance from 50 Ω .

However, the total radiation efficiency is shown to increase due to a reduction in conductor and dielectric losses.

Conductor and dielectric losses decrease during antenna H-plane bending because less power is transferred to the antenna due to increased input impedance mismatch, and due to the fringing fields that contribute to the overall radiation pattern are less exposed to the dielectric substrate, AMC cells and the ground plane. The ground plane is not performing as an adequate electromagnetic shield to its planar configuration. This is highlighted from the fact that the back lobe radiation increases as the bending angle increases as shown in Figure 6.35.

The power transmitted P^t into the conductive elements of the patch antenna is dissipated as heat within a very short distance from the surface. This is calculated as [35]

$$P^t = \frac{2|E_0|^2 R_s}{\eta_0^2} \quad (6.24)$$

Where η is the impedance of free space (Ω), E_0 is the peak electric field vector (V/m) and R_s is the surface resistance of the metal given as

$$R_s = \frac{1}{\sigma \delta_s} \quad (6.25)$$

Where σ is the conductivity (S/m) and δ_s is the skin depth (m). Increasing the bending angle marginally decreases the surface resistance of the metal since less of the electric field vector is exposed to it.

The dielectric loss of the antenna's substrate can be characterized by the loss tangent which is defined as

$$\tan \delta = \frac{\epsilon_r''}{\epsilon_r'} \quad (6.26)$$

where the real part of the permittivity ϵ_r' indicates how much energy from an external field is stored in the dielectric material and the imaginary part ϵ_r'' represents the losses within the dielectric in presence of an external field. The average power loss due to the dielectric P_d (W) can be expressed as [36]

$$P_d = \omega \epsilon_0 \epsilon'' |E_0|^2 V \quad (6.27)$$

where ω is the angular frequency (Hz), ε_0 is the permittivity of free space and V is the volume of the substrate (m^3). The dielectric power loss is given in Figure 6.41, which shows a gradual decrease in power loss for increasing H-plane bending angles. Overall, Figure 6.42 shows that the conductor loss reduction is small for increasing bending angles, while the dielectric loss reduction shown in Figure 6.41 is more significant reducing the losses by up to 0.2 W.

H-plane bending the TPA-MF is shown to detune the resonant frequency, however at its detuned frequency the reflection loss does not degrade when compared to the RPA and TPA for increasing bending angles. This is attributed to the width of the inset feed being smaller. The inset feed introduces a junction capacitance which is designed to tune the input impedance to 50Ω . A smaller inset feed gap results in the input impedance of the line to be less affected during H-plane bending.

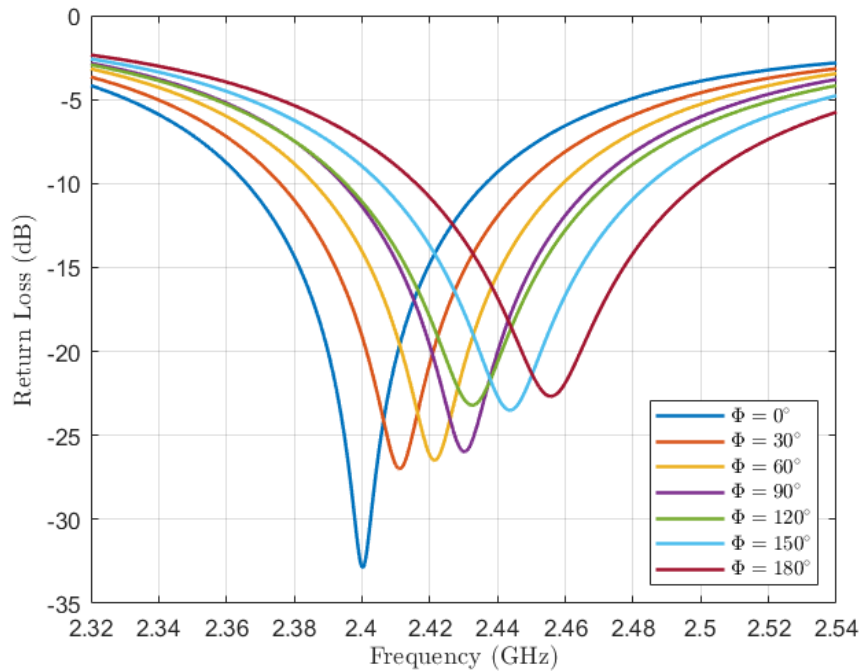


Figure 6.43: Simulated input reflection coefficient for E-plane bending of the TPA-MF.

Bending Angle θ (°)	f_{res} (GHz)	BW (MHz)	e_0 (%)	D (dBi)	G_{re} (dBi)
0	2.400	70.6	52.7	8.70	5.92
30	2.411	72.7	53.0	8.49	5.74
60	2.421	73.4	53.5	8.11	5.39
90	2.430	74.4	54.5	7.55	4.91
120	2.433	77.7	55.6	6.88	4.33
150	2.444	79.5	56.6	6.05	3.57
180	2.456	83.2	57.7	5.23	2.86

Table 6.7: Performance Parameters Comparison for YZ Plane Bending Angles.

Bending along the YZ or E-plane results in a larger shift in the resonant frequency when compared to H-plane bending. From its flat condition to a bending angle Φ of 180-degrees a total resonant frequency shift of 56 MHz occurs. The gradual shift in resonant frequency can be shown in Figure 6.44. Just like for H-plane bending, the bandwidth also increases for increasing bending angles like that of H-plane bending as shown in Figure 6.46. However, for bending angles $\Phi > 150^\circ$, the intended operational frequency of 2.4 GHz is no longer part of the antenna's bandwidth. In Chapter 5, for E-plane bending the RPA was also shown to completely detune its intended operating frequency's bandwidth for $\Phi > 150^\circ$. The TPA performed the best regarding E-plane bending with the bandwidth not including 2.4 GHz for a $\Phi = 150^\circ$. The comparison between the resonant frequency shifts for the RPA, TPA and TPA-MF is shown in Figure 6.44. The TPA-MF is shown to have the largest frequency detune for increasing E-plane bending angles.

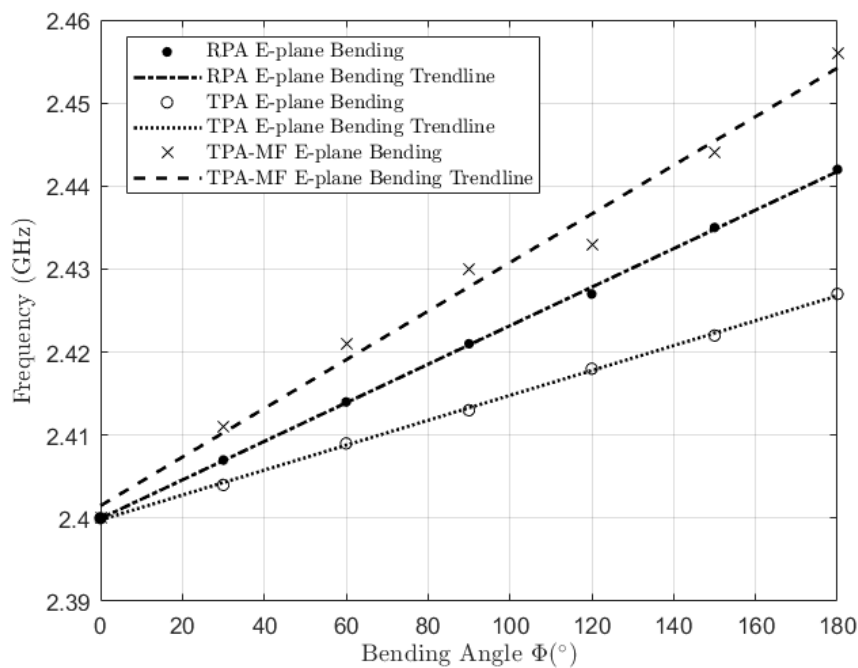
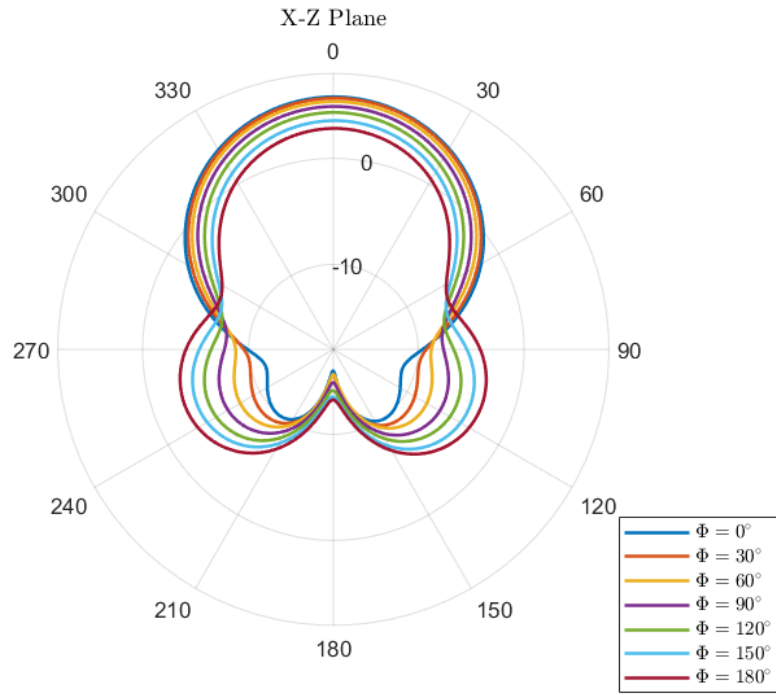


Figure 6.44: Resonant frequency shift comparison between the RPA, TPA and TPA-MF with respect to increasing bending angles across the E-plane.



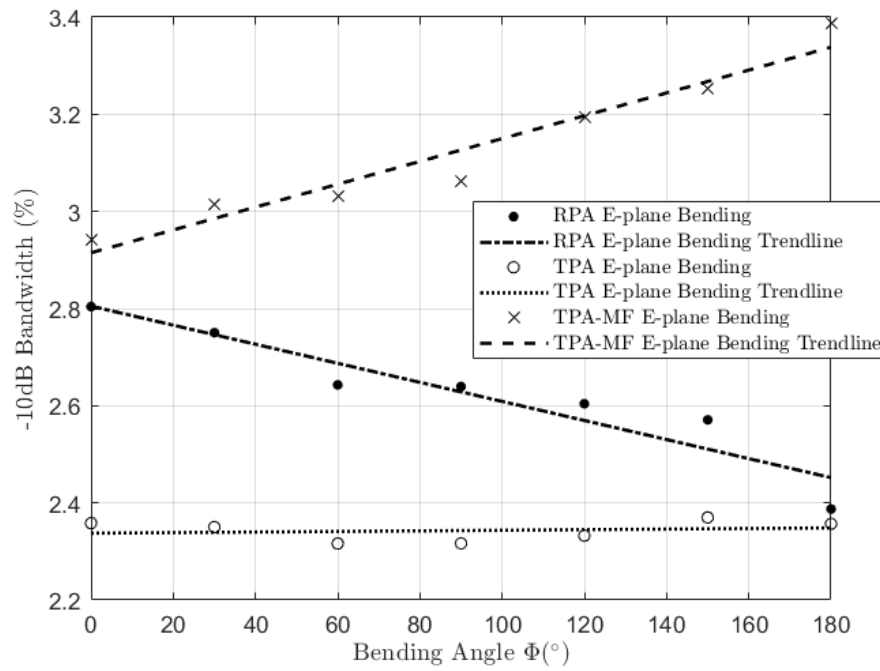


Figure 6.46: Bandwidth comparison between the RPA, TPA and TPA-MF while bending along the E-plane.

An observation can be made regarding the bandwidth increase for increasing bending angles for the TPA-MF in Figure 6.46. That is maximum bandwidth is not achieved at the resonant condition but rather with some mismatch with the antenna's input impedance. This bandwidth enhancement of up to 0.44% is at an expense of higher frequency detuning. Another way of saying this is the antenna's Q-factor reduces for increasing bending angles. Bending the RPA, which has the largest radiating element, deteriorates the bandwidth by 0.42% as the E-plane bending angle increases to 180 degrees, while there are marginal changes regarding the TPA's bandwidth which stays at a bandwidth of 2.3% for all bending angles investigated. The RPA's bandwidth is shown to be most affected due to input impedance mismatch for increasing E-plane bending.

Figure 6.47 shows the change in realized gain for the various bending angles. The realized gain is shown to decrease quite drastically when compared to H-plane bending with the realized gain reducing to 2.86 dBi for a 180-degree bending angle. The radiation efficiency increases just like for H-plane bending with similar power losses for both the dielectric and conductive material.

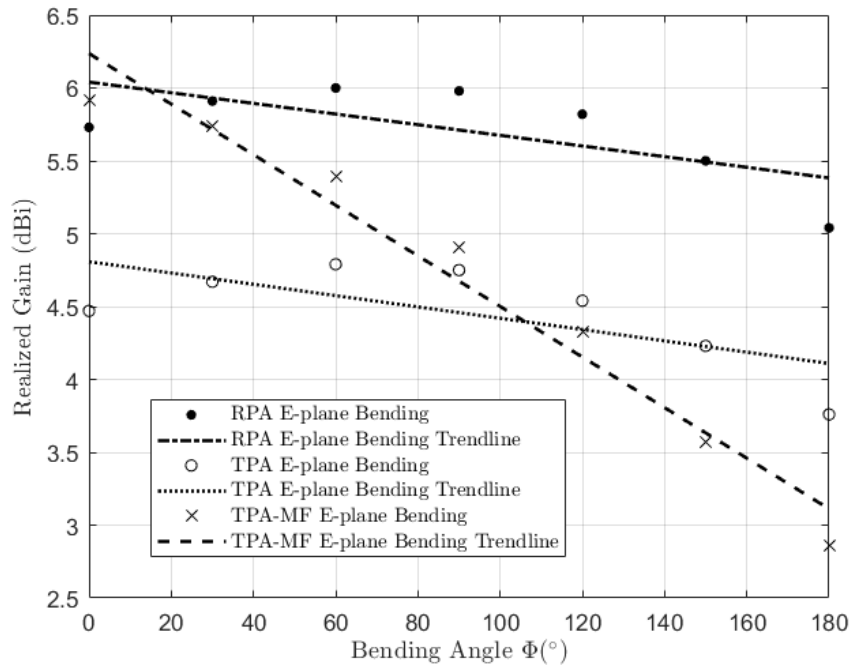


Figure 6.47: Simulated maximum realized gain comparison between the RPA, TPA and TPA-MF for various E-plane bending angles Φ .

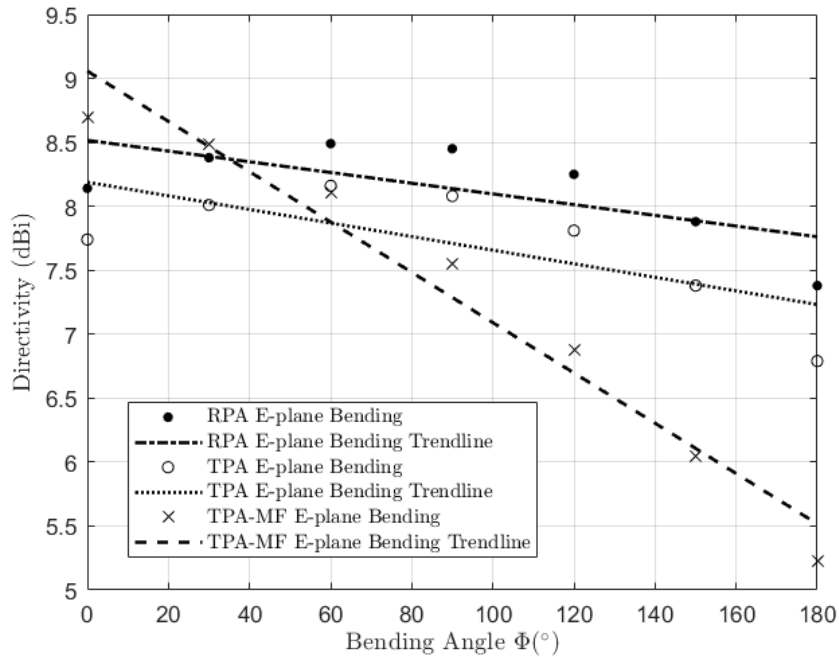


Figure 6.48: Simulated maximum directivity comparison between the RPA, TPA and TPA-MF for various E-plane bending angles Φ .

Whilst the TPA-MF has superior radiation performance compared to the RPA and TPA in its unbent form, E-plane bending results in rapid degradation of the antenna's directivity and realized gain. Figure 6.48 shows that the directivity of the TPA-MF for bending angles $\Phi >$

60° results in increasingly worse performance than the TPA. The realized gain in Figure 6.47 suffers a lot more for the TPA-MF for bending angles $\geq 120^\circ$ when compared to the TPA. The RPA is shown to outperform both the TPA-MF and TPA for large E-plane bending angles.

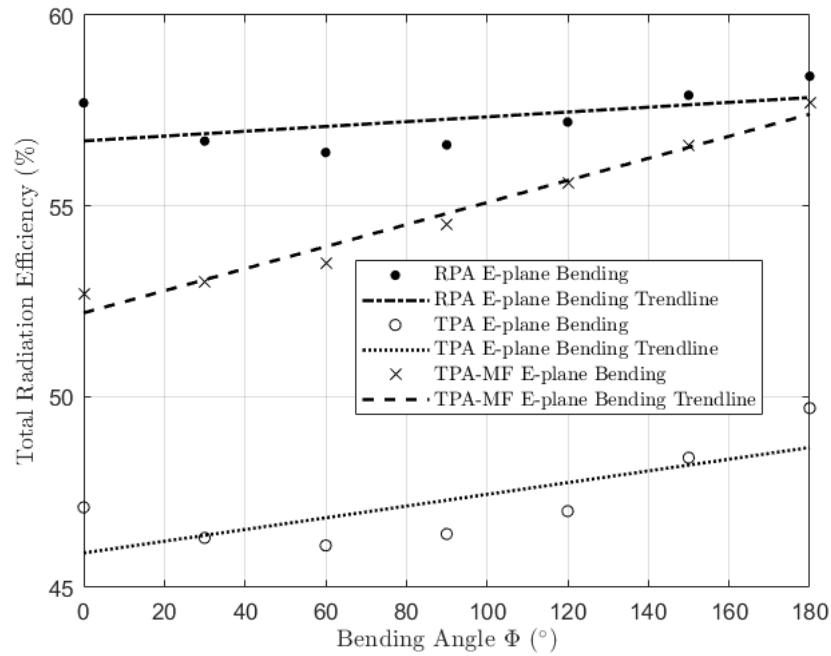


Figure 6.49: Simulated Total Radiation Efficiency comparison between the RPA, TPA and TPA-MF for various E-plane bending angles Φ .

The dielectric and conductor loss are shown in Figure 6.50 and Figure 6.51. For E-plane bending, the maximum magnitude for both the dielectric and conductor losses is similar to H-plane bending. The difference is that the maximum magnitude of these losses occurs at different frequencies since E-plane bending causes more resonant frequency detuning compared to H-plane bending. The explanation for the reduction in the conductor and dielectric is the same given for H-plane bending, which is that the fields that propagate along the sides of the patch are less exposed to the ground plane and the dielectric material as the bending angle increases. Also, as bending increases input impedance mismatch affects the amount of power transmitted to the antenna. If less power is transmitted to the antenna, the radiated electromagnetic fields will interact less strongly with the conductive and dielectric elements. These effects are highlighted by the larger back lobe radiation in Figure 6.35 and Figure 6.45.

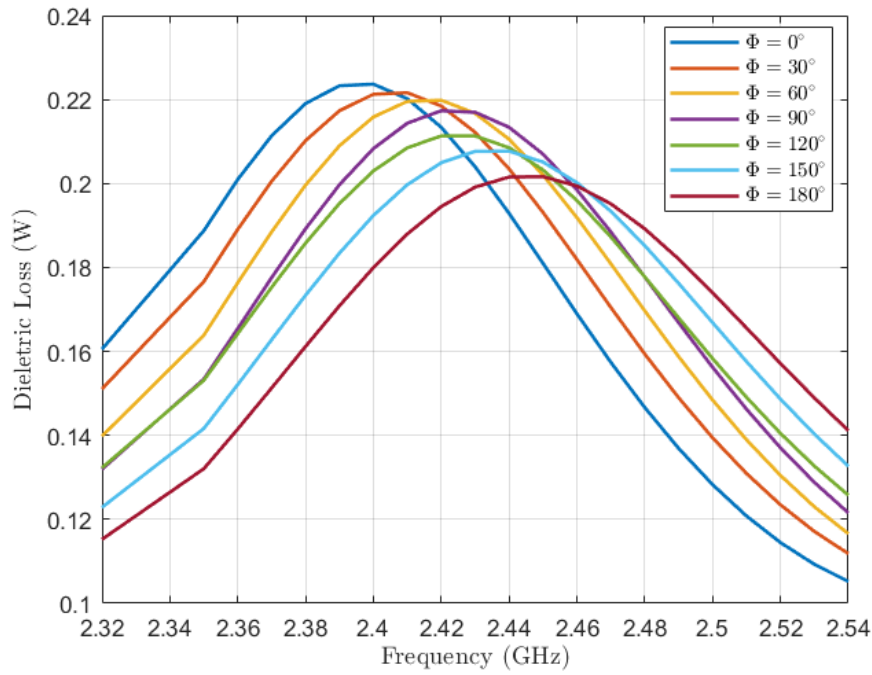


Figure 6.50: dielectric losses for the various bending angles Φ along the YZ-plane or E-plane.

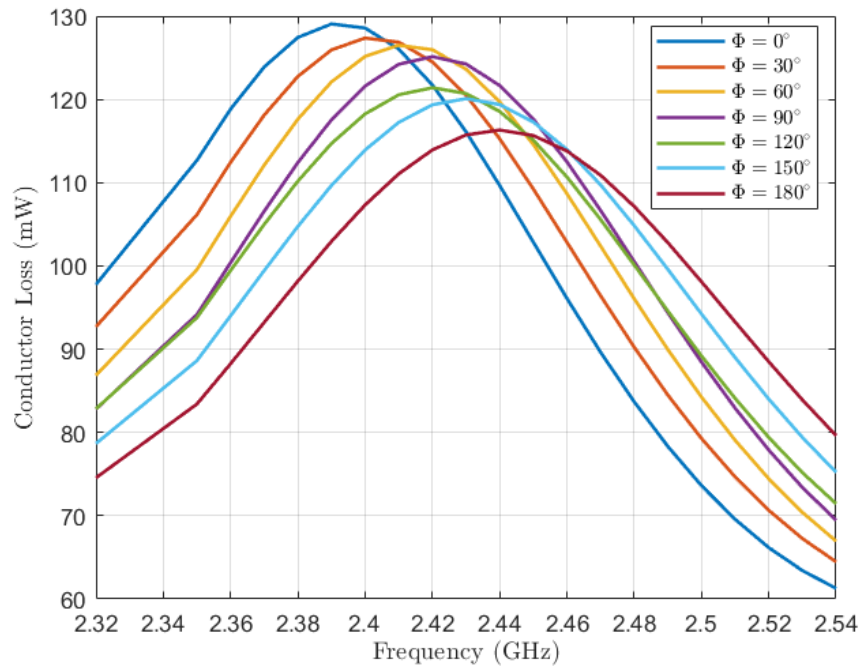


Figure 6.51: Conductor losses for the various bending angles Φ along the YZ-plane or E-plane.

6.6 Conclusion

Bending the TPA-MF affects the antenna's resonant frequency, bandwidth, radiation efficiency, directivity and realized gain. The TPA-MF results in a performance enhancement when compared to a TPA. There is an increase in bandwidth, directivity and realized gain. The

proposed design also reduces the radiating element's surface area by 26.3 % compared to the TPA.

The disadvantage of the TPA-MF is that resonant detune for both types of bending is greater than for a TPA. H-plane bending for a TPA from a 0-degree bend to a 180-degree bend shifts by 6 MHz, while E-plane bending shifts by 30 MHz. The TPA-MF has a resonant detune of 21 MHz for H-plane bending and 56 MHz for E-plane bending.

For both the TPA and TPA-MF bent along the H-plane, the -10 dB bandwidth is within that of the intended operational frequency of 2.4 GHz for all bending angles up to 180-degrees. For E-plane bending, the TPA for a 180-degree bend has completely detuned in which the -10 dB bandwidth does not contain 2.4 GHz. The TPA-MF for bending angles of 150 and 180-degrees have -10 dB bandwidths that do not contain 2.4 GHz. Therefore, while the TPA-MF has an increased bandwidth, resonant frequency detuning is affected to a slightly higher degree when compared to the RPA and TPA for E-plane bending. Thus, orientation of the TPA-MF is an important factor to consider. Bending the TPA-MF across its H-plane is shown to be resilient to radiation performance degradation.

6.7 References

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7 Simulation of the Wearable Trapezoidal Patch Antenna on the Human Body

Placing Antennas in proximity to lossy tissue raises challenges which include propagation losses, resonant detuning, fading and shadowing effects further influenced by variations in body posture and antenna positioning [1].

Since antenna measurements cannot be made on the human body directly due to uncertainty in exposure safety, measurements are made on physical phantoms, tissue probes and laboratory animals. Building a physical phantom body model containing liquids that mimic human tissue is an expensive and difficult method for radiofrequency dosimetry analysis. Physical models of body parts include the Specific Anthropomorphic Mannequin (SAM) head phantom which models the human head as a plastic shell filled with liquid mimicking the tissue within, and CTIA hand models which are posable for a variety of grips.

Due to the increased computational capabilities of modern hardware as well as efficient numerical techniques, computational human models offer an inexpensive alternative and can be simulated with submillimetre resolutions. The advancement in numerical dosimetry have led to it being an accepted method for specific absorption rate (SAR) compliance testing, provided ‘the computation properly captures everything that is required by the relevant exposure standard’ [2]. It is an ideal choice for situations where experimental SAR assessments is extremely complex [3].

Computational body models range from simple block models to complex voxel models which capture all the various tissues within the human body. Simple cubic or cylindrical models are layered with the various types of tissue such as skin, fat, muscle, and bone. A homogenous body model consists of a single layer in which the dielectric properties are averaged. A non-homogenous body model consists of discretized tissue each having its own dielectric properties [4].

CST offers a variety of bio models, from full body poseable homogenous phantoms to a variety of detailed high-resolution models such as voxel body models and FEM human models. The voxel body models are available in a range of ages, sizes, weight, number of tissues and resolutions of the voxels. During the import process the number of tissues can be customized to simplify the structure.

This chapter briefly introduces how the dielectric properties of human body tissue are realised in CST Microwave Studio followed by a short introduction to the specific absorption rate (SAR) and the legal limits for it. The main section of this chapter presents the investigation of the effect of body proximity on the proposed antennas performance and the SAR for different configurations. The configurations include a Cubic Tissue Model, Cylindrical Tissue Model, Homogenous Phantom Model, and the FEM Human Model. The tissue models are in order of increasing geometric complexity to represent the human arm.

7.1 Dielectric Properties of Human Body Tissue

Dielectric properties of human tissue vary with tissue type, frequency, and temperature. The non-frequency dependant dielectric properties of human tissue can be characterized by

$$\varepsilon_r = \frac{\varepsilon}{\varepsilon_0} = \varepsilon'_r - j\varepsilon''_r = \varepsilon'_r - \frac{j\sigma}{\varepsilon_0\omega} \quad (7.1)$$

where the real part of the permittivity ε'_r indicates how much energy from an external field is stored in the dielectric material and the imaginary part ε''_r represents the losses within the dielectric in presence of an external field. Conductivity σ has units of (S/m) while the angular frequency is in (rads/s).

Dielectric loss is defined by the dissipation factor $\tan \delta$ where

$$\tan \delta = \frac{\varepsilon''_r}{\varepsilon'_r} \quad (7.2)$$

As the frequency changes various polarization mechanisms occur. Molecular polarization or relaxations are associated with the translation, rotational, and vibratory modes of molecules within the medium. The field of dielectric spectroscopy is focused on observing these molecular relaxations. In the context of human tissue, the dielectric spectrum is characterized by three main relaxation regions, α , β and γ dispersion [5]. α dispersion occurs at frequencies less than 100kHz [6] and is associated with ionic diffusion at the site of the cellular membrane [7]. β dispersion occurs between 100 kHz ~ 500 MHz [8] and is due to the polarization of cellular membranes that act as barriers to the flow of ions between intra and extra cellular media [7]. γ dispersion occurs at the gigahertz range which will be the focus in this research. It occurs due to the polarization of water molecules [7].

The frequency dependence for tissue within each of the polarization or dispersion regions is expressed by the Cole-Cole equation

$$\varepsilon(\omega) = \varepsilon_{\infty} + \sum_{n=1} \frac{\Delta\varepsilon_n}{1 + (j\omega\tau_n)^{(1-\alpha)}} + \frac{\sigma_i}{j\omega\varepsilon_0} \quad (7.3)$$

where σ_i is the ionic conductivity, τ_n is the time constant or relaxation time, $\Delta\varepsilon_n$ is the magnitude of dispersion and α is an introduced term to describe the broadening of the dispersion [7].

Fitting the Cole-Cole equation to experimental data is shown to be accurate when compared to measured data. For frequencies in the γ dispersion region the predicted frequencies can be used with confidence [7]. Experimental data on dielectric properties of human tissue by C. Gabriel can be found in [9]. The dielectric property definitions for biological material in CST for a given frequency is based on the Cole-Cole equation and the measured results found in [9].

7.2 Specific Absorption Rate (SAR) and Recommended Human Exposure Limits

For human exposure to frequencies between 0 Hz to 300 GHz there are three biological effects to consider. Nerve stimulation, changes in permeability of membranes and temperature rise. At low frequencies between 0 Hz to 10 MHz, nerve and muscle stimulation effects can include induced ventricular fibrillation and inhibited breathing [10]. Furthermore, electromagnetic pulsed waveforms at low frequencies (up to 5 MHz [11]) can alter cell membrane permeability, which can lead to permanent damage to cells [12]. At higher frequencies between 100 kHz and 6 GHz heat is generated when the electromagnetic waves are exposed to human tissue.

Before talking about the effects of the induced temperature rise, a distinction must be made between a steady-state temperature rise and a brief temperature rise. Assuming an electromagnetic wave between 100 kHz and 6 GHz which is absorbed into the human body, a steady state temperature rise refers to heat dissipation over a large tissue mass and for thermoregulatory processes to counter the increase in temperature, while a brief temperature rise refers to a small region where there may not be enough time for the heat to dissipate. These small areas can result in a large temperature increase. A steady-state temperature increase of 1°C outside of the normothermic range ($37 \pm 1^\circ\text{C}$) of the body core

temperature can result in hyperthermia [13]. While localized temperature increase can cause tissue damage at temperatures greater than 41°C – 43°C. Above 6 GHz, heating predominantly occurs within the skin [13]. For a temperature rise due to exposure of an electromagnetic wave between 100 kHz and 6 GHz, the specific absorption rate (SAR) is the parameter of interest.

The SAR is defined as the “rate of incremental energy absorbed by an incremental mass contained in a volume element of a given density” [11].

The SAR can be expressed by

$$\text{SAR} = \frac{d}{dt} \left(\frac{dW}{\rho dV} \right) \quad (7.4)$$

and can be related to the root mean square (rms) of the E-field at a point by

$$\text{SAR} = \frac{\sigma |\mathbf{E}|^2}{2\rho} \quad (7.5)$$

The SAR has units in (W/Kg) where W is the energy (J), V is the volume (m^3), ρ is the mass density of the tissue (Kg/m^3), σ is the tissue conductivity (S/m) and $\mathbf{E}$ is the peak electric field vector (V/m) in the tissue.

The SAR can also be expressed in terms of a temperature increase at a point

$$\text{SAR} \approx \frac{c \Delta T}{\Delta t} \quad (7.6)$$

Where ΔT is the change in temperature ($^{\circ}\text{C}$), Δt is the duration of exposure (s) and c is the specific heat capacity ($\text{J}/\text{Kg}^{\circ}\text{C}$).

SAR depends on the properties of the incident electromagnetic field as well as the properties of the exposed body. For example, conductive elements and the dielectric substrate of a body-worn patch antenna cause attenuation to the radiated fields. The fields that propagate into human tissue due to edge diffraction at the ground plane are already attenuated and further attenuation occurs when they interact with human tissue. For simplicity, if the edge diffracted fields are modelled as plane waves, they have the form

$$\mathbf{E}(\mathbf{r}) = E_0 e^{j\mathbf{\tilde{k}}\mathbf{r}} \quad (7.7)$$

Where E_0 is the amplitude of the electric field, r is the distance from the observation point to the source point and $\tilde{k}$ is the complex propagation constant, given as

$$\tilde{k} = k - j\alpha \quad (7.8)$$

where α is the attenuation coefficient (Np/m) which is a measure of the spatial rate of decay of the wave in the medium [14] and k is the propagation constant which corresponds to the number of wave cycles per unit length.

Incorporating attenuation into (7.5) and assuming a linear, isotropic, homogenous, lossy medium, the SAR or absorbed power due to an attenuated plane wave can be expressed as

$$\text{SAR} = \frac{\sigma E_0^2}{2\rho} e^{-2\alpha r} \quad (7.9)$$

The attenuation coefficient α captures all the loss parameters and can be expressed in terms of the waveforms frequency and the medium's dielectric and conductive properties as [14]

$$\alpha = \omega \sqrt{\frac{\mu\epsilon}{2} \left(\sqrt{1 + \left(\frac{\sigma}{\omega\epsilon}\right)^2} - 1 \right)} \quad (7.10)$$

This example assumes the non-frequency dependant dielectric properties of the medium.

To limit the human exposure to emitted radio/microwave frequencies various standards have been developed for personal protection. The standards provide guidance on the maximum levels of permissible electromagnetic radiation exposure to protect against an excessive temperature rise in any part of the body that might result from local or nonuniform radiation exposure [11]. Standards such as the International Commission on Non-Ionizing Radiation Protection (ICNIRP) or the Institute of Electrical and Electronic Engineers International Committee on Electromagnetic Safety (IEEE ICES) provide guidance on limiting radiation exposure. National authorities can also have their own standards depending on the country [1]. From the IEEE, for human electromagnetic exposure between frequencies 0 Hz to 300 GHz there are two standards which complement each other, first is the C95.1-2019 [11] which is a standard to provide a criterion to protect against exposure to electromagnetic fields. The second is the C95.3-2021 [2] standard which provides 'concepts, techniques and

instruments that can be applied to computation of the SAR or electric field strength in humans and phantom models.’ This standard also provides various computational techniques on human exposure to electromagnetic fields. For human exposure, the C95.1-2019 recommends compliance with the dosimetric reference level (DRL) or the exposure reference level (ERL). The metrics provided for the DRL is given by the specific absorption rate (SAR) (W/Kg) or the Epithelial power density (W/m^2) which is associated with surface heating.

The metrics associated with the ERL is the rms electric and magnetic field strengths, power density and plane wave equivalent power density (S_E, S_H). For frequencies between 100 kHz and 6 GHz, compliance with measuring the parameters associated with the ERL implies compliance with the DRLs. At higher frequencies between 6 GHz to 300 GHz, both the DRLs and ERLs should be evaluated to demonstrate compliance. Recommended ERL limits can be found in [11].

Another standard for exposure criteria is defined in the 2013 directive by the European parliament and the council of the European union [15] which was established to limit human exposure to radiation in the frequency range from 0Hz to 300GHz.

Human exposure guidelines are also defined by the International Commission on Non-ionizing Radiation Protection (ICNIRP) which is a non-governmental organization formally recognized by the World Health Organisation (WHO) [1].

A summary of the SAR exposure limits from the various standards is shown in Table 7.1 below. The restricted environments refer to where the general public are not permitted unless subject to an applicable safety program.

Standard	Exposure environment	Frequency Range	Whole Body Average SAR (Wkg^{-1})	Local Head/Torso SAR (Wkg^{-1})	Local Limbs SAR (Wkg^{-1})
IEEE C95.1-2019	Restricted	100 kHz – 6 GHz	0.4	10	20
IEEE C95.1-2019	General Public	100 kHz – 6 GHz	0.08	2	4
ICNIRP	Restricted	100 kHz – 300 GHz	0.4	10	20
ICNIRP	General Public	100 kHz – 300 GHz	0.08	2	4
DIRECTIVE 2013/35/EU	-	100 kHz – 6 GHz	0.4	10	20

Table 7.1: Summary of SAR exposure limits from various standards.

Hirata and Fujiwara [16] found that a 10g cubic mass provides a good representation of EMF induced temperature changes. Therefore, local SAR is averaged over a 10g cubic mass. For the IEEE C95.1-2019 and the ICNIRP the whole-body average SAR exposure is for over 30 minutes while local exposure is for averaged over 6 minutes. The DIRECTIVE 2013/35/EU

average SAR exposure for the whole-body and local exposure is averaged over 6 minutes. Overall, electromagnetic exposure of frequencies between 100 kHz to 300 GHz has been researched extensively.

While one of the aims of this research is to investigate how the radiation performance parameters of the proposed antennas are affected in the presence of lossy dielectric tissue, computing SAR is an essential parameter which must comply with exposure limits defined in Table 7.1.

7.3 Effect of Body Proximity on Antenna Performance and SAR

In this section the effect of body proximity and bending of the antenna around the human arm is investigated to show how it affects the antennas performance parameters and specific absorption rate (SAR).

Initially for simplicity, a cubic and cylindrical multilayer human tissue model of the upper arm of a human is used from [17], where the volumes for both models are equivalent. The density, thermal and dispersive properties for each layer of tissue are already defined in the CST BioModel library extension, their definitions are included in the appendix. When designing the cubic and cylindrical tissue model, each layer is given a material definition with skin, fat, muscle, and bone. They each have a respective thickness of 1.5 mm, 8.5 mm, 27.5 mm and 12.5 mm [17]. These dimensions are consistent with other tissue models reported in literature for the upper arm of a human [18]–[20]. Antenna bending is incorporated when simulating the antenna on the cylindrical multilayer tissue model, homogenous phantom model, and FEM human model.

The homogenous body model represents the human tissue as a single material averaging the properties of muscle and fat, while the voxel and FEM meshable model is more complex and realistic with material definitions for individual tissue within the body. The homogenous body model is imported using the Bio Extension macro. The macro gives a choice of various body model poses in which the arm down pose is chosen.

Since the voxel tissue model is not available for tetrahedral meshing, only the FEM meshable body model is used for simulations which involves the frequency domain solver. These models represent the most detailed tissue body models available for tetrahedral meshing. CST Microwave Studio offers two models ‘Nelly’ and ‘Hugo’ which are optimized for tetrahedral meshing. Nelly and Hugo are based on a female and male frozen corpse from the

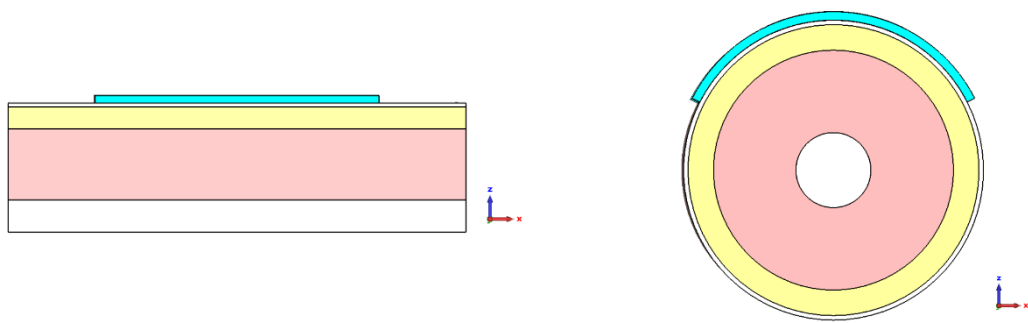
US National Library of Medicine's Visible Human Project which was dissected and photographed and converted to a 3D CAD model. Nelly is 59 years old with a body mass of 86 kg, while Hugo is a 38 years old, with a mass of 103 kg.

For large structures the frequency domain solvers memory requirements increase quadratically with the number of tetrahedron mesh cells. In Chapter 5 and 6 the frequency domain using a tetrahedral mesh was shown to simulate bending effects well. The transient solver using a hexahedral mesh requires an extremely dense mesh to capture the angular changes in the structure effectively. The Finite Difference Time Domain (FDTD) solver also known as the Transient Solver is known to be suited for electrically large simulations and detail rich structures [22]. The solver also allows the unique feature of being able to simulate the antennas with voxel body models. However due to not being able capture small changes in antenna bending effectively, the Transient Solver is not used. Since the memory requirements of the frequency domain will be high for a full body simulation and will exceed the available memory (RAM) during the simulation, adjustments to the bounding box of the simulation are made. For the Homogenous Phantom Model and FEM Tetrahedral Meshable Body Model the bounding box for the simulation is extended by an extra 10 cm in all directions from the antenna which captures the simulation of the upper arm, part of the torso and part of the lower arm. Each of the tissue model configurations are shown in Figure 7.1.

The diameter of the cylindrical tissue model, upper arm of the homogenous phantom, upper arm of the voxel model and upper arm of the FEM meshable body model are respectively 100 mm, 90 mm, 120 mm and 140 mm. The arm of the homogenous phantom, voxel model and FEM meshable body model does not form perfect cylinders. The antenna is bent around the arm and is placed as closely as possible to the arm without intersecting the antenna into the tissue. A 110° bending angle is applied across the cylindrical tissue model and homogenous phantom since the bending angle results in the antenna conforming to the surface of the upper arm well. The FEM meshable body model has a larger upper arm diameter of 140 mm and thus a bending angle of 90° is used. Bending along the E and H-plane is investigated for each of these models.

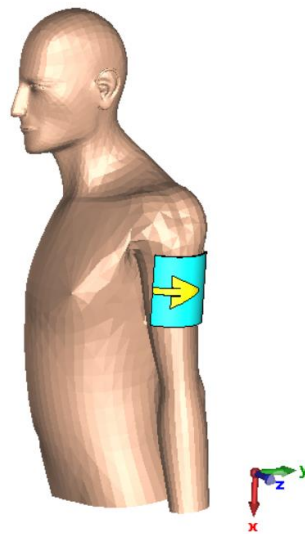
The antennas performance parameters are compared with the various body model configurations. Since one of the aims is to investigate the antenna's suitability for wearable applications, the specific absorption rate (SAR) is a crucial parameter that must be investigated to ensure that there are no health risks to the human body. SAR is expected to

be low and well within the standards recommendations as the patch antenna configurations contain a relatively large ground plane. The SAR calculations are performed using the IEEE/IEC 62704-1 standard averaging method [3] provided in the SAR calculation tool in CST Microwave Studio.

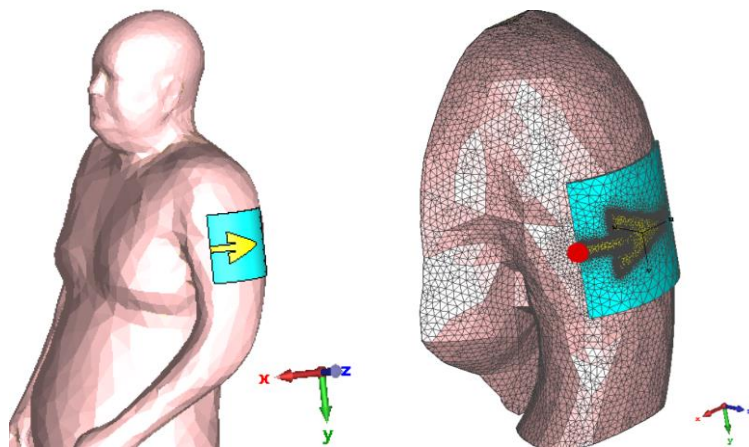


(a) Cubic Tissue Model

(b) Cylindrical Tissue Model



(c) Homogenous Phantom Model



(d) FEM Tetrahedral Meshable CAD Body Model 'Nelly'

Figure 7.1: Body model configurations to test the various antennas performance in the presence of human tissue.

The antenna simulations without any tissue implemented is labelled ‘free space’, which is included as a benchmark to compare how exposure to human tissue and antenna bending at the same time affects the radiation performance parameters.

For full wave simulations performed using the frequency domain solver, the TPA simulated on the cubic tissue model and compared to the TPA simulated in free space. The performance parameter results are shown in Table 7.2. The results show a decrease in the antenna’s bandwidth by 0.45% compared to the TPA in free space. The introduction of lossy tissue results in a radiation efficiency reduction by 6.2%. However, there is a performance enhancement regarding the antenna’s directivity and realized gain. The directivity increases by 0.7 dBi and the realized gain increases by 0.15 dBi.

Antenna Type	Tissue Model Configuration	f_{res} (GHz)	Bandwidth %	Directivity (dBi)	Total Radiation Efficiency %	Realized Gain (dBi)	SAR 10g (W/Kg)
TPA	Free Space	2.400	2.30	7.74	47.1	4.47	-
TPA	Cubic	2.399	1.85	8.44	40.9	4.57	0.135
TPA-MF	Free Space	2.400	2.94	8.70	54.0	5.92	-
TPA-MF	Cubic	2.393	3.07	9.72	48.3	6.56	0.105

Table 7.2: Comparison between the TPA and TPA-MF simulated in free space and placed on the cubic tissue model.

Simulating the TPA-MF on the cubic tissue model displays a performance enhancement in all the radiation performance parameters compared to the TPA-MF simulated in free space, including the antenna’s bandwidth which was not seen for the TPA on the cubic tissue model.

Antenna Type	Tissue Model Configuration	Mesh Cells (Tetrahedrons)	Peak Memory Used (GB)	Total Simulation Time (s)
TPA	Free Space	88778	1.181	86
TPA	Cubic	362072	15.67	892
TPA-MF	Free Space	187056	9.17	317
TPA-MF	Cubic	202718	9.39	726

Table 7.3: Comparison of the computational requirements using the frequency domain solver for the TPA and TPA-MF simulated in free space and placed on the cubic tissue model.

Table 7.3 shows the computational requirements comparing the proposed TPA and TPA-MF simulated in free space and placed on the cubic tissue model. A limiting factor in the antenna simulations with human tissue models is the memory requirements. The frequency domain is a finite elements type solver with numerical requirements which increase exponentially with larger simulation models. The peak memory requirements encompass the frequency domain solver’s equation system set up, system solver and mesh refinement over the tetrahedral

meshing of the model. The maximum amount of mesh cells for a simulation is determined by the level of mesh refinement the model requires which is set by the threshold convergence value for the S-parameters. The TPA and TPA-MF in free space and placed on the cubic tissue model have short simulation times with the maximum taking approximately 15 minutes, and moderate memory (RAM) requirements with the maximum peaking just under 16 GB of RAM.

Bending is now incorporated into the analysis and a comparison is made between the proposed TPA and TPA-MF in free space and placed on the FEM Human Model. The simulations are performed for E and H-plane bending where the results are shown in Table 7.4 and Table 7.5.

Antenna Type	Bending Angle (°)	Tissue Model Configuration	f_{res} (GHz)	Bandwidth (%)	Directivity (dBi)	Total Radiation Efficiency (%)	Realized Gain (dBi)	SAR 10g (W/Kg)
TPA	90	Free Space	2.413	2.31	8.08	46.4	4.75	-
TPA	90	FEM Human	2.414	1.73	8.31	42.0	4.53	0.335
TPA-MF	90	Free Space	2.430	3.06	7.55	54.5	4.91	-
TPA-MF	90	FEM Human	2.425	2.96	8.49	48.5	5.35	0.526

Table 7.4: E-plane bending comparison between the TPA and TPA-MF simulated in free space and placed on the FEM Human Model.

In Chapter 5 and 6, frequency detuning was shown to be more significant for E-plane bending. Simulating E-plane antenna bending with the FEM Human tissue model shows the same amount of resonant frequency detuning as the free space configuration for a 90° bend. For the TPA the radiation efficiency is reduced by 4.4% and the bandwidth is reduced by 0.58%. However, there is a marginal increase in the TPA's directivity by 0.23 dBi.

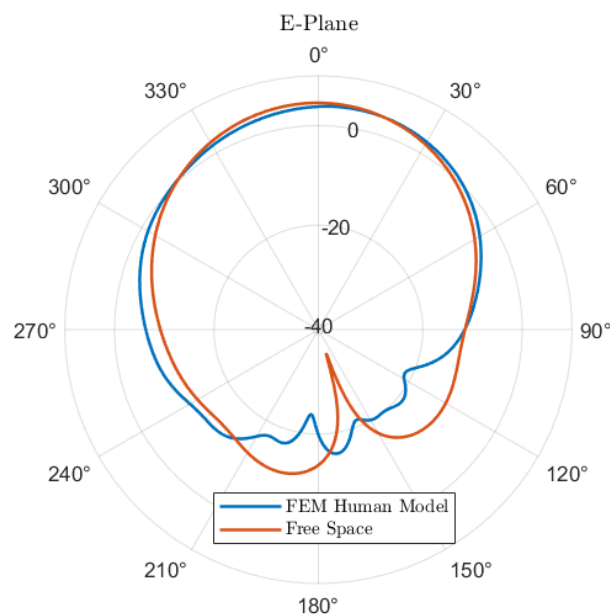


Figure 7.2: E-plane far-field cut in terms of realized gain for TPA while bending along its H-plane.

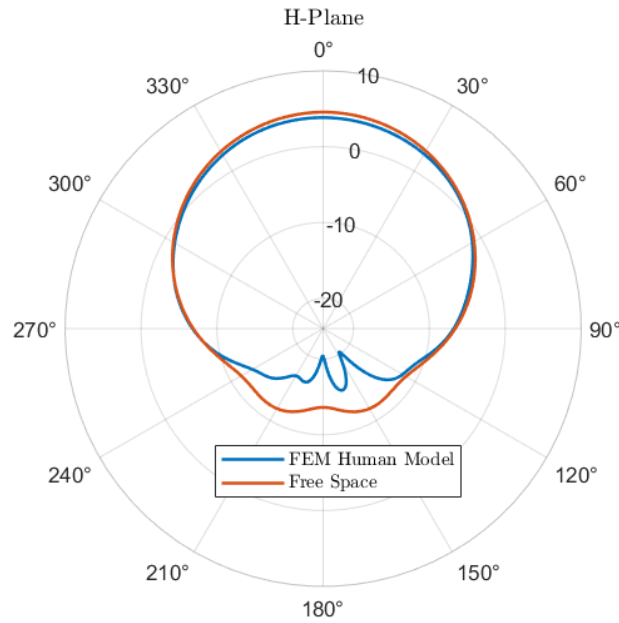


Figure 7.3: H-plane far-field cut in terms of realized gain for TPA while bending along its H-plane.

A comparison of the far-field cuts for the TPA simulated in free space and with the FEM Human Model is shown in Figure 7.2 and Figure 7.3 for the E-plane and H-plane cuts. For the TPA on the FEM Human Model, the E and H-plane is shown to have smaller back lobes compared to the TPA in free space. The magnitude at the major lobes also decreases for both cuts. The E-plane cut shows broadening of the major lobe for the FEM Human Model, while the H-plane does not.

The TPA-MF shows a small reduction in the antenna's bandwidth of 0.1%. The radiation efficiency shows to reduce by 6% which is more than for the TPA. The TPA-MF shows an enhancement to the antenna's directivity by 0.94 dBi. The realized gain is marginally enhanced by 0.44 dBi since the increase in the antenna's directivity supersedes the decrease in radiation efficiency.

Antenna Type	Bending Angle (°)	Tissue Model Configuration	f_{res} (GHz)	Bandwidth %	Directivity (dBi)	Total Radiation Efficiency %	Realized Gain (dBi)	SAR 10g (W/Kg)
TPA	90	Free Space	2.402	2.45	6.63	53.3	3.90	-
TPA	90	FEM Human	2.402	1.95	6.98	48.2	3.83	0.0721
TPA-MF	90	Free Space	2.411	3.16	8.15	55.9	5.63	-
TPA-MF	90	FEM Human	2.410	3.19	8.44	55.9	5.91	0.0385

Table 7.5: H-plane bending comparison between the TPA and TPA-MF simulated in free space and placed on the FEM Human Model.

H-plane bending the TPA placed on the FEM Human Model is shown to cause no changes to the resonant frequency and very small enhancement to the directivity. Due to a reduction in the TPA's radiation efficiency by 5.1%, the radiation efficiency's reduction supersedes the

increase in the antenna's directivity and thus this time the realized gain reduces. The TPA's bandwidth reduces by 0.5%.

Despite the increased resonant frequency detuning for the TPA-MF compared to the TPA, the TPA-MF is shown to maintain its radiation performance for all the parameters. There is even a small improvement in the TPA-MF's directivity and realized gain.

A comparison of the far-field cuts for H-plane bending the TPA-MF in free space and placed on the FEM Human Model is shown in Figure 7.4 and Figure 7.5. For both cuts at the major lobe, the magnitudes are almost equivalent. For the TPA-MF placed on the FEM Human Model, the H-plane cut shows a reduction in magnitude at the back lobe for the antenna placed on the FEM Human Model. It can also be seen that ripples are introduced at the back lobe due to the interference between the radiation from the back lobe and the reflected waveforms from the ground plane and human tissue. Also, the major lobe becomes marginally broader in the E-plane cut.

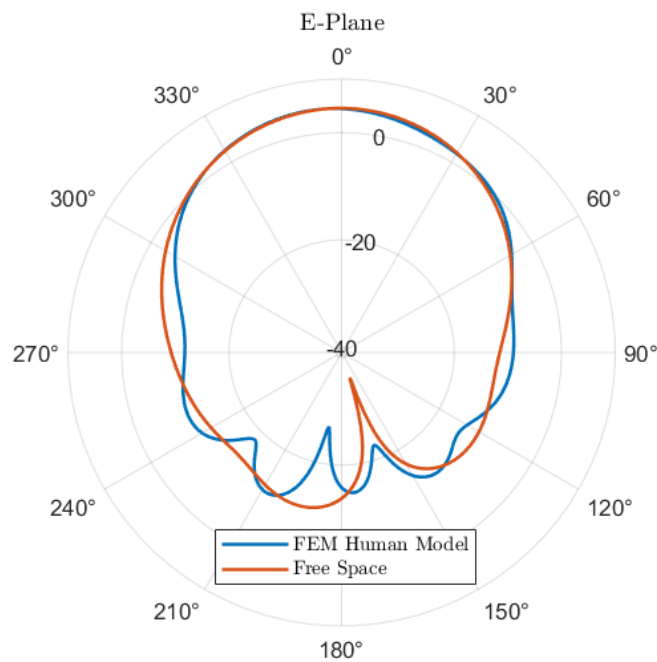


Figure 7.4: E-plane far-field cut in terms of realized gain for TPA-MF while bending along its E-plane.

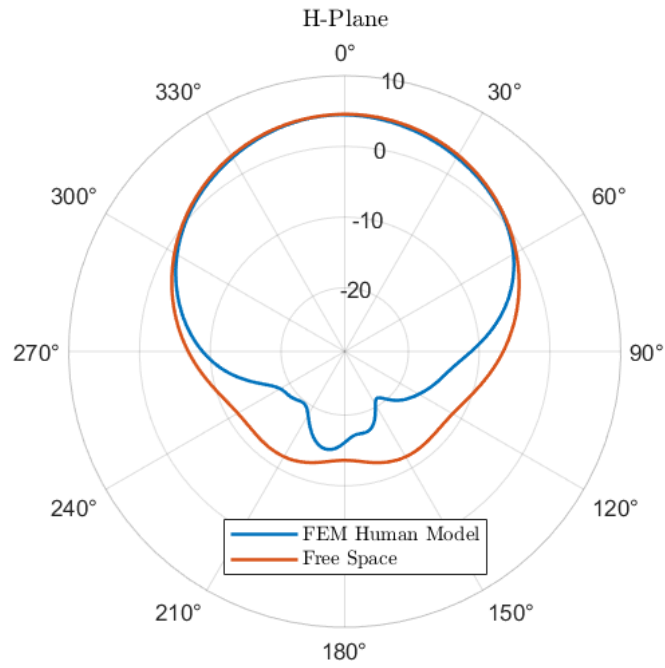


Figure 7.5: H-plane far-field cut in terms of realized gain for TPA-MF while bending along its E-plane.

Antenna Type	Bending Type	Bending Angle (°)	Tissue Model Configuration	Mesh Cells (Tetrahedrons)	Peak Memory Used (GB)	Total Simulation Time (s)
TPA	E-Plane	90	Free Space	85001	3.89	253
TPA	E-Plane	90	FEM Human	900378	19.63	4237
TPA	H-Plane	90	Free Space	85609	1.41	208
TPA	H-Plane	90	FEM Human	903605	20.16	4407
TPA-MF	E-Plane	90	Free Space	942538	8.10	609
TPA-MF	E-Plane	90	FEM Human	512736	42.53	4463
TPA-MF	H-Plane	90	Free Space	182558	9.27	414
TPA-MF	H-Plane	90	FEM Human	510415	44.82	5555

Table 7.6: Comparison of the computational requirements using the frequency domain solver for bending the TPA and TPA-MF in free space and placed on the FEM Human Model.

Adding the FEM Human Model drastically increases the computational effort requiring over 40 GB of RAM and taking a significantly longer time to simulate. The longest simulation time is recorded for the TPA-MF H-plane bent over the FEM Human Model which lasted just under 2 hours. The largest amount of mesh cells belongs to both E and H-plane bending the TPA over the FEM Human Model. Convergence of the S-parameters was shown not to be found early in the mesh refinement process.

The various simulated radiation performance parameters for the TPA and TPA-MF for the other tissue model configurations such as the cylindrical and homogenous phantom

models for E and H-plane bending are given in Table C.1 and Table C.2 whereby similar trends are seen.

SAR is computed for each tissue model for the proposed TPA and TPA-MF. The results are shown in Table C.1 and Table C.2. SAR is shown to be low and complies well within the recommended limits for all antenna model configurations. The result is expected due to the antenna consisting of a large ground plane acting as an EM shield against backpropagating radiation. There is shown to be some variance in the SAR results between the models for both E and H-plane bending. For the cylindrical tissue model, SAR for both E and H-plane bending is shown to be almost equivalent. For the homogenous phantom model, SAR is shown to be slightly larger by 0.2 W/Kg, for H-plane bending, while the FEM human model shows the opposite behaviour with SAR being higher for E-plane bending. Differences in SAR can be attributed to how the antenna is placed on a cylindrical surface which is non uniform. The air gaps between the skin tissue and the ground plane vary and thus SAR will also vary based on the separation distance. SAR for the TPA-MF simulated on the FEM human model is shown to be highest at a value of 0.526 W/Kg when E-plane bending the antenna and lowest for H-plane bending with a value of 0.0385 W/Kg.

7.4 Conclusion

Overall, introducing human tissue models and bending effects to the proposed TPA and TPA-MF show small changes to the antenna's performance parameters. Introducing the human tissue beneath the antennas shows an improvement in the antenna's directivity compared to its free space counterpart.

Chapter 5 and 6 found that E and H-plane bending results in a performance degradation in different ways. For E-plane bending, the resonant frequency of the antenna was shown to increase more than for H-plane bending, while H-plane bending shows to cause impedance mismatch due to bending across the feedline and a more significant degradation for the antennas realized gain than E-plane bending. Therefore, the orientation of the antenna placed on a curved surface on the human body is an important factor to consider.

In this chapter, incorporating cylindrical type tissue models with increasing geometric complexity to replicate a human arm shows marginal changes to the antenna's performance parameters. This can be attributed to the antennas consisting of a large ground plane.

The key results that have been identified in this analysis are:

- Frequency detuning occurred for both antennas simulated in the presence of the FEM human tissue model, however, the TPA-MF has the largest frequency detune shifting upwards of up to 2.43 GHz for a 90-degree E-plane bending angle. Despite this shift the operational bandwidth is large enough to operate at the intended resonant frequency of 2.4 GHz.
- H-plane bending is shown to reduce the TPA's directivity by 1.33 dBi compared to E-plane bending when in the presence of the FEM human tissue model.
- The proposed TPA-MF simulated on the FEM human tissue model is shown to maintain its directivity performance of 8.4 dBi for both E and H-plane bending operations.
- A bandwidth of 3% is maintained for both E and H-plane bending the TPA-MF in free-space and placed on the FEM human tissue model.
- Introducing the FEM human body model results in a reduction in the total radiation efficiency, which is shown to be more drastic for E-plane bending the TPA, reducing from 46.4% to 42%
- Total radiation efficiency is shown to be superior for the TPA-MF than the TPA simulated in free space and in the presence of tissue models.
- E-plane bending the TPA-MF in the presence of the FEM Human model results in a total radiation efficiency reduction by 6% however it remains being a more efficient radiator compared to the TPA by 6.5%.

It was also found that a detailed tissue model such as the FEM Human Model can be simulated without the use of a high-performance workstation, provided that the simulation is confined to the area of interest around the antenna. The area of interest in this analysis was located on the arm of the FEM Human Model and thus mesh refinement processes on the antennas can be performed to achieve a geometrically accurate simulation of the antenna without exceeding memory (RAM) capability.

7.5 References

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8 Conclusions

This thesis has addressed the challenge of miniaturizing and optimizing the patch antenna for suitability in wearable applications, where the antenna can still perform optimally under H and E-plane bending deformation. The Trapezoidal Patch Antenna (TPA) and the Trapezoidal Patch Antenna with an AMC Metafilm (TPA-MF) are proposed as edge and inset fed miniaturized variants of the microstrip patch antenna and are compared to the traditional Rectangular Patch Antenna (RPA).

This chapter provides a chapter-by-chapter summary of the thesis followed by suggestions for future work and a final reflection.

8.1 Summary

8.1.1 Chapter 4

The TPA has been previously presented as a miniaturization technique however, there have been no analytical methods presented giving insight into its radiation characteristics. An approximate analytical analysis based on the Aperture Array Model (AAM) is presented for the TPA. A far-field normalized power pattern plot can be computed analytically, while the radiated power and directivity of the antenna is evaluated numerically. The aperture array model shows that directivity decreases if the size of a radiating edge decreases.

A full-wave analysis shows that transforming the radiating element of a rectangular patch to a trapezoidal patch can result in a surface area reduction by up to 38.9 % while operating at the same frequency.

The size reduction of the radiating element's aperture is shown to be a design consideration for low-profile patch antenna applications. The decrease in the radiating element's size supersedes the marginal decrease in the antenna's radiation performance. The reduction in performance of the antenna's directivity is less than what the analytical aperture array analysis predicts.

8.1.2 Chapter 5

The Cylindrical Aperture Array Model (CAAM) is used to analyse the radiation characteristics of the proposed trapezoidal patch antenna (TPA). The CAAM shows that patch antenna bending degrades the directivity of the antenna while increasing back lobe radiation and broadening the side lobes.

Using the Cylindrical-Rectangular Cavity Model the resonant frequency detune for a rectangular patch antenna can be investigated for both E and H-plane bending. The model represents the cross-section of the patch as a sector of an annulus between two concentric circular cylinders of radius R_0 and h . The results show that bending along the E-plane of the antenna increases the resonant frequency linearly from 2.4 GHz for a 0° bending angle to 2.73 GHz for a 240° bending angle. Bending along the H-plane has no effect on the resonant frequency. This model assumes a ground plane that extends the width and length of the aperture.

A full wave simulation analysis is performed on the TPA which is compared to the insights gained from the analytical models. The definition of antenna bending for the full wave simulation analysis is now changed to account for the antenna's ground plane which is larger than the radiating element. E-plane bending is shown to increase the antennas resonant frequency the most. However, the proposed TPA results in a resonant frequency detune which is 15 MHz less than the RPA for E-plane bending and 4 MHz less than the RPA for H-plane bending. The simulated farfield pattern is plotted in terms of realized gain. The farfield pattern is consistent with the power pattern plotted using the CAAM, showing a decrease in magnitude at the main lobe, increasing back lobe and broadening side lobes.

H-plane bending for a bending angle between 0° and 180° reduces the realized gain linearly for the TPA from 4.47 dBi to 3.49 dBi. While for E-plane bending the realized gain increases for small bending angles up to a bending angle of 60° . For bending angles greater than 60° the realized gain starts degrading. Overall, E-plane bending does not degrade the antennas realized gain as much as H-plane bending.

8.1.3 Chapter 6

The Trapezoidal Patch Antenna loaded with an AMC MetaFilm (TPA-MF) is proposed as a performance enhancement method. Compared to the RPA, the TPA-MF results in a 55 % reduction in the radiating element's surface area, while maintaining the antenna's bandwidth and realized gain.

The Generalized Sheet Transition Conditions (GSTCs) is argued as the appropriate method of characterizing the metafilm's dielectric properties and thus a comparison is made between the effective medium model and the generalized sheet transition condition (GSTCs) method to characterize the AMC metafilm. There are differences in the computed

susceptibilities between the two characterization models which needs to be validated with measured results.

Bending analysis for both E and H-plane bending results in more resonant detuning when compared to the TPA. For E-plane bending the TPA-MF detunes to 2.456 MHz for a bending angle of 180° , while the TPA only detunes to 2.427 GHz for the same bending angle. The realized gain for the TPA-MF suffers more than the RPA and TPA for increasing bending angles. E-plane bending improved the realized gain and directivity for the TPA and RPA for small bending angles up to 60° , however for the TPA-MF the realized gain and directivity falls sharply. For an E-plane bending angle of 120° the realized gain of the TPA outperforms the TPA-MF. The directivity of the TPA outperforms the TPA-MF after a bending angle of 90° . The TPA-MF has an increasing bandwidth for increasing bending angles which is shown to outperform both the RPA and the TPA.

For H-plane bending, the TPA-MF has a resonant shift to 2.421 GHz for a bending angle of 180° . The antenna's realized gain, directivity and bandwidth decrease as the bending angle increases but not as much when compared to E-plane bending. The TPA-MF maintains superior performance to that of the TPA and RPA over all the bending angles analysed.

E-plane bending the TPA-MF for a bending angle of 150° results in the intended operating frequency of 2.4 GHz detuning outside of the antenna's bandwidth. This is the same for the RPA. The TPA is shown to be the most resilient to E-plane bending only detuning outside of the antenna's bandwidth for a bending angle of 180° .

Overall, it can be shown that the radiation performance of the TPA-MF exceeds that of the RPA and TPA. However, when bending is applied to the antenna the performance of the TPA-MF varies depending on the type of bending. E-plane bending for the TPA-MF is shown to have more detrimental effects than H-plane bending.

8.1.4 Chapter 7

Introducing human tissue to the proposed TPA and TPA-MF has a small performance enhancing effect on the TPAs and TPA-MFs directivity. Introducing human tissue models behave as if it were extending the antenna's ground plane. Due to the lossy nature of human tissue, the total radiation efficiency of the antenna reduces compared to its free space counterpart.

For the antenna operating at 2.4 GHz, SAR is shown to be low and complies well within the recommended limits for all antenna model configurations.

Introducing antenna bending by placing the antenna on a curved surface such as the human arm model shows that it is important to consider the orientation of the antenna. Chapter 6 showed that there are significant differences between H-plane and E-plane bending for the TPA and TPA-MF. Simulating the TPA and TPA-MF on the FEM Human Model did not show much change compared to the free space equivalent simulations in Chapter 6.

When the antennas are placed on a curved cylindrical surface such as an arm, leg, torso etc, H-plane bending the antenna across the surface is recommended as opposed to E-plane bending for the proposed TPA-MF. H-plane bending the TPA-MF results in less resonant frequency detuning and maintains a superior realized gain for increasing bending angles compared to the TPA.

Modelling the FEM Human Model is computationally expensive. It was shown that this type of detailed tissue model can be simulated accurately with mesh refinement techniques without the use of a high-performance workstation provided the simulation is confined to the area of interest.

8.2 Future Research Directions

8.2.1 Fabrication and Measurements

The research presented in this thesis can be taken forward in various ways. First and foremost a natural progression from the research undertaken so far would be to focus on fabrication and measurements for the proposed TPA and TPA-MF using a digitally controlled sewing machine to create a fully textile antenna.

A prototype was generated for the TPA which is designed as a partial textile with a cotton substrate and copper tape used for the conductive elements. So far only reflection parameters have been obtained using a NanoVNA. Measurements for the radiation pattern, realized gain, directivity, radiation efficiency, impedance are required to validate the simulated results. A prototype for the TPA-MF has also been created however no measurements have been obtained as of now.

Measurements are also required to validate the characterization of the AMC metafilm's susceptibilities using the GSTC method.

8.2.2 Analytical Analysis

The AAM in Chapter 4 only considers radiation in the upper hemisphere of the TPA's radiation pattern. Future work for this model includes incorporating edge diffraction effects around the ground plane using straight edge diffraction equations derived from Uniform Theory of

Diffraction (UTD). Future work also includes incorporating H-plane bending to the Cylindrical-AAM. In H-plane bending the apertures themselves are bent resembling a sector of an annulus. Therefore, the form of the electric field of the radiating apertures will not be the same as for E-plane bending.

8.2.3 Further Miniaturization

Miniaturization in this research was focused solely on the patch antenna's radiating element, while the adjacent structures such as the size of the ground plane remain constant. This type of miniaturization is regarded as a secondary miniaturization. Further miniaturization of the antenna can be achieved by reducing the size of the ground plane or the number of AMC unit cells used in the TPA-MF. It may be worth investigation if there is a critical size which enables the proposed TPA and TPA-MF to remain compact yet have optimal radiation performance parameters.

8.2.4 Applying Further Bandwidth Enhancement Techniques

The TPA-MF was shown to have a bandwidth enhancement. However, this can be increased by investigating bandwidth enhancing techniques using parasitic elements or slots to create resonances around the patch's resonant frequency and without alteration to the antenna's ground plane.

8.2.5 Further Simulation of the Proposed Antennas on the Human Body

Full-wave simulations with the antennas on a full body model need to be performed using a high-performance workstation to determine whether a partial body model is adequate for an accurate characterization on how the antenna interacts with the other parts of the body.

8.2.6 Method for Implementing Circular Polarization

A circularly polarized patch is known to be favourable for wearable applications due to their improved signal robustness with human body movement [1]. Furthermore, bending circularly polarized antennas is known to remove the circular polarization characteristics [2]. Therefore, a method to circularly polarize the TPA and TPA-MF while maintaining circular polarization characteristics during bending operations is to be investigated.

8.3 Reflection

This PhD research started as an investigation of textile microstrip patch antennas with the focus on modelling and simulation. The aim was to investigate whether woven textiles influenced the patch antenna's radiation performance parameters.

Although not formally reported, an attempt to simulate woven structures in CST Microwave Studio was explored. Attempts were made to generate textile models in the software TexGen and exporting them into CST. There were also attempts to manually build a model of the woven textile patch antenna in CST. However due to the exceedingly large amount and poor quality of mesh cells during the discretization process of the model, it did not seem possible to achieve an accurate simulation while incorporating mesh refinement processes with the available computational capabilities. A decision was made not to pursue the simulation challenges further.

Consequently, the research focus moved to other challenges related to wearable textile patch antenna's, specifically, investigating methods for antenna bending and techniques to miniaturize the antenna without sacrificing performance in the presence of surrounding human tissue. This journey taken on pushing the developments of a suitable wearable textile patch antenna has led to the proposed inset-fed TPA being miniaturized and more resilient to antenna bending compared the traditional RPA. The TPA-MF was also proposed which achieved a performance enhancement regarding the antenna's directivity, realized gain and bandwidth, allowing for improvements in operational performance. A performance comparison between the proposed antennas and various textile patch antennas is presented in Table 8.1. The TPA-MF has a high realized gain of up to 6.56 dBi which outperforms most of the reported antennas. The probe fed textile rectangular patch antenna reported in [3] has marginally better gain, however the proposed TPA and TPA-MF use a microstrip feed which guarantees a flat structure and the ability to use a relatively thin substrate. The microstrip feed also allows the assembly of electronic components directly on the fabric in antenna proximity [4]. The TPA-MF also showed a bandwidth enhancement compared to the TPA without using bandwidth enhancing techniques such as an aperture coupled feedline [5] and introducing slots into the resonator [6].

As discussed in Chapter 1, various application examples for wearable patch antennas were highlighted such as on-body power harvesting and location transmitters. In this research the TPA-MF was shown to miniaturize the radiating element to that of a conventional RPA while also maintaining bandwidth, radiation efficiency and high gain performance. In addition to the peak gain of the TPA-MF, it is also important to consider how its coverage is distributed around the body [7]. A body-worn patch antenna with a narrow beamwidth is more sensitive to the wearer's movements and will experience a significant decrease in performance as a

result. Multiple antenna elements placed strategically around the body can result in a radiation pattern with a high gain over a wide field-of-view. The TPA-MF is a good candidate for an on-body distributed array for off-body communication links as it is low-profile, resilient to antenna H-plane deformations and the influence of lossy human tissue. Such distributed array applications include on-body power harvesting [8], space communication and navigation system for astronauts [7] and body-worn MIMO systems [9].

Ref	Patch Antenna Configuration	f_{res} (GHz)	ϵ_r	$\tan \delta$	Antenna Size (mm ³)	RESA (mm ²)	Gain (dBi)	Bandwidth (%)	ϵ_0 (%)	Year
[10]	Textile Rectangular-Ring	2.45	1.25	-	78 x 73.5	2548	-	6.9	70	2006
[3]	Textile Rectangular Patch	2.44	1.9	0.0045	115 x 73 x 4	1358	6.6	6.2	78	2008
[11]	Inset-Fed Rectangular Patch	2.45	1.6	0.02	100 x 100 x 2.808	1922	-	-	41.4	2012
[12]	Dual Band Patch with AMC	2.45/5.20	1.2	0.044	100 x 100 x 3	2087	2.5/4	-	>40	2014
[13]	Textile Dual-Band Patch	2.40/5.20	1.3	0.044	50 x 20 x 6.34	200	-3.8/6.2	5.51/13.7	-	2015
[14]	Textile Rectangular Patch	2.40	2.1	-	100 x 100 x 1.06	2050	-	1.25	-	2017
[5]	Aperture Coupled Textile Patch	2.45	1.35	0.044	100 x 100 x 3.34	1267	5.42	9.72	50.8	2023
This work	Inset-Fed Trapezoidal Patch	2.40	1.6	0.02	110 x 110 x 2.87	1749	4.57*	1.84	40.9	2023
This work	Inset-Fed Trapezoidal Patch with AMC	2.40	1.6	0.02	110 x 110 x 2.87	1288	6.56*	3.07	48.3	2023

Note: RESA = Radiating Element Surface Area

*Realized Gain

Table 8.1: Proposed antennas compared with some textile patch antennas reported in Chapter 2.

8.4 References

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Appendix

A. MATLAB Codes

A.1 Dual-AAM to compute far-field patterns for the TPA

This section provides a walkthrough of the implementation of the Aperture Array Model (AAM) presented in Chapter 4 using MATLAB.

```
clc

format long;

% units
meters      = 1;
seconds      = 1;
hertz        = 1/seconds;
millimeters  = 1e-3 * meters;
gigahertz    = 1e9 * hertz;
seconds      = 1;
degrees      = pi/180;

% constants
c0 = 299792458 * meters/seconds;
e0 = 8.854187812813e-12 * 1/meters;
u0 = 1.256637062121e-6 * 1/meters;
N0 = 376.7303136686;
er = 1.6; %relative permittivity of patch substrate

h = 2.8*millimeters; %height of aperture

f = 2.4*gigahertz; %resonant frequency
lambda = c0/f; %wavelength
k = (2*pi)/(lambda) ; %wavenumber

% Computation of the patch antennas width and length and distance to
% farfield
w = (c0/(2*f))*sqrt(2/(er+1)); %Aperture width

e_reff = ( (er+1)/2 ) + ( (er-1)/2 )*( 1+ 12*(h/w))^(-1/2); %effecitive dieletric
constant

l_ext = ( 0.412*(h)) * ((e_reff +0.3) * ( (w/h) + 0.264 )) ...,
        / ((e_reff -0.258) * ( (w/h) + 0.8 )) ; %length extension due to fringe fields

l = ( 1 / (2*f*sqrt(e_reff)*sqrt(u0*e0)) ) - 2*l_ext ; %length between apertures
l_eff = l + 2*l_ext; %total length including fringe fields
r = 2*(w)^2/(lambda) ; %farfield distance (diagonal distance of trapezoid)

% E-plane
PHI = 0;
THETA=(0:360)*pi/180; % convert degrees to radians

%Forming the vector of various widths for the aperture of width W'
w_prime_sweep = linspace(5*millimeters, w,2);

no_sweep = numel(w_prime_sweep); %number of elements in the sweep
P_n_container = zeros(no_sweep,numel(THETA)); %preallocated container

index = 1;

for w_prime = w_prime_sweep %calculating P_n for each value of W'
```

```

%calculating the diagonal length l_d of equilateral trapezoid
g = ( w - w_prime ) ./ 2 ;
l_d = sqrt( (l)^2 + (g).^2 );

%angle of triangle that forms the sides of the trapezoid
alpha = asind(g./l_d);
alpha_rad = alpha .* degrees;

%midline distance in trapezoid
w_c = (w+w_prime)/2;

d = (k*h)/(2*pi*r);

X = (k*(w/2)).*sin(THETA).* (sin(PHI));           %argument X
Y = (k*(w_prime/2)).*sin(THETA).*sin(PHI);        %argument Y

w_fun1 = w.*sincX(X) ;
w_fun2 = w_prime.*sincX(Y) ;
exp_w = 1i*k*l_eff*cos(THETA)/2;
exp_wprime = -1i*k*l_eff*cos(THETA)/2;

%E-fields for the apertures across the widths of the patch
E_theta = -1i*d .*cos(PHI).*(exp(-1i*k*r)/r) .* ...,
    ( (w_fun1.*exp(exp_w)) + (w_fun2.*exp(exp_wprime)) );

E_phi = 1i*d .*cos(THETA).*sin(PHI).*(exp(-1i*k*r)/r) .* ...,
    ( (w_fun1.*exp(exp_w)) + (w_fun2.*exp(exp_wprime)) );

E_theta_total = E_theta ;
E_phi_total = E_phi ;

absE_thetasq1 = abs( E_theta).^2; %absolute value squared for E_theta
absE_phisq1 = abs( E_phi).^2; %absolute value squared for E_phi

absE_thetasq2 = abs( E_theta_total).^2; %absolute value squared for E_theta for
Quad-AAM
absE_phisq2 = abs( E_phi_total).^2; %absolute value squared for E_phi for Quad-AAM

W_rad = ( 1/(2*N0) ) .* ( absE_thetasq2 + absE_phisq2);

U = r^2 .* W_rad;

P_n_container(index,:) = 10*log10((U)./(max(U))); %normalized power pattern or
radiation intensity plot

index = index + 1;

end

figure(1)
plot((180/pi)*THETA, (P_n_container(1,:)), 'LineWidth', 1.5,
'Color', 'k', 'LineStyle', '-');
hold on
plot((180/pi)*THETA, (P_n_container(2,:)), 'LineWidth', 1.5 ,
'Color', 'k', 'LineStyle', '--');

xlim([0 90]);
ylabel('$ \mathrm{U}(\theta, \phi) / \mathrm{U}_{\max} \backslash ; \mathrm{dB} \}$', 'Interpreter', 'latex', 'FontSize', 12);
xlabel('$\theta$ $ (^{\circ}) $', 'Interpreter', 'latex', 'FontSize', 12);
title('E-plane', 'Interpreter', 'latex', 'FontSize', 12)
grid on;
legend({'$W_2 = 5$ mm', '$W_2 = W_1 = 54.8$ mm'}, 'Location', 'Southeast',
'Interpreter', 'latex');

```

```

% H-plane
figure(2)
THETA = pi/2;
PHI=(0:360)*pi/180;    % convert degrees to radians

%Forming the vector of various widths for the aperture of width W'
w_prime_sweep = linspace(5*millimeters, w,2);

no_sweep = numel(w_prime_sweep); %number of elements in the sweep
P_n_container2 = zeros(no_sweep,numel(PHI)); %preallocated container

index = 1;

for w_prime = w_prime_sweep %calculating P_n for each value of W'

%calculating the diagonal length l_d of equilateral trapezoid
g = ( w - w_prime ) ./ 2 ;
l_d = sqrt( (l)^2 + (g).^2 );

%angle of triangle that forms the sides of the trapezoid
alpha = asind(g./l_d);
alpha_rad = alpha .* degrees;

%midline distance in trapezoid
w_c = (w+w_prime)/2;

X = (k*(w/2)).*sin(THETA).*sin(PHI);    %argument X
Y = (k*(w_prime/2)).*sin(THETA).*sin(PHI);    %argument Y

w_fun1 = w.*sincX(X) ;
w_fun2 = w_prime.*sincY(Y) ;
exp_w = 1i*k*l*cos(THETA)/2;
exp_wprime = -1i*k*l*cos(THETA)/2;

%E-fields for the apertures across the widths of the patch
E_theta = -1i*d .*cos(PHI).* sincX(k*h*sin(THETA).*cos(PHI)/2) .* (exp(-1i*k*r)/r)
.* ... ,
    ( (w_fun1.*exp(exp_w)) + (w_fun2.*exp(exp_wprime)) );

E_phi = 1i*d .*cos(THETA).*sin(PHI).*sincX(k*h*sin(THETA).*cos(PHI)/2) .* (exp(-
1i*k*r)/r) .* ... ,
    ( (w_fun1.*exp(exp_w)) + (w_fun2.*exp(exp_wprime)) );

%E-fields for the non-radiating apertures across the length of the patch

% sin taper
E_phi_Ld = ((k*l_d)^2/pi) * (sin(2*THETA)) .* sincX(k*h*sin(THETA).*cos(PHI)/2)
.* ... ,
    .* ( cos(k*l_d*cos(THETA)/2) ./ ((pi)^2 - (k*l_d*cos(THETA)).^2 ) ).*... ,
    (cos((1/2)*k*w_c * sin(THETA).*cos(PHI)) ).*(exp(-1i*k*r)/r);

E_theta_Ld = 0;

E_theta_total = E_theta + (E_theta_Ld) ;
E_phi_total = E_phi + (E_phi_Ld);

absE_thetasq1 = abs( E_theta).^2; %absolute value squared for E_theta
absE_phisq1 = abs( E_phi).^2; %absolute value squared for E_phi

absE_thetasq2 = abs( E_theta_total).^2; %absolute value squared for E_theta for
Quad-AAM
absE_phisq2 = abs( E_phi_total).^2; %absolute value squared for E_phi for Quad-AAM

W_rad = ( 1/(2*N0) ) .* ( absE_thetasq2 + absE_phisq2 );

```



```

u0 = 1.256637062121e-6 * 1/meters;
N0 = 376.7303136686;

er = 1.6; %relative permittivity of patch substrate
h = 2.8*millimeters; %height of aperture

f = 2.4*gigahertz; %resonant frequency
lambda = c0/f; %wavelength
k = (2*pi)/(lambda) ; %wavenumber

% Computation of the patch antennas width and length and distance to
% farfield
w = (c0/(2*f))*sqrt(2/(er+1)); %Aperture width

e_reff = ( (er+1)/2 ) + ( (er-1)/2 )*( 1+ 12*(h/w))^(1/2); %effective dielectric
constant

l_ext = ( 0.412*(h) ) * ((e_reff +0.3) * ( (w/h) + 0.264 )) ... ,
/ ((e_reff -0.258) * ( (w/h) + 0.8 )) ; %length extension due to fringe fields

l = ( 1 / (2*f*sqrt(e_reff)*sqrt(u0*e0)) ) - 2*l_ext ; %length between apertures
l_eff = l + 2*l_ext;

r = ( 2/(lambda) ) * (w)^2; %farfield distance calculated with the largest
dimension of the antenna

% The upper and lower limits of  $\theta$  and  $\phi$  in degrees are defined.
% The upper limit for  $\theta = 90^\circ$  defines that
% radiation is propagating normal to the patch in the upper hemisphere

Theta_L=0;
Theta_U=90;
Phi_L=0;
Phi_U=360;

%Creating the meshgrid

theta=(Theta_L:Theta_U)*pi/180; % convert degrees to radians
theta=theta-0.0001;
phi=(Phi_L:Phi_U)*pi/180; % convert degrees to radians
phi=phi-0.0001; % term added since first element cannot be zero.

%trailing edge incrementation for theta and phi
dth=theta(2)-theta(1);
dph=phi(2)-phi(1);

[THETA,PHI]=meshgrid(theta,phi);

%Forming the vector of various widths for the aperture of width W'
w_prime_sweep = linspace(0.1*millimeters, w,100);

no_sweep = numel(w_prime_sweep); %number of elements in the sweep

%preallocated container for computed directivities using the dual and quad
%aperture array model

dual_direct_container_dB = zeros(no_sweep,1);
non_radiating_edges_direct_container_dB = zeros(no_sweep,1);
index = 1;

for w_prime = w_prime_sweep %calculating the directivity for each value of W'

%calculating the diagonal length l_d of equilateral trapezoid
g = ( w - w_prime ) ./ 2 ;

```

```

l_d = sqrt( (l)^2 + (g).^2 );

% angle of triangle that forms the sides of the trapezoid
% alpha = asind(g./l_d);
% alpha_rad = alpha .* degrees;

%midline distance in trapezoid
w_c = (w+w_prime)/2;

%some variables defined to shorten the expression for the E-fields
d = (k*h)/(2*pi*r);
X = (k*(w/2)).*sin(THETA).*sin(PHI);          %argument X
Y = (k*(w_prime/2)).*sin(THETA).*sin(PHI);      %argument Y

w_fun1 = w.*sincX(X) ;
w_fun2 = w_prime.*sincX(Y) ;
exp_w = li*k*l_eff*cos(THETA)/2;
exp_wprime = -li*k*l_eff*cos(THETA)/2;

%E-fields for the apertures across the widths of the patch
E_theta = -li*d .*cos(PHI).* sincX(k*h*sin(THETA).*cos(PHI)/2) .* (exp(-li*k*r)/r)
.* ... ,
    ( (w_fun1.*exp(exp_w)) + (w_fun2.*exp(exp_wprime)) );

E_phi = li*d .*cos(THETA).*sin(PHI).*sincX(k*h*sin(THETA).*cos(PHI)/2) .* (exp(-
li*k*r)/r) .* ... ,
    ( (w_fun1.*exp(exp_w)) + (w_fun2.*exp(exp_wprime)) );

%non-radiating apertures (horizontal components)

E_theta_g = (-li*k*h*g/2*pi) * cos(PHI) .* sincX(k*h*sin(THETA).*cos(PHI)/2) ... ,
    .*sincX(k*g*sin(THETA).*cos(PHI)).* (2*cos(k*w_c*cos(THETA)/2)) .* (exp(-
li*k*r)/r);
E_phi_g = (-li*k*h*g/2*pi) * cos(THETA).*sin(PHI) .*
sincX(k*h*sin(THETA).*cos(PHI)/2) ... ,
    .*sincX(k*g*sin(THETA).*cos(PHI)).* (2*cos(k*w_c*cos(THETA)/2)) .* (exp(-
li*k*r)/r);

%E-fields for the non-radiating apertures across the length of the patch
%(vertical component)
% cosine taper
E_phi_Ld = (-li*k*h*(l)^2) * (sin(THETA)) .* sincX(k*h*sin(THETA).*cos(PHI)/2) ... ,
    .* ( cos(k*l*cos(THETA)/2) ./ ((pi)^2 - (k*l*cos(THETA)).^2 ) ) .* ... ,
    (2*cos((l/2)*k* w_c * cos(acos(sin(THETA).*sin(PHI)) ) ) ) ).*(exp(-
li*k*r)/r);

E_theta_Ld = 0;

E_theta_total = E_theta + E_theta_g + E_theta_Ld ;
E_phi_total = E_phi + E_phi_g + E_phi_Ld;

absE_thetasq1 = abs( E_theta).^2; %absolute value squared for E_theta
absE_phisq1 = abs( E_phi).^2; %absolute value squared for E_phi

absE_thetasq2 = abs( E_theta_total).^2; %absolute value squared for E_theta for
Quad-AAM
absE_phisq2 = abs( E_phi_total).^2; %absolute value squared for E_phi for Quad-AAM

W_rad1 = ( 1/(2*N0) ) .* ( absE_thetasq1 + absE_phisq1 ); %Dual-AAM average power
density
W_rad2 = ( 1/(2*N0) ) .* ( absE_thetasq2 + absE_phisq2); %Quad-AAM average power
density

U1 = r^2 .* W_rad1; %radiation intensity for Dual-AAM
U2 = r^2 .* W_rad2; %radiation intensity for Quad-AAM

%numerical computation of the radiated power for the Dual-AAM and Quad-AAM
Prad1=sum(sum(U1.*sin(THETA)*dth*dph));

```

```

D1=4*pi*max(max(U1))/Prad1; %linear directivity calculation
D_db1=10*log10(D1); %directivity in dB

Prad2=sum(sum(U2.*sin(THETA)*dth*dph));
D2=4*pi*max(max(U2))/Prad2;
D_db2=10*log10(D2);

%maximum directivities for varyin W' collected here
dual_direct_container_dB(index) = D_db1; %Dual-AAM
non_radiating_edges_direct_container_dB(index) = D_db2; %Quad-AAM

index = index + 1;

end

figure(1)
plot(w_prime_sweep*1e3,dual_direct_container_dB, 'LineWidth', 1.5 ,
'Color','k','LineStyle','--');
xlim([5 55]);
% ylim([3 12]);
ylabel('Directivity (dBi)', 'Interpreter','latex');
xlabel('Aperture Width W$_2$(mm)', 'Interpreter','latex');
grid on;
hold on
plot(w_prime_sweep*1e3,non_radiating_edges_direct_container_dB, 'LineWidth', 1.5 ,
'Color','k','LineStyle','-');
hold on
% W_prime1 = [54.8 45 40 35 30 25 20 15 10 5]; %values for W' used in full-wave
simulations
W_prime1 = [5 10 15 20 25 30 35 40 45 54.8];
Dir_fd = [8.14 8.07 7.99 7.96 7.89 7.85 7.80 7.77 7.78 7.74 ]; %full-wave
simulated directivites using frequency domain solver
Dir_td = [8.38 8.30 8.26 8.22 8.22 8.1 8.05 8.04 8.04 8.03 ]; %full-wave simulated
directivities using time-domian solver

% legend({'Dual-Aperture Array Model', 'Quad-Aperture Array Model'}, 'Location',
'Southeast', 'Interpreter','latex');

coefficients = polyfit(W_prime1, flip(Dir_fd), 1);
xFit = linspace(min(W_prime1), max(W_prime1), 1000);
yFit = polyval(coefficients , xFit);
plot(W_prime1, flip(Dir_fd), 'k.', 'MarkerSize', 15); % Plot training data.
hold on; % Set hold on so the next plot does not blow away the one we just drew.
plot(xFit, yFit, 'k-', 'LineWidth', 1.5); % Plot fitted line.
grid on;

% plot(W_prime,Band_w_td, 'lineWidth', 1.5, 'Color','k','LineStyle','-');

hold on;
coefficients2 = polyfit(W_prime1, flip(Dir_td), 1);
yFit2 = polyval(coefficients2 , xFit);
plot(W_prime1, flip(Dir_td), 'ko', 'MarkerSize', 5); % Plot training data.
hold on; % Set hold on so the next plot does not blow away the one we just drew.
plot(xFit, yFit2, 'k:', 'LineWidth', 1.5); % Plot fitted line.

% plot(W_prime1,Dir_fd, 'LineWidth', 1.5, 'Color','k','LineStyle',':');
% plot(W_prime1,Dir_td, 'LineWidth', 1.5, 'Color','k','LineStyle','-');
legend({'Dual-Aperture Array Model', 'Quad-Aperture Array Model', 'Simulated
Frequency Domain',...,
'Frequency Domain Trendline','Simulated Time Domain', 'Time Domain Trendline'
}, 'Location', 'Southeast', 'Interpreter','latex');

figure(2)
plot(w_prime_sweep*1e3,dual_direct_container_dB, 'LineWidth', 1.5,
'Color','k','LineStyle','-');

```

```

xlim([5 55]);
% ylim([3 12]);
ylabel('Directivity (dBi)', 'Interpreter','latex', 'FontSize', 12);
xlabel('Aperture Width  $\mathrm{W}_2$  (mm)', 'Interpreter','latex', 'FontSize', 12);
grid on;

function out = sincX(x)

    out = sin(x)./(x);
    if x == 0
        out = 1;
    end

end

```

A.3 Quad-AAM to compute far-field pattern for the TPA

```

close all;
clc
format long;

% DEFINED UNITS
meters      = 1;
seconds      = 1;
hertz        = 1/seconds;
millimeters = 1e-3 * meters;
gigahertz    = 1e9 * hertz;
seconds      = 1;
degrees      = pi/180;

% CONSTANT DEFINITIONS
c0 = 299792458 * meters/seconds;
e0 = 8.854187812813e-12 * 1/meters;
u0 = 1.256637062121e-6 * 1/meters;
N0 = 376.7303136686;
er = 1.6; %relative permittivity of patch substrate

h = 2.8*millimeters; %height of aperture
% l = 52*millimeters; %maximum dimension of patch
E0 = 1; %normalized

f = 2.4*gigahertz; %resonant frequency
lambda = c0/f; %wavelength
k = (2*pi)/(lambda) ; %wavenumber
% Computation of the patch antennas width and length and distance to
% farfield
w = (c0/(2*f))*sqrt(2/(er+1)); %Aperture width

e_reff = ( (er+1)/2 ) + ( (er-1)/2 )*( 1+ 12*(h/w))^(1/2); %effective dielectric
constant

l_ext = ( 0.412*(h) ) * ((e_reff +0.3) * ( (w/h) + 0.264 )) ... ,
        / ((e_reff -0.258) * ( (w/h) + 0.8 )) ; %length extension due to fringe fields

l = ( 1 / (2*f*sqrt(e_reff)*sqrt(u0*e0)) ) - 2*l_ext ; %length between apertures
l_eff = l + 2*l_ext; %total length including fringe fields
r = 2*(70*millimeters)^2/(lambda) ; %farfield distance (diagonal distance of
trapezoid)

% E-plane
PHI = 0;
THETA=(0:360)*pi/180; % convert degrees to radians

%Forming the vector of various widths for the aperture of width W'

```

```

w_prime_sweep = linspace(5*millimeters, w,2);

no_sweep = numel(w_prime_sweep); %number of elements in the sweep
P_n_container1 = zeros(no_sweep,numel(THETA)); %preallocated container
P_n_container2 = zeros(no_sweep,numel(THETA));

index = 1;

for w_prime = w_prime_sweep %calculating P_n for each value of W'

%calculating the diagonal length l_d of equilateral trapezoid
g = ( w - w_prime ) ./ 2 ;
l_d = sqrt( (l)^2 + (g).^2 );

%angle of triangle that forms the sides of the trapezoid
alpha = asind(g./l_d);
alpha_rad = alpha .* degrees;

%midline distance in trapezoid
w_c = (w+w_prime)/2;

d = (k*h)/(2*pi*r);

X = (k*(w/2)).*sin(THETA).* (sin(PHI)); %argument X
Y = (k*(w_prime/2)).*sin(THETA).*sin(PHI); %argument Y
% X = 0;
% Y = 0;

w_fun1 = w.*sincX(X) ;
w_fun2 = w_prime.*sincX(Y) ;
exp_w = 1i*k*l_eff*cos(THETA)/2;
exp_wprime = -1i*k*l_eff*cos(THETA)/2;

%E-fields for the apertures across the widths of the patch
E_theta = -1i*d .*cos(PHI).* sincX(k*h*sin(THETA).*cos(PHI)/2) .* (exp(-1i*k*r)/r)
.* ... ,
( (w_fun1.*exp(exp_w)) + (w_fun2.*exp(exp_wprime)) );

E_phi = 1i*d .*cos(THETA).*sin(PHI).*sincX(k*h*sin(THETA).*cos(PHI)/2) .* (exp(-
1i*k*r)/r) .* ... ,
( (w_fun1.*exp(exp_w)) + (w_fun2.*exp(exp_wprime)) );

%non-radiating apertures (horizontal components)

E_theta_g = (-1i*k*h*g/2*pi) * cos(PHI) .* sincX(k*h*sin(THETA).*cos(PHI)/2) ... ,
.*sincX(k*g*sin(THETA).*cos(PHI)).* (2*cos(k*w_c*cos(THETA)/2)) .* (exp(-
1i*k*r)/r);
E_phi_g = (-1i*k*h*g/2*pi) * cos(THETA).*sin(PHI) .*
sincX(k*h*sin(THETA).*cos(PHI)/2) ... ,
.*sincX(k*g*sin(THETA).*cos(PHI)).* (2*cos(k*w_c*cos(THETA)/2)) .* (exp(-
1i*k*r)/r);

%E-fields for the non-radiating apertures across the length of the patch
%(vertical component)
% cosine taper
E_phi_Ld = (-1i*k*h*(l)^2) * (sin(THETA)) .* sincX(k*h*sin(THETA).*cos(PHI)/2) ... ,
.* ( cos(k*l*cos(THETA)/2) ./ ((pi)^2 - (k*l*cos(THETA)).^2 ) ).*... ,
(2*cos((1/2)*k* w_c * cos(acos(sin(THETA).*sin(PHI)) ) ) ) ).*(exp(-
1i*k*r)/r);

E_theta_Ld = 0;

E_theta_total = E_theta + E_theta_g + E_theta_Ld ;
E_phi_total = E_phi + E_phi_g + E_phi_Ld;

E_thetasq1 = ( E_theta).^2; %absolute value squared for E_theta
E_phisq1 = ( E_phi).^2; %absolute value squared for E_phi

```

```

E_thetasq2 = ( E_theta_total).^2; %absolute value squared for E_theta for Quad-AAM
E_phisq2 = ( E_phi_total).^2; %absolute value squared for E_phi for Quad-AAM

W_rad1 = ( abs(E_thetasq1 + E_phisq1) ); %Dual-AAM average power density
W_rad2 = ( abs(E_thetasq2 + E_phisq2) ); %Quad-AAM average power density

U1 = W_rad1; %radiation intensity for Dual-AAM
U2 = W_rad2; %radiation intensity for Quad-AAM

P_n_container1(index,:) = 20*log10((U1)/(max(U1))); %normalized power pattern or
radiation intensity plot
P_n_container2(index,:) = 20*log10((U2)/(max(U2)));

index = index + 1;

end

figure(1)
% plot((180/pi)*THETA, (P_n_container1(1,:)), 'LineWidth', 1.5,
'Color','k','LineStyle','-');
% hold on
% plot((180/pi)*THETA, (P_n_container1(2,:)), 'LineWidth', 1.5 ,
'Color','k','LineStyle',':');
%
% hold on

plot((180/pi)*THETA, (P_n_container2(2,:)), 'LineWidth', 1.5 ,
'Color','k','LineStyle','-','...',
'Marker', 'none',
'MarkerIndices',1:3:length(P_n_container2(2,:)), 'MarkerSize',4);
hold on

load('e_plane_rpa_1d.txt');
x3 = e_plane_rpa_1d(:,1);
y3 = e_plane_rpa_1d(:,2);
plot(flip(x3),y3, 'LineWidth', 1.5, 'LineStyle',':','Color','k', 'Marker',
'none', 'MarkerSize',4, ...,
'MarkerIndices',1:3:length(P_n_container2(2,:)));

hold on
load('e_plane_rpa_TD_1d.txt');
x3 = e_plane_rpa_TD_1d(:,1);
y3 = e_plane_rpa_TD_1d(:,3);
plot(flip(x3),y3, 'LineWidth', 1.5, 'LineStyle','-','Color','k', 'Marker',
'none', 'MarkerSize',4, ...,
'MarkerIndices',1:3:length(P_n_container2(2,:)));

xlim([0 90]);
ylabel('$ |\mathrm{E}(\theta, \phi)| / \mathrm{E}_{\max} | \mathrm{dB} )$', 'Interpreter','latex', 'FontSize', 12);
xlabel('$\theta$ ($^\circ$)', 'Interpreter','latex', 'FontSize', 12);
title('E-plane', 'Interpreter','latex', 'FontSize', 12)

grid on;

legend({'Quad-AAM RPA',...,
'Simulated RPA Frequency Domain Solver', 'Simulated RPA Time Domain Solver'},
...,
'Location', 'Southeast', 'Interpreter','latex');

figure(2)
plot((180/pi)*THETA, (P_n_container2(1,:)), 'LineWidth', 1.5,
'Color','k','LineStyle','-','...',
'Marker', 'none',
'MarkerIndices',1:3:length(P_n_container2(2,:)), 'MarkerSize',5);
hold on

```

```

load('e_plane_tpa_1d.txt');
x4 = e_plane_tpa_1d(:,1);
y4 = e_plane_tpa_1d(:,2);
plot(flip(x4),y4, 'LineWidth', 1.5, 'LineStyle',':', 'Marker','none','Color','k',
....,
      'MarkerIndices',1:3:length(P_n_container2(2,:)));

hold on
load('e_plane_tpa_TD_1d.txt');
x3 = e_plane_tpa_TD_1d(:,1);
y3 = e_plane_tpa_TD_1d(:,3);
plot(flip(x3),y3, 'LineWidth', 1.5, 'LineStyle','-','Color','k', 'Marker',
'none','MarkerSize',4,...,
      'MarkerIndices',1:3:length(P_n_container2(2,:)));

xlim([0 90]);
ylabel('$ |\mathrm{E}(\theta,\phi)| / \mathrm{E}_{\max} | \backslash; \mathrm{(dB)} $',
'Interpreter','latex', 'FontSize', 12);
xlabel('$\theta$ $(^\circ)$ $', 'Interpreter','latex', 'FontSize', 12);
title('E-plane','Interpreter','latex', 'FontSize', 12)

grid on;
legend({'Quad-AAM TPA',...,
      'Simulated TPA Frequency Domain Solver', 'Simulated TPA Time Domain Solver'},
....,
      'Location', 'Southeast', 'Interpreter','latex');

% figure(3)
% polarplot(THETA,P_n_container,'LineWidth', 1.5);
% rlim([-55 0])

%H-plane
THETA = pi/2;
PHI=(0:360)*pi/180;    % convert degrees to radians

%Forming the vector of various widths for the aperture of width W'
w_prime_sweep = linspace(5*millimeters, w,2);

no_sweep = numel(w_prime_sweep); %number of elements in the sweep
P_n_container3 = zeros(no_sweep,numel(PHI)); %preallocated container
P_n_container4 = zeros(no_sweep,numel(PHI)); %preallocated container

index = 1;

for w_prime = w_prime_sweep %calculating P_n for each value of W'

%calculating the diagonal length l_d of equilateral trapezoid
g = ( w - w_prime ) ./ 2 ;
l_d = sqrt( (l)^2 + (g).^2 );

%angle of triangle that forms the sides of the trapezoid
alpha = asind(g./l_d);
alpha_rad = alpha .* degrees;

%midline distance in trapezoid
w_c = (w+w_prime)/2;

X = (k*(w/2)).*sin(THETA).*sin(PHI);           %argument X

```

```

Y = (k*(w_prime/2)).*sin(THETA).*sin(PHI);    %argument Y

w_fun1 = w.*sincX(X) ;
w_fun2 = w_prime.*sincY(Y) ;
exp_w = 1i*k*l_eff*cos(THETA)/2;
exp_wprime = -1i*k*l_eff*cos(THETA)/2;

%E-fields for the apertures across the widths of the patch
E_theta = -1i*d .*cos(PHI).* sincX(k*h*sin(THETA).*cos(PHI)/2) .* (exp(-1i*k*r)/r)
.* ... ,
    ( (w_fun1.*exp(exp_w)) + (w_fun2.*exp(exp_wprime)) );

E_phi = 1i*d .*cos(THETA).*sin(PHI).*sincX(k*h*sin(THETA).*cos(PHI)/2) .* (exp(-
1i*k*r)/r) .* ... ,
    ( (w_fun1.*exp(exp_w)) + (w_fun2.*exp(exp_wprime)) );

%E-fields for the non-radiating apertures across the length of the patch

% sin taper
% E_phi_Ld = ((k*l_d)^2/pi) * (sin(2*THETA)) .* sincX(k*h*sin(THETA).*cos(PHI)/2)
... ,
%      .* ( cos(k*l_d*cos(THETA)/2) ./ ((pi)^2 - (k*l_d*cos(THETA)).^2 ) ).*... ,
%      (cos((1/2)*k*w_c * sin(THETA).*cos(PHI)) ).*(exp(-1i*k*r)/r);

% sin taper correction
E_phi_Ld = ((k*l_d)^2/pi) * (sin(2*THETA)) ... ,
    .* ( cos(k*l_d*cos(THETA)/2) ./ ((pi)^2 - (k*l_d*cos(THETA)).^2 ) ).*... ,
    (cos((1/2)*k*w_c * cos(acos(sin(THETA).*sin(PHI)) ) ) ).*(exp(-1i*k*r)/r);

E_theta_Ld = 0;

E_theta_total = E_theta + E_theta_g + E_theta_Ld ;
E_phi_total = E_phi + E_phi_g + E_phi_Ld;

E_thetasq1 = ( E_theta).^2; %absolute value squared for E_theta
E_phisq1 = ( E_phi).^2; %absolute value squared for E_phi

E_thetasq2 = ( E_theta_total).^2; %absolute value squared for E_theta for Quad-AAM
E_phisq2 = ( E_phi_total).^2; %absolute value squared for E_phi for Quad-AAM

W_rad1 = ( abs(E_thetasq1 + E_phisq1) ); %Dual-AAM average power density
W_rad2 = ( abs(E_thetasq2 + E_phisq2) ); %Quad-AAM average power density

U1 = W_rad1; %radiation intensity for Dual-AAM
U2 = W_rad2; %radiation intensity for Quad-AAM

P_n_container3(index,:) = 20*log10((U1)./(max(U1))); %normalized power pattern or
radiation intensity plot
P_n_container4(index,:) = 20*log10((U2)./(max(U2)));

index = index + 1;
end

P_n_container3(1,91) = NaN;
P_n_container3(2,91) = NaN;
P_n_container4(1,91) = NaN;
P_n_container4(2,91) = NaN;
% plot((180/pi)*PHI,(P_n_container3(1,:)), 'LineWidth', 1.5 ,
'Color','k','LineStyle','-');
% hold on
% plot((180/pi)*PHI,(P_n_container3(2,:)), 'LineWidth', 1.5 ,
'Color','k','LineStyle','--');
% hold on

figure(3)

```

```

plot((180/pi)*PHI, (P_n_container4(2,:)), 'LineWidth', 1.5 ,
'Color', 'k', 'LineStyle', '-.', ...
'Marker', 'none',
'MarkerIndices', 1:3:length(P_n_container4(2,:)), 'MarkerSize', 4);
hold on

load('h_plane_rpa_1d.txt');
x1 = h_plane_rpa_1d(:,1);
y1 = h_plane_rpa_1d(:,2);
plot(x1,y1, 'LineWidth', 1.5, 'LineStyle', ':', 'Color', 'k');
hold on

load('h_plane_rpa_TD_1d.txt');
x1 = h_plane_rpa_TD_1d(:,1);
y1 = h_plane_rpa_TD_1d(:,3);
plot(x1,y1, 'LineWidth', 1.5, 'LineStyle', '-.', 'Color', 'k');

xlim([0 90]);
ylabel('$ |\mathrm{E}(\theta, \phi)| / \mathrm{E}_{\max}| \; \mathrm{dB} \{ \}$', 'Interpreter', 'latex', 'FontSize', 12);
xlabel('$ \phi $ (^{\circ}) $', 'Interpreter', 'latex', 'FontSize', 12);
title('H-plane', 'Interpreter', 'latex', 'FontSize', 12)
grid on;

legend({'Quad-AAM RPA', ...,
'Simulated RPA Frequency Domain Solver', 'Simulated RPA Time Domain Solver'},
...,
'Location', 'Southeast', 'Interpreter', 'latex');

figure(4)
plot((180/pi)*PHI, (P_n_container4(1,:)), 'LineWidth', 1.5 ,
'Color', 'k', 'LineStyle', '-');
hold on

hold on
load('h_plane_tpa_1d.txt');
x2 = h_plane_tpa_1d(:,1);
y2 = h_plane_tpa_1d(:,2);
plot(x2,y2, 'LineWidth', 1.5, 'LineStyle', ':', 'Color', 'k');

hold on
load('h_plane_tpa_TD_1d.txt');
x2 = h_plane_tpa_TD_1d(:,1);
y2 = h_plane_tpa_TD_1d(:,3);
plot(x2,y2, 'LineWidth', 1.5, 'LineStyle', '-.', 'Color', 'k');

xlim([0 90]);
ylabel('$ |\mathrm{E}(\theta, \phi)| / \mathrm{E}_{\max}| \; \mathrm{dB} \{ \}$', 'Interpreter', 'latex', 'FontSize', 12);
xlabel('$ \phi $ (^{\circ}) $', 'Interpreter', 'latex', 'FontSize', 12);
title('H-plane', 'Interpreter', 'latex', 'FontSize', 12)
grid on;

legend({'Quad-AAM TPA', ...,
'Simulated TPA Frequency Domain Solver', 'Simulated TPA Time Domain Solver'},
...,
'Location', 'Southeast', 'Interpreter', 'latex');

function out = sincX(x)

    out = sin(x)./(x);
    if x == 0
        out = 1;
    end

end

```

```

function out = sincY(y)

    out = sin(y)./(y);
    if y == 0
        out = 1;
    end
end

```

A.4 Cylindrical-AAM to compute directivity for the TPA

```

clc

format long;

% DEFINED UNITS
meters      = 1;
seconds     = 1;
hertz       = 1/seconds;
millimeters = 1e-3 * meters;
gigahertz   = 1e9 * hertz;
seconds     = 1;

% CONSTANT DEFINITIONS
c0 = 299792458 * meters/seconds;
e0 = 8.854187812813e-12 * 1/meters;
u0 = 1.256637062121e-6 * 1/meters;
N0 = 376.7303136686;
er = 1.6; %relative permittivity of patch substrate

h = 2.8*millimeters; %height of aperture

f = 2.4*gigahertz; %resonant frequency
lambda = c0/f; %wavelength
k = (2*pi)/(lambda) ; %wavenumber
% k = k0*sqrt(er);

% Computation of the patch antennas width and length and distance to
% farfield
w = (c0/(2*f))*sqrt(2/(er+1)); %Aperture width
w_prime = 5*millimeters;
%horizontal component of l_d
g = ( w - w_prime ) ./ 2 ;

e_reff = ( (er+1)/2 ) + ( (er-1)/2 )*( 1 + 12*(h/w))^(1/2); %effecitve dieletric
constant

l_ext = ( 0.412*(h) ) * ((e_reff +0.3) * ( (w/h) + 0.264 ) ) ... ,
        / ((e_reff -0.258) * ( (w/h) + 0.8 ) ) ; %length extension due to fringe fields

l = ( 1 / (2*f*sqrt(e_reff)*sqrt(u0*e0)) ) - 2*l_ext ; %length between apertures
r = ( 2/(lambda) ) * (l^2); %farfield distance

%The upper and lower limits of theta and phi in degrees are defined.
%The upper limit for theta defines that radiation is propagating normal
%to the patch in the upper hemisphere
Theta_L=0;
Theta_U=90;
Phi_L=0;
Phi_U=360;

%Creating the meshgrid
theta=(Theta_L:Theta_U)*pi/180; % convert degrees to radians
theta=theta-0.0001;
phi=(Phi_L:Phi_U)*pi/180; % convert degrees to radians

```

```

phi=phi-0.0001; % term added since first element cannot be zero.

%trailing edge incrementation for theta and phi
dth=theta(2)-theta(1);
dph=phi(2)-phi(1);

[THETA,PHI]=meshgrid(theta,phi);

%antenna E-plane bending implementation for angles between 1 and 240
%degrees.
bend_angles_degrees = linspace(1,240,100);
bend_angles_rad = (pi/180) .* bend_angles_degrees; %convert to radians

radius_sweep = 1./ bend_angles_rad; %convert bending angles to equivalent radius
no_sweep = numel(radius_sweep); %number of elements to prepare sweep
direct_container_dB = zeros(no_sweep,1); %preallocated directivity container

index = 1;
for radius = radius_sweep

d = h/((pi^2).*radius_sweep(index));

X = (k*w/2)*cos(THETA);
w_fun1 = w*sincX(X);

Y = (k*w_prime/2)*cos(THETA);
w_fun2 = w_prime*sincX(Y);

Z = (k*g/2)*cos(THETA);
w_fun3 = g*sincX(Z);

gamma = k.*radius_sweep(index).*sin(THETA);

csf = zeros(size(THETA));
csf2 = zeros(size(THETA));

for n = 1:200 %computation of cylindrical space factor

    csf = csf + ((
(1i).^n)./besselh(n,2,gamma)).*cos(n.*1./2*radius_sweep(index)).*cos(n.*PHI) );
    csf2 = csf2 + (( (1i).^n)./besselh(n,2,gamma)).*cos(n.*PHI) );

end

E_phi_T = -d*(exp(-1i*k*r)./r) .* (w_fun1 + w_fun2 + 2*w_fun3)...,
.* ( 1./besselh(0,2,gamma) + 2 .* csf ).* (1./besselh(0,2,gamma) + 2 .* csf2 )
; %computing phi component of E-field

absE_phisq = abs( E_phi_T).^2; %absolute squared of phi component of E-field

W_rad = ( 1/(2*N0) ) .* (absE_phisq); %average power density
U = r^2 .* W_rad; %radiation intensity

Prad=sum(sum(U.*sin(THETA)*dth*dph)); %numerical computation of the radiated power
D=4*pi*max(max(U))/Prad; %linear directivity calculation
D_db=10*log10(D); %directivity in dB

direct_container_dB(index) = D_db; %directivities for varying the cylinder's
radius collected here

index = index + 1;

end

figure(1)
plot(bend_angles_degrees ,direct_container_dB, 'k-', 'LineWidth', 1.5);
ylabel('Directivity (dBi)', 'Interpreter','latex');
xlabel('Bending Angle $\Phi$ $(^\circ)$', 'Interpreter','latex');

```

```

xlim([0 240]);
grid on;

hold on

w_prime = w;
g = ( w - w_prime ) ./ 2 ;

e_reff = ( (er+1)/2 ) + ( (er-1)/2 )*( 1+ 12*(h/w))^(1/2); %effective dielectric
constant

l_ext = ( 0.412*(h)) * ((e_reff +0.3) * ( (w/h) + 0.264 )) ... ,
/ ((e_reff -0.258) * ( (w/h) + 0.8 )) ; %length extension due to fringe fields

l = ( 1 / (2*f*sqrt(e_reff)*sqrt(u0*e0)) ) - 2*l_ext ; %length between apertures
r = ( 2/(lambda) ) * (l^2); %farfield distance

%The upper and lower limits of theta and phi in degrees are defined.
%The upper limit for theta defines that radiation is propagating normal
%to the patch in the upper hemisphere
Theta_L=0;
Theta_U=90;
Phi_L=0;
Phi_U=360;

%Creating the meshgrid
theta=(Theta_L:Theta_U)*pi/180; % convert degrees to radians
theta=theta-0.0001;
phi=(Phi_L:Phi_U)*pi/180; % convert degrees to radians
phi=phi-0.0001; % term added since first element cannot be zero.

%trailing edge incrementation for theta and phi
dth=theta(2)-theta(1);
dph=phi(2)-phi(1);

[THETA,PHI]=meshgrid(theta,phi);

%antenna E-plane bending implementation for angles between 30 and 240
%degrees.
bend_angles_degrees = linspace(1,240,100);
bend_angles_rad = (pi/180) .* bend_angles_degrees; %convert to radians

radius_sweep = 1./ bend_angles_rad; %convert bending angles to equivalent radius
no_sweep = numel(radius_sweep); %number of elements to prepare sweep
direct_container_dB = zeros(no_sweep,1); %preallocated directivity container

index = 1;
for radius = radius_sweep

d = h/((pi^2).*radius_sweep(index));

X = (k*w/2)*cos(THETA);
w_fun1 = w*sincX(X);

Y = (k*w_prime/2)*cos(THETA);
w_fun2 = w_prime*sincX(Y);

Z = (k*g/2)*cos(THETA);
w_fun3 = g*sincX(Z);

gamma = k.*radius_sweep(index).*sin(THETA);

csf = zeros(size(THETA));
csf2 = zeros(size(THETA));

for n = 1:200 %computation of cylindrical space factor

```

```

        csf = csf + ((
(1i).^(n)./besselh(n,2,gamma)).*cos(n.*1./2*radius_sweep(index)).*cos(n.*PHI) );
        csf2 = csf2 + (( (1i).^(n)./besselh(n,2,gamma)).*cos(n.*PHI) );

end

E_phi_T = -d*(exp(-1i*k*r)./r) .* (w_fun1 + w_fun2 + 2*w_fun3)...,
.* ( 1./besselh(0,2,gamma) + 2 .* csf ).* (1./besselh(0,2,gamma) + 2 .* csf2 )
; %computing phi component of E-field

absE_phisq = abs( E_phi_T).^2; %absolute squared of phi component of E-field

W_rad = ( 1/(2*N0) ) .* (absE_phisq); %average power density
U = r^2 .* W_rad; %radiation intensity

Prad=sum(sum(U.*sin(THETA)*dth*dph)); %numerical computation of the radiated power
D=4*pi*max(max(U))/Prad; %linear directivity calculation
D_db=10*log10(D); %directivity in dB

direct_container_dB(index) = D_db; %directivities for varying the cylinder's
radius collected here

index = index + 1;

end

plot(bend_angles_degrees ,direct_container_dB, 'k--', 'LineWidth', 1.5);
% hold on
%
% full wave simulation results used to compare
% bend_ang = [0 30 60 90 120 150 180]; %bending angles investigated in full-wave
simulation analysis
% dirH = [7.74 7.36 6.98 6.63 6.35 6.07 5.86]; %full-wave directivities computed
for H-plane bending
% dirE = [7.74 8.01 8.16 8.08 7.81 7.38 6.79]; %full-wave directivities computed
for E-plane bending
%
%
% plot(bend_ang ,dirH, 'k-.', 'LineWidth', 1.5);
% hold on;
% plot(bend_ang ,dirE, 'k:', 'LineWidth', 1.5);
% legend({'Cylindrical Aperture Array Model', 'Simulated H-plane Bending',
'Simulated E-plane Bending'}, 'Location', 'Southeast', 'Interpreter','latex');
legend({'TPA E-plane Bending', 'RPA E-plane Bending' }, 'Location', 'Southeast',
'Interpreter','latex');
% legend({'CAAM TPA E-plane Bending', 'Simulated H-plane Bending', 'Simulated E-
plane Bending'}, 'Location', 'Southeast', 'Interpreter','latex');

function out = sincX(x)

    out = sin(x)./(x);
    if x == 0
        out = 1;
    end

end

end

```

A.5 Cylindrical-AAM to compute far-field patterns for the TPA

```

clc

format long;
close all

% DEFINED UNITS
meters      = 1;
seconds     = 1;
hertz       = 1/seconds;
millimeters = 1e-3 * meters;
gigahertz   = 1e9 * hertz;
seconds     = 1;

% CONSTANT DEFINITIONS
c0 = 299792458 * meters/seconds;
e0 = 8.854187812813e-12 * 1/meters;
u0 = 1.256637062121e-6 * 1/meters;
N0 = 376.7303136686;
er = 1.6; %relative permittivity of patch substrate

h = 2.8*millimeters; %height of aperture

f = 2.4*gigahertz; %resonant frequency
lambda = c0/f; %wavelength
k = (2*pi)/(lambda) ; %wavenumber

% Computation of the patch antennas width and length and distance to farfield
w = (c0/(2*f))*sqrt(2/(er+1)); %Aperture width #1
w_prime = 5*millimeters; %Aperture width #2

%horizontal component of l_d
g = ( w - w_prime ) ./ 2 ;

e_reff = ( (er+1)/2 ) + ( (er-1)/2 )*( 1+ 12*(h/w))^(1/2); %effective dielectric
constant

l_ext = ( 0.412*(h)) * ((e_reff +0.3) * ( (w/h) + 0.264 )) ... ,
        / ((e_reff -0.258) * ( (w/h) + 0.8 )) ; %length extension due to fringe fields

l = ( 1 / (2*f*sqrt(e_reff)*sqrt(u0*e0)) ) - 2*l_ext ; %length between apertures
r = ( 2/(lambda) ) * (l^2); %farfield distance

% H-plane
PHI = 0;
THETA=(0:360)*pi/180; % convert degrees to radians

%antenna E-plane bending implementation
bend_angles_degrees = 30:30:240; %bending angles between 30 & 240 degrees
bend_angles_rad = (pi/180) .* bend_angles_degrees;

radius_sweep = 1./ bend_angles_rad;
no_sweep = numel(radius_sweep);
P_n_container = zeros(no_sweep,numel(THETA)); %H-plane container

index = 1;

for radius = radius_sweep

d = h/((pi^2).*radius_sweep(index));

X = (k*w/2)*cos(THETA);
w_fun1 = w*sinc(X);

```

```

Y = (k*w_prime/2)*cos(THETA);
w_fun2 = w_prime*sincX(Y);

Z = (k*g/2)*cos(THETA);
w_fun3 = g*sincX(Z);

gamma = k.*radius_sweep(index).*sin(THETA);

csf = zeros(size(THETA));
csf2 = zeros(size(THETA));

for n = 1:200 %computation of cylindrical space factor

    csf = csf + ((
(1i).^n)./besselh(n,2,gamma)).*cos(n.*1./2*radius_sweep(index)).*cos(n.*PHI) );
    csf2 = csf2 + (( (1i).^n)./besselh(n,2,gamma)).*cos(n.*PHI) );

end

E_phi_T = -d*(exp(-1i*k*r)./r) .* (w_fun1 + w_fun2 + 2*w_fun3)...,
.* ( 1./besselh(0,2,gamma) + 2 .* csf ).* (1./besselh(0,2,gamma) + 2 .* csf2 )
; %computing phi component of E-field

absE_phisq = abs( E_phi_T).^2; %absolute squared of phi component of E-field

W_rad = ( 1/(2*N0) ) .* (absE_phisq); %average power density
U = r^2 .* W_rad; %radiation intensity
P_n = 10*log10((U)./(max(U))); %normalized power pattern plot in dB

P_n_container(index,:) = P_n; %collected power pattern plots for different bending
angles

%Linear Plots
figure(1)
plot((180/pi)*THETA,P_n, 'LineWidth', 1.5); %H-plane
xlim([0 180]);
ylim( [-30 0]);
xticks(0:45:360);
ylabel('$ \mathrm{U}(\theta,\phi) / \mathrm{U}_{\max} $ (dB)', 'Interpreter', 'latex');
xlabel('$ \theta (^{\circ}) $', 'Interpreter', 'latex'); %H-plane
legend('Bending Angle$ = 30^{\circ}$' , 'Bending Angle$ = 60^{\circ}$' , ...,
'Bending Angle$ = 90^{\circ}$' , 'Bending Angle$ = 120^{\circ}$' , ...,
'Bending Angle$ = 150^{\circ}$' , 'Bending Angle$ = 180^{\circ}$' , ...,
'Bending Angle$ = 210^{\circ}$' , 'Bending Angle$ = 240^{\circ}$' , ...,
'Bending Angle$ = 270^{\circ}$' , 'Interpreter', 'latex'); hold on
grid on;

%Polar Plots
figure(2)
polarplot(THETA,P_n, 'LineWidth', 1.5);
rlim([-110 0])
legend('Bending Angle$ = 30^{\circ}$' , 'Bending Angle$ = 60^{\circ}$' , ...,
'Bending Angle$ = 90^{\circ}$' , 'Bending Angle$ = 120^{\circ}$' , ...,
'Bending Angle$ = 150^{\circ}$' , 'Bending Angle$ = 180^{\circ}$' , ...,
'Bending Angle$ = 210^{\circ}$' , 'Bending Angle$ = 240^{\circ}$' , ...,
'Bending Angle$ = 270^{\circ}$' , 'Interpreter', 'latex');
hold on

index = index + 1;
end

pax = gca;
pax.ThetaTickLabel = string(pax.ThetaTickLabel) + char(176);
% pax.ThetaZeroLocation = 'left';

```

```

% pax.ThetaDir = 'clockwise';
pax.RAxis.Label.String = 'Radiation Intensity (dB)';

% E-plane
THETA = pi/2;
PHI=(0:360)*pi/180;    % convert degrees to radians

P_n_container = zeros(no_sweep,numel(PHI));    %E-plane

index = 1;

for radius = radius_sweep

d = h/((pi^2).*radius_sweep(index));

X = (k*w/2)*cos(THETA);
w_fun1 = w*sincX(X);

Y = (k*w_prime/2)*cos(THETA);
w_fun2 = w_prime*sincX(Y);

Z = (k*g/2)*cos(THETA);
w_fun3 = g*sincX(Z);

gamma = k.*radius_sweep(index).*sin(THETA);

csf = zeros(size(THETA));
csf2 = zeros(size(THETA));

for n = 1:200 %computation of cylindrical space factor

    csf = csf + ((
    (1i).^n)./besselh(n,2,gamma)).*cos(n.*1./2*radius_sweep(index)).*cos(n.*PHI) );
    csf2 = csf2 + (( (1i).^n)./besselh(n,2,gamma)).*cos(n.*PHI) );

end

E_phi_T = -d*(exp(-1i*k*r)./r) .* (w_fun1 + w_fun2 + 2*w_fun3)...,
.* ( 1./besselh(0,2,gamma) + 2 .* csf ).* (1./besselh(0,2,gamma) + 2 .* csf2 )
; %computing phi component of E-field

absE_phisq = abs( E_phi_T).^2; %absolute squared of phi component of E-field

W_rad = ( 1/(2*N0) ) .* (absE_phisq); %average power density
U = r^2 .* W_rad; %radiation intensity
P_n = 10*log10((U)./(max(U))); %normalized power pattern plot in dB
P_n_container(index,:) = P_n; %collected power pattern plots for different
bending angles

%Linear Plots
figure(3)
plot((180/pi)*PHI,P_n, 'LineWidth', 1.5); %E-plane
xlim([0 180]);
ylim( [-30 0]);
xticks(0:45:360);
ylabel('$ \mathrm{U}(\theta,\phi) / \mathrm{U}_{\max}$ (dB)','Interpreter','latex');
xlabel('$\phi(^{\circ})$', 'Interpreter','latex'); %E-plane
legend('Bending Angle$ = 30^{\circ}$' , 'Bending Angle$ = 60^{\circ}$' , ...,
'Bending Angle$ = 90^{\circ}$', 'Bending Angle$ = 120^{\circ}$' , ...,
'Bending Angle$ = 150^{\circ}$', 'Bending Angle$ = 180^{\circ}$' , ...,
'Bending Angle$ = 210^{\circ}$', 'Bending Angle$ = 240^{\circ}$' , ...,
'Bending Angle$ = 270^{\circ}$', 'Interpreter','latex');
hold on
grid on;

```

```

%Polar Plots
figure(4)
polarplot(PHI,P_n,'LineWidth', 1.5);
rlim([-90 0])
legend('Bending Angle$ = 30^\circ$', 'Bending Angle$ = 60^\circ$', ...,
    'Bending Angle$ = 90^\circ$', 'Bending Angle$ = 120^\circ$', ...,
    'Bending Angle$ = 150^\circ$', 'Bending Angle$ = 180^\circ$', ...,
    'Bending Angle$ = 210^\circ$', 'Bending Angle$ = 240^\circ$', ...,
    'Bending Angle$ = 270^\circ$', 'Interpreter','latex');
hold on

index = index + 1;
end

pax = gca;
pax.ThetaTickLabel = string(pax.ThetaTickLabel) + char(176);
pax.ThetaZeroLocation = 'top';
pax.ThetaDir = 'clockwise';
pax.RAxis.Label.String = 'Radiation Intensity (dB)';

function out = sincX(x)

    out = sin(x)./(x);
    if x == 0
        out = 1;
    end

end

```

A.6 Cylindrical-Rectangular Cavity Model to Compute Resonant Frequencies

```

clear;
clc;

% DEFINE UNITS
meters      = 1;
millimeters = 1e-3 * meters;
seconds     = 1;
hertz       = 1/seconds;
gigahertz   = 1e9 * hertz;
degrees     = pi/180;

% CONSTANTS
c0 = 299792458 * meters/seconds;
e0 = 8.854187812813e-12 * 1/meters;
u0 = 1.256637062121e-6 * 1/meters;
N0 = 376.7303136686;

L = 49.4*millimeters; %Length of patch
W = 49.4*millimeters; %Width of patch

h = 2.8e-3; % 2.8 mm thickness

epsilon_r = 1.6; %dielectric constant of substrate
fr = 2.4*gigahertz; % resonant frequency for unbent antenna

%TM Mode(mli -> 011 for TM fundamental
m = 0;
l = 1;

thetasweep = linspace(0.1,240,10)./180*pi; % bending angle sweep in rads
no_sweep = numel(thetasweep); %number of elements for sweep

```

```

fres_container=zeros(no_sweep,1); %container collecting resonant frequencies for
varying bending angle
index = 1;

for theta = thetasweep

v = m*pi./(theta);

a_plus_h = L./(theta);
a = a_plus_h - h;

kmi_search_range = [getkmi(epsilon_r, fr,l,L), getkmi(epsilon_r, 2*fr,l,L)];
%Limits for kmi wavenumbr

kmi_min = min(kmi_search_range);
kmi_max = max(kmi_search_range);

%solving transcendental equation for TM or H-plane bending
determinant_TM =@(kmi) JM(v,kmi*a).*YM(v,kmi*a_plus_h) -
JM(v,kmi*a_plus_h).*YM(v,kmi*a) ;

%locating zeros of transcendental equation
%requires chebfun package
functiontozero = chebfun(determinant_TM,[kmi_min,kmi_max],'vectorize');
zerocrossing = roots(functiontozero,'norecursion');
fres = getfres(epsilon_r, zerocrossing,l,L);
kres = zerocrossing;

if isempty (zerocrossing)
    kres = 0;
    fres = getfres(epsilon_r, kres,l,L);
end

fres_container(index) = fres(1);
index = index + 1;

end

figure(1)
plot(thetasweep .* (180/pi) , fres_container * 1e-9, 'k--','LineWidth', 1.5);
ylabel('Frequency (GHz)', 'Interpreter','latex');
xlabel('Bending Angle  $\Phi$  ( $^\circ$ )', 'Interpreter','latex');
grid on;
hold on;

%TE Mode TE mode (mli -> 101 for TE fundamental)
m = 1;
l = 0;

fres_container2=zeros(no_sweep,1);
index = 1;

for theta = thetasweep

v = m*pi./(theta); %order of bessell function

a_plus_h = W./(theta); %radius from centre of annulus to ground plane + height of
substrate
a = a_plus_h - h; %radius from centre of annulus to ground plane

kmi_search_range = [getkmi(epsilon_r, 0.8*fr,l,W), getkmi(epsilon_r, 2*fr,l,W)];
%Limits for kmi wavenumbr

kmi_min = min(kmi_search_range);
kmi_max = max(kmi_search_range);

```

```

% solving transcendental equation for TE or E-plane bending
determinant_TE = @(kmi) (JMp(v, kmi*a) .* YMp(v, kmi*a_plus_h)) -
(JMp(v, kmi*a_plus_h) .* YMp(v, kmi*a)) ;

% locating zeros of transcendental equation
functiontozero = chebfun(determinant_TE, [kmi_min, kmi_max], 'vectorize');
zerocrossing = roots(functiontozero, 'norecursion');
fres = getfres(epsilon_r, zerocrossing, l, W); % extracting resonant frequency from
the zeros
kres = zerocrossing;

if isempty (zerocrossing)
    kres = 0;
    fres = getfres(epsilon_r, kres, l, L);
end

fres_container2(index) = fres(1);
index = index + 1;

end

plot(thetasweep .* (180/pi) , fres_container2 .* 1e-9, 'k-', 'LineWidth', 1.5);
xlim([0 240])
xticks([0 30 60 90 120 150 180 210 240])
legend('H-Plane Bending', 'E-Plane Bending');

function fres = getfres(epsilon_r, kmi, l, b)

    eps0=8.854187817e-12;
    mu0=4e-7*pi;

    fres = (1/(2*pi*sqrt(epsilon_r*eps0*mu0))).*...
        sqrt(kmi.*kmi + (l*pi/(b))*(l*pi/(b)));

end

% function to get kmi from resonant frequency
function kmi = getkmi(epsilon_r, fres, l, b)

    eps0=8.854187817e-12;
    mu0=4e-7*pi;

    kmi = sqrt( ...
        (fres*(2*pi*sqrt(epsilon_r*eps0*mu0))).^2 ...
        - (l*pi/(b))*(l*pi/(b)) ...
    );

end

% Bessel functions and Hankel function with their derivatives
function out = JM(m, x) % besel function J order m
    out = besselj(m, x);
end

function out = JMp(m, x) % besel function derivative J order m
    out = 0.5.*(besselj(m-1, x)-besselj(m+1, x));
end

function out = YM(m, x) % besel function Y order m
    out = bessely(m, x);
end

function out = YMp(m, x) % besel function Y derivative order m
    out = 0.5.*(bessely(m-1, x)-bessely(m+1, x));
end

```

B. Measurements

B.1 Measured return loss for the fabricated RPA and TPA

This section showcases the prototypes of the RPA and TPA that were developed. A NanoVNA-F V2 was used to characterize the antennas return loss.

The antenna prototypes were fabricated using copper foil with a conductive adhesive placed on a cotton substrate. The dimensions for each antenna are based on the simulated results in Table 4.3 using the frequency domain solver.

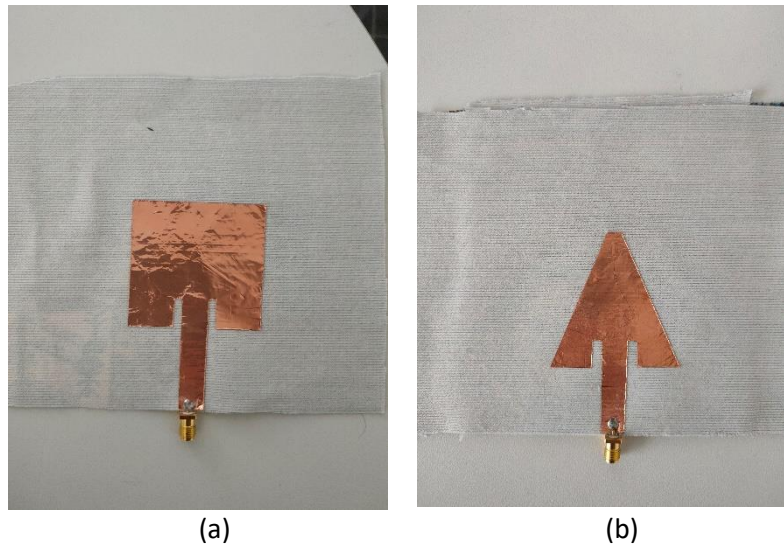


Figure B.1: Photographs of the fabricated prototype for the (a) RPA and (b) TPA.

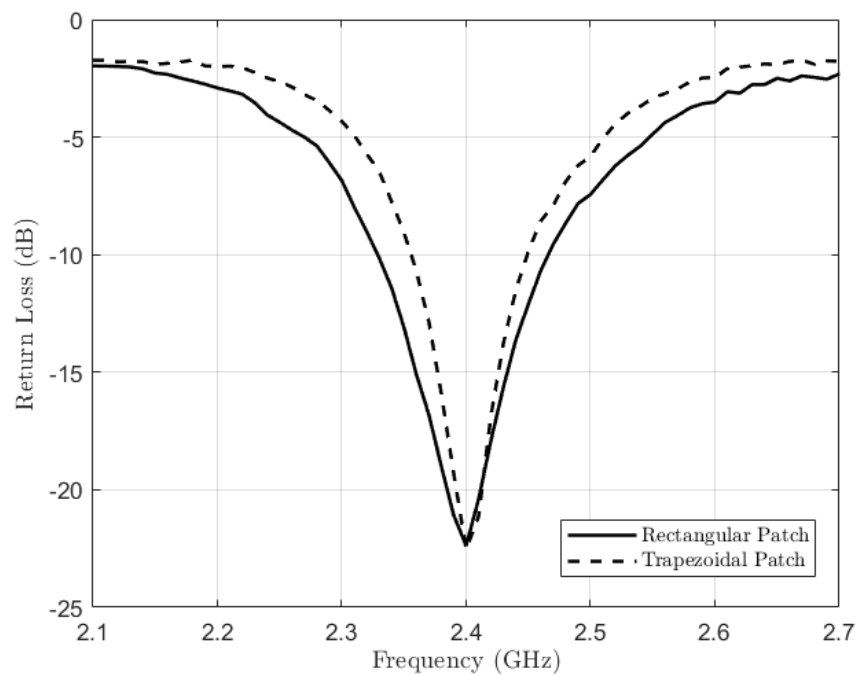
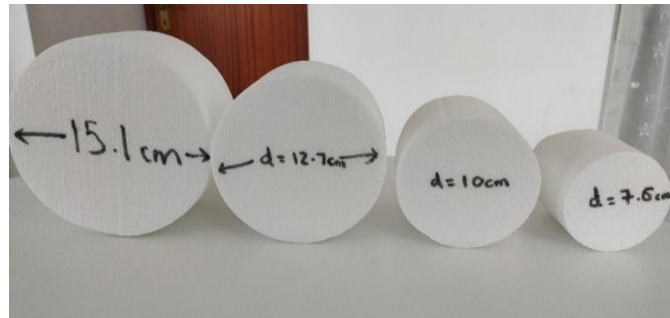


Figure B.2: Measured S_{11} comparing RPA with the TPA.

It was found that the chosen cotton fabric substrate did not have similar dielectric properties to the sample used for the full-wave simulation analysis. To correct for this to obtain a resonance at 2.4 GHz the fabric thickness for both antenna prototypes were increased to 4 mm. The measured return loss or S_{11} is shown in Figure B.2 which shows a bandwidth reduction for the TPA. This was also shown in the full-wave simulation analysis however, the measured bandwidths supersede the results obtained using the full-wave simulation solver. The measured -10 dB bandwidth for the RPA is 5.72 %, while for the TPA it is 3.91%. Despite these differences, miniaturization of the radiating element is achieved with the TPA radiating at the same frequency of 2.4 GHz.

B.2 Implementing Antenna E-plane Bending

E-plane bending is investigated using the prototypes built for the RPA and TPA. The antennas are mounted on a cylindrical polystyrene fixture of various diameters to investigate bending. Four cylindrical fixtures are used with a diameter of 7.6 cm, 10 cm, 12.7 cm and 15.1 cm as shown in Figure B.3(a). These correspond to antenna bending angles Φ of 166° , 126° , 99° and 83.5° .



(a)



(b)

Figure B.3: Incorporation of E-plane bending using the polystyrene fixtures.

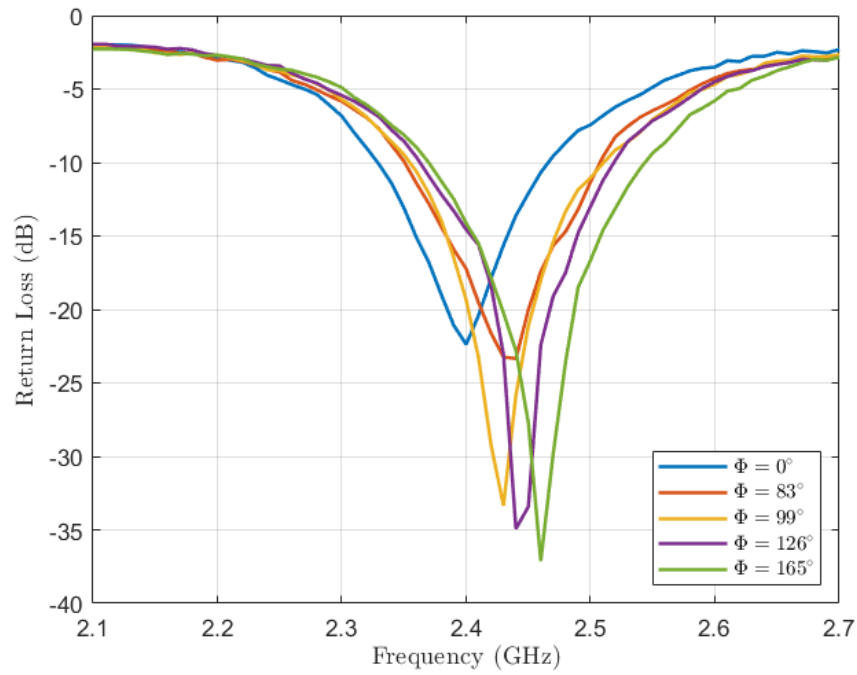


Figure B.4: Measured S_{11} for E-plane bending the RPA for various bending angles Φ .

Antenna Type	Bending Type	Bending Angle (°)	f_{res} (GHz)	-10 dB Bandwidth (%)
RPA	None	0	2.40	5.72
RPA	E-Plane	83	2.43	6.47
RPA	E-Plane	88	2.44	6.35
RPA	E-Plane	126	2.44	6.39
RPA	E-Plane	165	2.46	7.07

Table B.1: Resonant behaviour for E-plane bending the RPA over the cylindrical fixtures.

The measured return loss is shown in Figure B.4 for the RPA bent over various E-plane bending angles. The results show an increase in the resonant frequency for increasing bending angles. Also, it is shown that antenna's bandwidth increases for increasing E-plane bending angles.

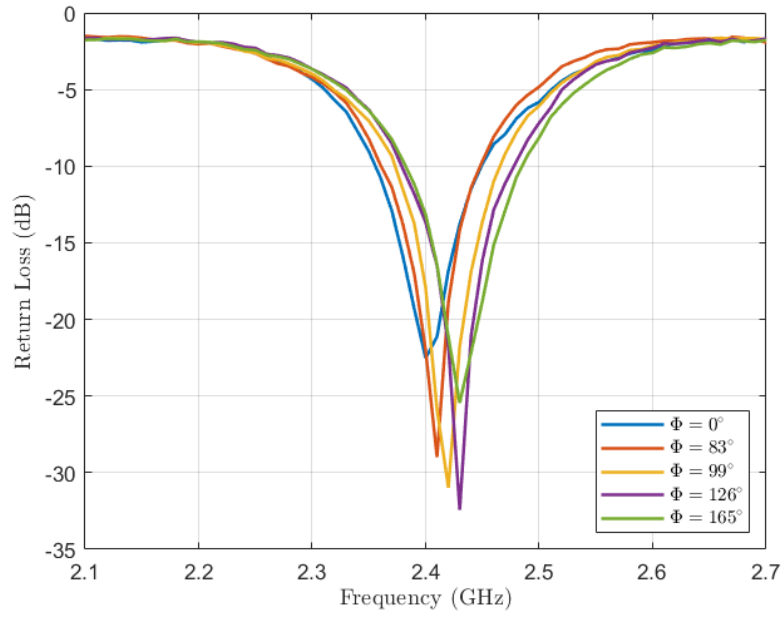


Figure B.5: Measured S_{11} for E-plane bending the TPA for various bending angles Φ .

Antenna Type	Bending Type	Bending Angle (°)	f_{res} (GHz)	-10 dB Bandwidth (%)
TPA	None	0	2.40	3.91
TPA	E-Plane	83	2.41	3.63
TPA	E-Plane	88	2.42	3.81
TPA	E-Plane	126	2.43	4.07
TPA	E-Plane	165	2.43	4.24

Table B.2: Resonant behaviour for E-plane bending the TPA over the cylindrical fixtures.

The measured return loss for the TPA is shown in Figure B.5 which shows that less resonant frequency detuning occurs which agrees well with the simulated results in Chapter 5.

Comparing the values obtained from the measured results to the simulated results is difficult due to the antennas not being structurally the same since adjustments were made by increasing the substrates height. The dielectric characterization of the fabric used for measurements is not the same as the sample used for the full-wave simulations. However, the trend of increasing resonant frequency is shown to be less for the TPA which agrees well with the simulations. The TPA shows a decrease in the bandwidth when compared to the RPA, however both the RPA and TPA for bending angles of up to 165° contain 2.4 GHz as part of the -10 dB threshold. Another observation was that the measured bandwidth increases for increasing bending angles for both the RPA and TPA. The full-wave simulations showed the opposite behaviour.

C. Simulation Results

This section consists of the full-wave simulation results for the TPA and TPA-MF simulated in the presence of the cubic tissue model, cylindrical tissue model, homogenous phantom, and FEM human tissue model. The cylindrical type models incorporate antenna bending whereby E and H-plane bending are investigated. The equivalent antenna simulations performed in free space are added for comparison.

Antenna Type	Bending Type	Bending Angle (°)	Tissue Model Configuration	f_{res} (GHz)	Bandwidth (MHz)	D (dBi)	ϵ_0 (%)	G_{re} (dBi)	SAR 10g (W/Kg)
TPA	None	0	Free Space	2.400	56.5	7.74	47.1	4.47	-
TPA	None	0	Cubic	2.399	44.3	8.44	40.9	4.57	0.135
TPA	E-Plane	110	Free Space	2.415	56.1	7.93	46.7	4.62	-
TPA	H-Plane	110	Free Space	2.402	58.8	6.45	53.2	3.82	-
TPA	E-Plane	110	Cylindrical	2.416	42.8	8.49	47.4	4.72	0.197
TPA	H-Plane	110	Cylindrical	2.402	49.3	6.75	47.8	3.56	0.193
TPA	E-Plane	110	Homogenous Phantom	2.417	41.4	8.63	43.0	5.00	0.101
TPA	H-Plane	110	Homogenous Phantom	2.403	48.9	6.60	50.0	3.57	0.232
TPA	E-Plane	90	Free Space	2.413	55.9	8.08	46.4	4.75	-
TPA	H-Plane	90	Free Space	2.402	58.8	6.63	53.3	3.90	-
TPA	E-Plane	90	FEM Human	2.414	42.0	8.31	42.0	4.53	0.335
TPA	H-Plane	90	FEM Human	2.402	46.8	6.98	48.2	3.83	0.0721

Table C.1: Performance parameters of the TPA in the presence of various tissue models using the frequency domain solver in CST.

Antenna Type	Bending Type	Bending Angle (°)	Tissue Model Configuration	f_{res} (GHz)	Bandwidth (MHz)	D (dBi)	ϵ_0 (%)	G_{re} (dBi)	SAR 10g (W/Kg)
TPA-MF	None	0	Free Space	2.400	70.6	8.70	54.0	5.92	-
TPA-MF	None	0	Cubic	2.393	73.7	9.72	48.3	6.56	0.105
TPA-MF	E-Plane	110	Free Space	2.431	77.5	7.08	55.0	4.51	-
TPA-MF	H-Plane	110	Free Space	2.410	75.9	8.07	58.3	5.72	-
TPA-MF	E-plane	110	Cylindrical	2.428	67.4	8.02	46.7	4.72	0.202
TPA-MF	H-plane	110	Cylindrical	2.416	87.8	8.06	55.0	5.47	0.183
TPA-MF	E-plane	110	Homogenous Phantom	2.428	70.4	7.53	50.2	4.54	0.341
TPA-MF	H-plane	110	Homogenous Phantom	2.409	81.6	8.40	55.7	5.87	0.118
TPA-MF	E-Plane	90	Free Space	2.430	74.4	7.55	54.5	4.91	-
TPA-MF	H-Plane	90	Free Space	2.411	76.2	8.15	55.9	5.63	-
TPA-MF	E-plane	90	FEM Human	2.425	71.9	8.49	48.5	5.35	0.526
TPA-MF	H-plane	90	FEM Human	2.410	76.9	8.44	55.9	5.91	0.0385

Table C.2: Performance parameters of the TPA-MF in the presence of various tissue models using the frequency domain solver in CST.

Antenna Type	Bending Type	Bending Angle (°)	Tissue Model Configuration	Mesh Cells (Tetrahedrons)	Peak Memory Used (GB)	Total Simulation Time (s)
TPA	None	0	Free Space	88778	1.181	86
TPA	None	0	Cubic	362072	15.67	892
TPA	E-Plane	110	Free Space	69467	3.25	237
TPA	H-Plane	110	Free Space	89336	1.45	221
TPA	E-plane	110	Cylindrical	359098	6.77	1513
TPA	H-plane	110	Cylindrical	418454	15.35	1555
TPA	E-plane	110	Homogenous Phantom	713398	12.41	2515
TPA	H-plane	110	Homogenous Phantom	632249	12.38	2372
TPA	E-plane	90	Free Space	85001	3.89	253
TPA	H-plane	90	Free Space	85609	1.41	208
TPA	E-plane	90	FEM Human	900378	19.63	4237
TPA	H-plane	90	FEM Human	903605	20.16	4407
TPA-MF	None	0	Free Space	187056	9.17	317
TPA-MF	None	0	Cubic	202718	9.39	726
TPA-MF	E-plane	110	Free Space	531877	4.45	305
TPA-MF	H-plane	110	Free Space	180147	9.20	415
TPA-MF	E-plane	110	Cylindrical	484217	36.72	3648
TPA-MF	H-plane	110	Cylindrical	256623	23.17	2473
TPA-MF	E-plane	110	Homogenous Phantom	723611	14.81	4450
TPA-MF	H-plane	110	Homogenous Phantom	477974	51.86	4128
TPA-MF	E-plane	90	Free Space	942538	8.10	609
TPA-MF	H-plane	90	Free Space	182558	9.27	414
TPA-MF	E-plane	90	FEM Human	512736	42.53	4463
TPA-MF	H-plane	90	FEM Human	510415	44.82	5555

Table C.3: Computational costs using the frequency domain solver.

D. Human Tissue Material Definitions

This section provides the material definitions provided in CST's Bio Models extension for human tissue involved in the antenna simulations.

D.1 Cubic and Cylindrical Tissue Model

The cubic and cylindrical tissue models are the simplest tissue model configurations which consist of four different tissue layers. This includes a layer of skin, fat, muscle, and bone. The material definitions are shown in Table D.1.

Tissue Type	ρ (kg/m ³)	κ (W/K/m)	cp (J/K/kg)	α (m ² /s)
Skin	1100	0.3	3500	7.79e-08
Fat	900	0.2	2500	8.89e-08
Muscle	1080	0.5	3500	1.32e-08
Bone	1850	0.4	1300	1.66e-07

Table D.1: Material property definitions in CST Microwave Studio for human tissue in the cubic and cylindrical tissue model.

where ρ is the material density, κ is the thermal conductivity, cp is the specific heat capacity, and α is the thermal diffusivity. These parameters describe the thermal properties associated with human tissue and are important in the computation of SAR and characterize how a material reacts to a temperature change. The dielectric properties for each tissue in the cubic and cylindrical tissue model is shown in Figure D.1 and Figure D.2.

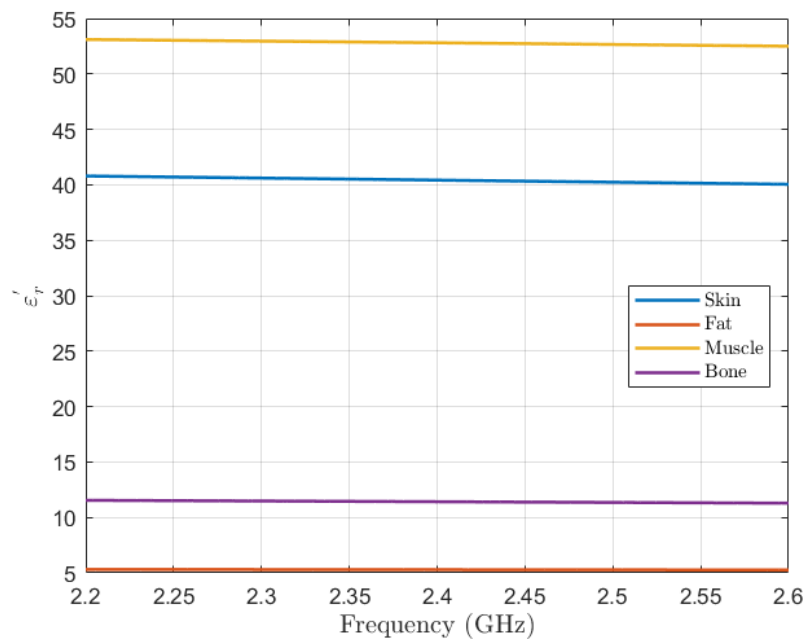


Figure D.1: Comparison of ϵ'_r for the layers of tissue used in the cubic and cylindrical tissue models.

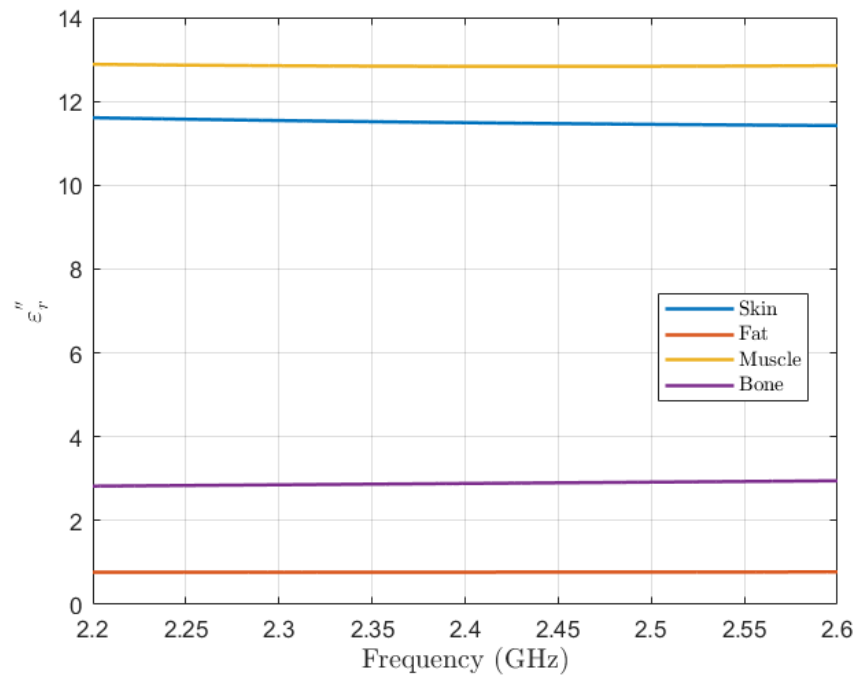


Figure D.2: Comparison of ϵ_r'' for the layers of tissue used in the cubic and cylindrical tissue models.

D.2 Homogenous Phantom Model

The Homogenous Phantom Model represents the human tissue as a single material averaging the properties of muscle and fat. The material definitions

Tissue Type	ρ (kg/m ³)	κ (W/K/m)	cp (J/K/kg)	α (m ² /s)
Muscle-Fat Average	1000	0.4	3000	1.33e-07

Table D.2: Material property definitions in CST Microwave Studio for Homogenous Phantom Model.

The dielectric properties for the muscle-fat averaged tissue in the Homogenous Phantom Model is shown in Figure D.3 and Figure D.4.

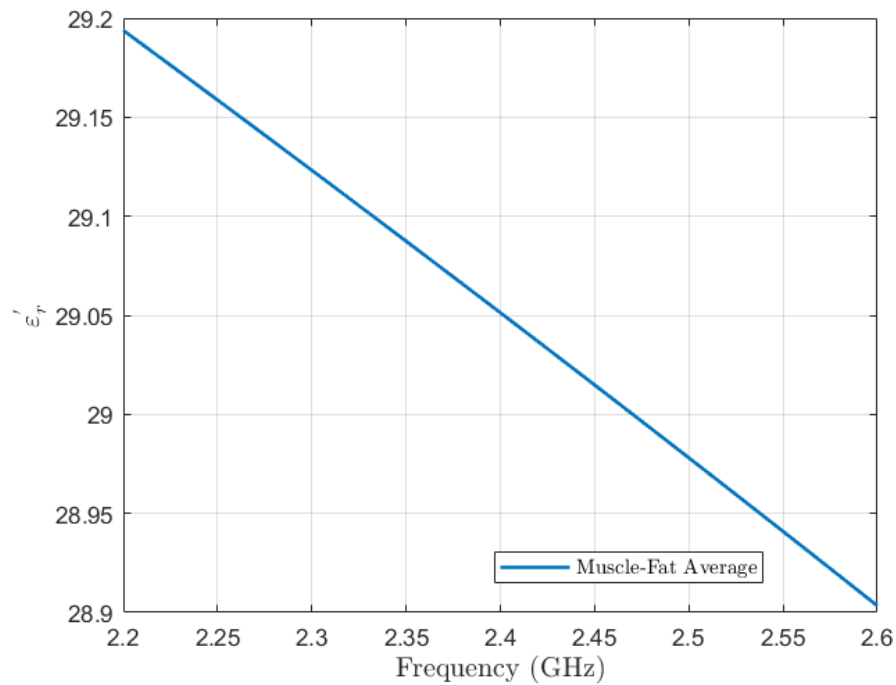


Figure D.3: ϵ'_r for the muscle-fat averaged tissue in the Homogenous Phantom Model.

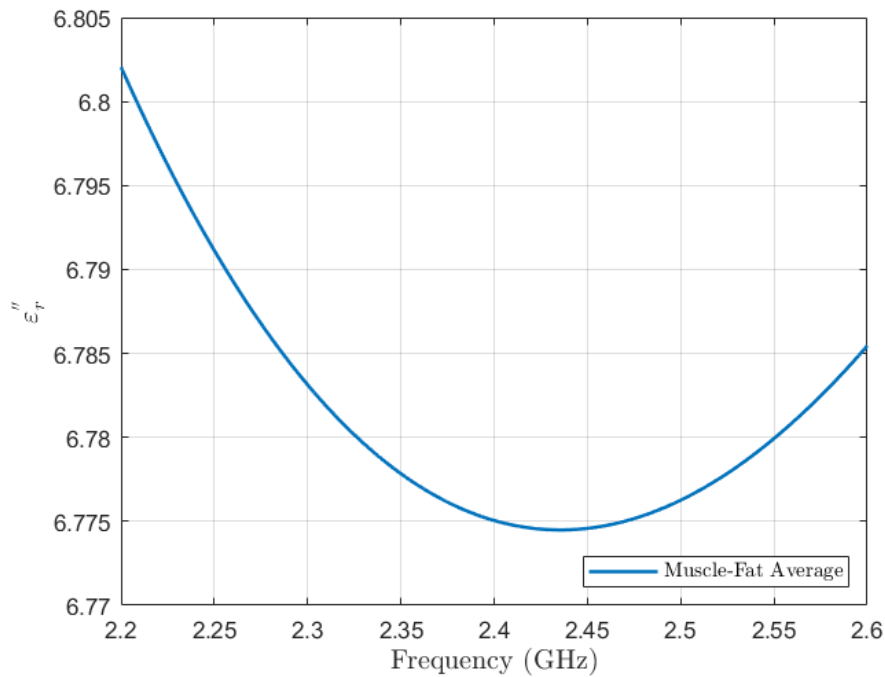


Figure D.4: ε_r'' for the muscle-fat averaged tissue in the Homogenous Phantom Model.

D.3 FEM Human Model

The FEM Human tissue model is the most detailed tissue model both geometrically and through its material definitions. The FEM Human tissue model within the bounding box of the simulations performed consists of skin, fat, muscle, muscle-fat averaged tissue and cortical bone definitions.

The material definitions are shown in Table D.3.

Tissue Type	ρ (kg/m ³)	κ (W/K/m)	cp (J/K/kg)	α (m ² /s)
Skin	1100	0.3	3500	7.79e-08
Fat	900	0.2	2500	8.89e-08
Muscle	1080	0.5	3500	1.32e-08
Muscle-Fat average	1000	0.4	3000	1.33e-07
Cortical Bone	2000	0.4	1300	1.54e-07

Table D.3: Material property definitions in CST Microwave Studio for the FEM Human tissue model.

The dielectric properties for each tissue in the cubic and cylindrical tissue model is shown in Figure D.5 and Figure D.6.

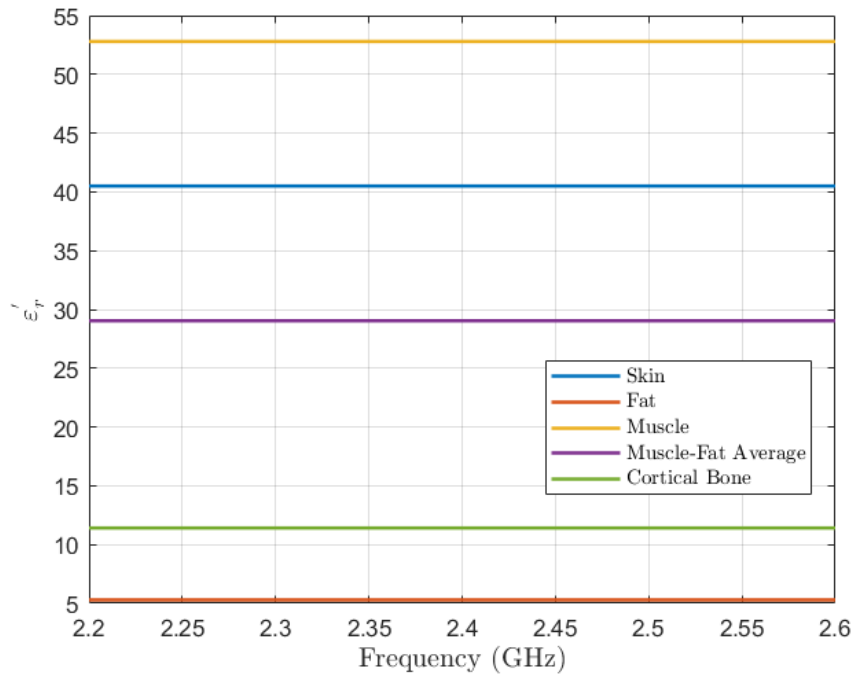


Figure D.5: Comparison of ϵ_r' for the layers of tissue used in the FEM Human model.

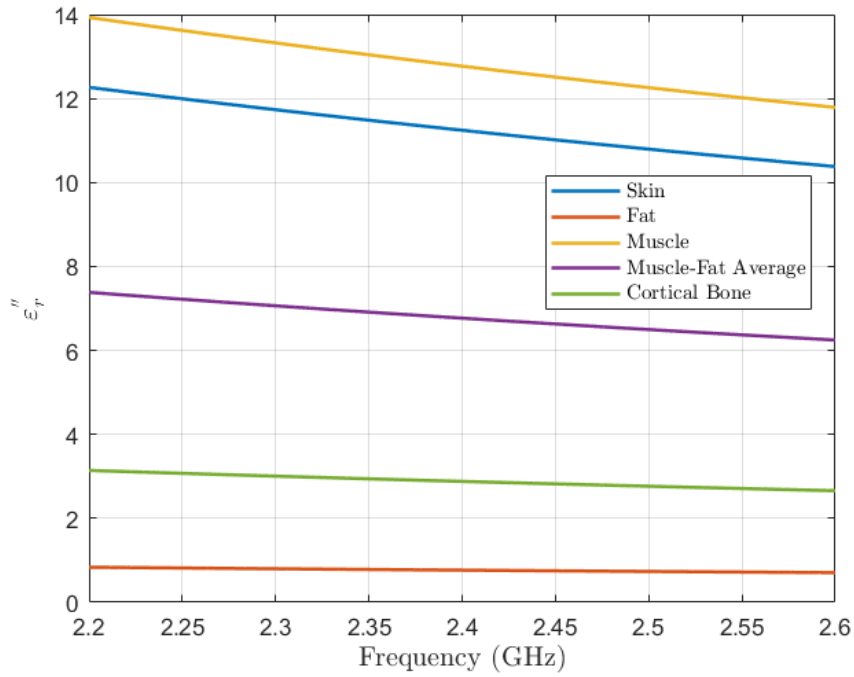


Figure D.6: Comparison of ϵ_r'' for the layers of tissue used in the FEM Human model.