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ANALYSIS AND DESIGN OF THE TWISTED LOOP ANTENNA TOPOLOGY FOR MOBILE COMMUNICATIONS

By

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A Doctoral Thesis

Submitted in partial fulfilment for the requirements for the degree of Doctor of Philosophy of Loughborough University

February 2004

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To my Family, the Cockroft Family, and in memory of Michelle Cockroft who sadly lost her battle with leukaemia in the duration of this research.
ABSTRACT

Keywords: Bifilar, Bluetooth, Quadrifilar, Parasitic Resonators, GPS, Dual-Polarisation

The handset product has been styled in successive years to reach more compact sizes and there has as a result been a reduction in volume available to house antennas; therefore size/performance trade-offs have had to become accommodated. Some of the issues antenna engineers are currently confronted with include; frequency shifting due to the antenna not being isolated from the handset, far field pattern deformation due to close proximity effects from the energy absorbing human tissues, distortion caused by noise from electronic components that share the handheld platform. What is required is antenna technology, which maintains a high enough performance despite the escalating restrictions imposed by the demands of the market. Research is performed on a twisted loop antenna topology that possesses an integral balun as part of its structure. Two rudimentary designs are utilised in the research, a simple bifilar structure that can be adapted for GSM, PCN, Bluetooth and W-LAN applications, and a quadrifilar helix structure for use in GPS. Both structures are based on existing industrial dielectric-loaded antenna structures but are modelled as novel air-loaded structures using a commercially available Method of Moments (MoM) electromagnetic simulator. In this fashion, the antennas could be generated quickly with low computational requirements. A parametric study is performed on the bifilar antenna structure to gain an enhanced understanding of the twisted loop topology. Once this understanding is achieved proposed modifications to the structure are implemented to improve the performance of the antenna. The main subject of improvement is the broadening of bandwidth as normally dielectric-loaded antennas have inherent narrow bandwidth. Any improvements made on the air-loaded structures could be tested on dielectric structures in future research. The most successful novel approach attempted to increase the bandwidth in the twisted loop structure was the insertion of parasitic helices to create a coupled multi-pole filter response. In conjunction with the work performed on the bifilar, an air-loaded GPS quadrifilar helix antenna was also modelled. A method for inducing circular polarisation is proposed and then by the insertion of parasitics into the quadrifilar helix design a novel dual-band dual-polarised antenna is presented. Finally measurements are made to demonstrate the advantageous properties the dielectric-loaded GPS antenna has over conventional GPS antennas.
PUBLICATIONS FROM THE RESEARCH


12. Leisten, O.P., Wingfield, A.P., ‘Miniature antennas with low proximity effects and filter response characteristics for Bluetooth and 3G applications’, IEE


ACKNOWLEDGEMENTS

My sincere gratitude goes out to Professor Yiannis Vardaxoglou for giving me the chance of pursuing a Doctorate in antennas for Mobile Telecommunications. His encouragement and guidance was always welcome and appreciated, and hope I get a further opportunity to work with Professor Vardaxoglou in the future. I would like to thank the electrical and electronic department of Loughborough University for my funding.

My greatest of thanks also goes to Dr. Oliver Leisten; his continued support of the way I think and his encouragement in pursuing and developing my ideas further was refreshing. His kind words, explanations, knowledge and motivation for my research was most appreciated. It is with great pleasure that I’ve been able to work with Dr. Leisten further, and have been given the chance to continue my research for further publications and ultimately for utilisation in industrial applications.

I would also like to thank Dr. M Jayawardene and Mr. P McEvoy for their help and invaluable contributions throughout my research and also concerning the knowledge of available electromagnetic numerical software packages. I wish to thank the antennas group at Loughborough university including Rob Seager, Gavin, Chin, Alford, Nico, Alex, Richard and Dave and wish them all the best of luck for the future in which ever direction they take.

Additional thanks also to Miss. Karen Violet for being a good friend, soul mate and for keeping me laughing during my research time. My gratitude to Mr. C. Collishaw, Mr. A. Woodall and Mr. Y. J. Knight also, for their invaluable advice at crucial times.

Finally and most importantly of all, I’d like to thank my own family and also my ‘foster’ family, the Cockroft family. Words cannot express my gratitude and love towards them and I wish them the best of health and happiness in the future. It is also with great regret that Michelle Cockroft passed away at the age of 19 during the course of my research having lost a battle with cancer; she will never be forgotten.
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CHAPTER 1.0

INTRODUCTION

1.0 Introduction to the Twisted Loop Structure

Recent years have seen a mass growth in the sales of mobile handsets around the world, and with each successive year the size of the handset has reached more compact size. This reduction in the size of the mobile handsets has meant a restriction in the available space in which to house its antenna. The modern market also demands multi-banding or even multi-tasking antenna properties; therefore the design of antennas for mobile telecommunications has become increasingly complex.

As telecommunication devices become more complex, manufacturers are demanding more novelty to keep customers attracted to their product. Current products are advertising Bluetooth or GPS capabilities in addition to having dual or triple-band properties. From the consumer’s standpoint, WAP phones have also become more commonplace as well as handsets with increasingly alluring features. Some may say these features, together with the complex technology and the increased usage of mobile handsets have come about at a price as over the last few years. Not only has the issue of SAR (Specific energy Absorption Rate) been repeatedly broached, the quality of reception in some of the more sophisticated handset models has been compromised.

From the SAR perspective, a number of computational simulations have shown a heating effect of various antennas from handsets [1-3] having on the side of the head where the handset is positioned. The debate as to whether the mobile handset antenna is harmful to the user continues because the application of the theory to prove the heating effects on the human head on a measurements platform is extremely complex. Nevertheless as energy is being absorbed by the consumer’s hand and head the antenna is not power efficient when in use, and makes the power consumption of the handset more demanding and therefore reduces battery life.
Although the debate on SAR and whether mobile phones pose a threat to human health has yet to be resolved, human interaction with mobile handsets is well documented. Antennas in mobile handsets are electrically connected to the handset such that when the handset is held, the antenna tunes to the appropriate resonant frequency as shown by Hill [4] in Figure 1. Consequently, as the human hand is not a standard size and varies from person to person, present antennas for mobile telecommunications cannot be tuned for optimal performance.

![Image](image_url)

Figure 1. Tuning Effects of the Hand and Head on the Handset Antenna

In addition to the antennas not being optimally tuned for resonance at the frequency for which they were designed, the near field of the antenna is known to interact with the field absorbing human hand as well as with it’s surroundings, this includes interaction with the mobile handset housing. The effects of the interaction with the user on the antennas far field pattern are investigated later in the thesis.

This means that from the modern day antenna designers’ perspective, what is required for mobile telecommunications to compete with the current market trend is an antenna with the following characteristics:

1. Small in size to account for the ever-decreasing size of antenna housing
2. A balanced feed system so the antenna has isolation from the handset to give the antenna a more robust antenna response as demonstrated by Leisten [5]
3. Low user proximity effects to minimise degradation of the far field pattern
4. Wide bandwidth
5. Low SAR

These are just some of the considerations required for the primary linearly polarised antenna in a mobile handset. The characteristics mentioned can also be applied to other linear polarised antennas for mobile telecommunications, in particular for usage on the Bluetooth application platform due to its diverse functionality. When the antenna is fitted onto cards for laptop computers for wireless connection, or onto wireless hands-free kits for mobile handsets, the need for isolation of the antenna from the device it’s connected to becomes apparent.

The containment of the near fields to reduce user or close body proximity effects is vital for these and other bluetooth applications. In the case where the bluetooth antenna is connected to a laptop computer, a further issue of common mode noise rejection needs to be confronted by the antenna designer. Without these considerations and properties built into the antenna the Bluetooth data-link connection between products will be unreliable.

The antenna topology proposed and researched for the linearly polarised applications of GSM, PCN and Bluetooth is the twisted loop antenna with integral balun, also known as the bifilar antenna as shown in Figure 2. The miniature dielectric-loaded bifilar antenna has been offered in the past as a low SAR and robust solution for mobile handset units. Its small size and containment of the near fields due to its high dielectric core makes it ideal for integration into small housings and for minimal close proximity effects from the user.

The bifilar antenna was invented by Leisten, who saw the opportunity to combine a twisted loop antenna to a sleeve balun. This would create two main loop modes of resonance each having a $360^\circ$ path length. The first resonant loop would be balanced and have complete isolation. The second resonant loop would not use the balun sleeve to choke to the currents, but would act as a path for the currents on which to propagate. Therefore much of the work performed with the bifilar has concentrated on the first resonant mode which uses the balun sleeve as a choke.
The loop antenna in the antennas topology was twisted as Leisten realised that an electromagnetic null would be formed which would be located near the users head when resonating in the balanced mode. The introduction of the dielectric core meant he could miniaturise the antenna and have further control of the antennas fields, leading to antenna suited to mobile telecommunications.

The twisted loops integral balun means the antenna is isolated from the device to which it is attached. The balun prevents currents flowing down the outer conductor of the feed cable, which would normally couple the antenna to the handset therefore providing isolation from the handset. When functioning optimally the antenna is therefore much less susceptible to the frequency shifting and far field pattern impairments experienced when human tissue holds the handset in proximity to a human head. These effects have been highlighted in the past for other personal telephone antennas; R Hill [4] also found 12dB, 15dB and 9dB impairments in the far field when a rod monopole, helical monopole and patch antenna in the PCN band were measured in normal talk positions in proximity to a human head.

![Figure 2. Proposed Twisted Loop Structure with Integral Balun](image)
The benefits of using the dielectric-loaded twisted loop structure on a handset platform have been previously published by Leisten et al [7-16] and by Rosenberger [17]. These publications demonstrated SAR measurements as well as simulated results that highlighted the characteristic electromagnetic null of the far field pattern. More recent publications on the dielectric loaded bifilar have also demonstrated the low SAR characteristics but just as importantly emphasised the importance of isolating the antenna from the handset [5].

GSM/PCN and Bluetooth applications aren’t the only mobile telecommunication functions that have necessitated additional deliberation when being designed. The handset product has been styled in successive years to reach more compact size and there has as a result been a reduction in volume available to house antennas. Whilst size/performance tradeoffs in the terrestrial system have accommodated the consequent reductions in radiation efficiency, there is no available way to mitigate this loss for the GPS system as it uses signals that are transmitted from satellites.

However with the introduction of the Federal Communications Commission (FCC) US E-911 emergency services position location by Global Positioning System (GPS) in October 2001 and with the impending introduction of E-112 in Europe, the impairment in radiation efficiency became a more critical issue. As the antennas remit for GPS is the reception of satellite signals there is currently no available way to mitigate for these impairments or to easily increase the signal strength.

GPS reception requires differences in the antenna characteristics from those exhibited by the main GSM/PCN antenna in the handset. Instead of a far field pattern directed at base stations, the GPS receivers are trying to obtain signals from satellites above the horizon. Just as importantly the signals being received are right hand circularly polarised and therefore for the GPS antenna also needs to be right hand circularly polarised for optimal reception. This means the linearly polarised antennas that are currently used for GSM in handsets would not have the optimum antenna response to provide a stand-alone positioning service.
Nevertheless to meet the deadlines set by the FCC for US E-911 emergency location some manufacturers have adopted a monopole antenna to function as their main network antenna and as the GPS antenna. A recent study performed by Leisten and Wingfield [6] has highlighted the potentially detrimental effects of utilising a linear polarised antenna with a single ended feed as a GPS antenna. What is needed is an antenna technology that maintains a high enough performance despite occupying a small volume, being in close proximity to energy absorbing human tissues of the user and in the presence of noise and distortion from the radios that share the handset platform. An antenna with a predictable and robust far field pattern despite its environment is especially crucial when utilised on a life saving platform. Many of the properties mentioned are common with those required for Bluetooth, GSM and PCN applications.

One of the reasons for the research being performed on the twisted loop structure, apart from the benefits it offers as a telecommunications antenna, is that by the introduction of an additional orthogonal loop to the structure as shown in Figure 3., the antenna can be made to radiate with circular polarisation. This gives a natural progression from the bifilar to the quadrifilar, although historically the quadrifilar came first. The quadrifilar helix, which is formed by a correctly phased orthogonal loop, was initially pioneered by Kilgus [18-21] who identified its far field characteristics. The characteristic cardioid pattern of the quadrifilar helix has good zenith gain and a broad beamwidth that makes it ideallic for GPS applications.

![Figure 3. Dielectric Loaded Quadrifilar Helix Antenna](image-url)
The function of the dielectric loaded quadrifilar helix as a GPS antenna is enhanced by the integral balun that is common to both this structure and the twisted loop structure, which again provides isolation from the handset box. This isolation in conjunction with the high dielectric core gives the antenna a robust and predictable pattern as recently demonstrated by Leisten and Wingfield [6].

Unfortunately, although the miniature dielectric loaded twisted loop and quadrifilar helix antennas have advantages over the conventional telecommunication antennas, it has been shown that a number of modelling issues are encountered when simulating these antennas [22]. Therefore to simplify the modelling and reduce the computational time of the simulations air loaded geometries can be utilised for the analysis of the structures. This is particularly advantageous, as only in recent years has the advancements in computational memory and speed required for accurate simulation of the dielectric structures been accomplished. In conjunction with the advancements of computational power, improvements have also been seen in the electromagnetic solvers but nevertheless the twisted loop and quadrifilar structures remain a challenge for any simulation software.

1.1 Structure of the thesis

The objective of Chapter 2 is to introduce the reader to the twisted loop antenna structure and to develop an accurate air loaded model of the GSM antenna structure. Due to the complications that arise when introducing the dielectric core to the simulation model an electrically equivalent air-loaded structure is modelled. This model would allow fast and accurate simulation of potential modifications to the antenna structure, before utilising a package with dielectrically loaded capabilities or measurements taken on a real model.

The air-loaded simulation is modelled using a Method of Moment (MoM) called MiniNEC Professional for Windows (MiniNEC), which uses wires to build up the required geometry. For additional simplicity the coaxial feed system of the antenna structure is initially removed, this allows for just the antenna structure to be modelled as accurately as possible. The results generated by this model are then compared to a
model generated using an in-house Finite Difference Time Domain (FDTD) package. A coaxial cable is generated using the MiniNEC and inserted as the feed system for the twisted loop structure, this allows the effects of the feed system to be monitored. Models were then built for measurement and comparison with the simulated results.

Having defined a robust air-loaded twisted loop geometry, two further geometries are developed for Bluetooth and PCN applications to be used later in the research. Finally a modelling technique is introduced for the coaxial feed system to avoid modelling issues that have been encountered and highlighted in previous work [21]. These modelling issues are reviewed in a comparative study of some electromagnetic packages and their ability to model the twisted loop structure.

Chapter 3 then looks more closely at the near and far field properties of the bifilar and ways of modifying the structure to achieve the required antenna response. This analysis is initiated using a parametric study to understand the primary modes of the antenna. In one study the far field is observed for changes in the characteristic Low Specific Absorption Rate (SAR) pattern of the balanced mode when the twist in the loop structure is altered. Another study seeks to define the current propagation of the single ended mode more precisely by the analysis of the near fields. The far field of the balanced mode is then analysed, and the realisation of the characteristic far field pattern by the simulation of electrical and magnetic dipoles is attempted.

To complete the bifilar twisted loop analysis, before moving onto the GPS twisted loop structure, modifications are made to the structure increase its bandwidth. This included techniques such as making the helices into bowtie structures, which are known for their broadbanding capabilities, and the introduction of parasitics that would establish additional modes of resonance. As the simulations of the twisted loop with parasitics showed potential, a more intense investigation was undertaken.

Chapter 4 progresses to the air-loaded GPS twisted loop antenna, which utilises two twisted loops orthogonal to each other to form a circularly polarised quadrifilar helix. Once again an integral balun is present to give the antenna the advantages of isolation from the telecommunications device to which it is attached. It will be seen that the
modelling of the quadrifilar helix geometry, with a single feed system, requires an awareness of modelling techniques to generate the phase quadrature for the characteristic circularly polarised far field pattern as demonstrated by Kilgus [18-21]. Once the air-loaded quadrifilar geometry with integral balun has been constructed and its far field pattern displays circular polarisation, comparisons with measured results are then undertaken. Finally parasitics are introduced into the design in an attempt to broaden the bandwidth covered by the antenna response, by this method a novel dual-band dual-polarised antenna is exposed.

The advantages of the quadrifilar helix antenna with integral balun as a GPS antenna is investigated in Chapter 5 by means of measurements involving the dielectric-loaded antenna. Many publications have highlighted the benefits of the dielectric-loaded quadrifilar helix [6, 22-26] by means of theory; therefore measurements were performed to demonstrate this theory.

Conclusions and future work emanating from the work carried out from this thesis are presented in Chapter 6.

1.2 References


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CHAPTER 2.0

CYLINDRICAL AIR-LOADED BIFILAR ANTENNAS

2.0 Introduction

The bifilar antenna was invented by Leisten, who saw the opportunity to combine a twisted loop antenna to a sleeve balun. This would create two main loop modes of resonance each having a $360^\circ$ path length. The first resonant loop would be balanced and have complete isolation. The second resonant loop would not use the balun sleeve to choke to the currents, but would act as a path for the currents on which to propagate. Therefore much of the work performed with the bifilar has concentrated on the first resonant mode which uses the balun sleeve as a choke.

The loop antenna in the structure was twisted, as Leisten realised that an electromagnetic null would be formed which would be located near the users head when resonating in the balanced mode. The introduction of the dielectric core meant he could miniaturise the antenna and have further control of the antennas fields, leading to antenna suited to mobile telecommunications. Many of these concepts are given in more detail in the next two chapters.

This chapter forms the foundation of the air-loaded bifilar antenna simulation models presented in this thesis. These models are based on the more complex dielectric loaded bifilar antenna pioneered by Leisten [1, 2, 3, 4]. Most importantly, it develops simulation-modelling techniques crucial for the design and manufacturing of twisted loop antennas. It also facilitates the modification of current designs and development of future generations.

Two air-loaded twisted loop antennas are presented in this chapter to establish a working simulation model that forms the basis of much of the work carried out in this thesis. These are based on a scaled up model of the dielectric-loaded bifilar with
dielectric constant 36 that was originally designed for GSM applications. The first model is configured with the coaxial cable feed excluded from the design to maintain simplicity in the development of the antenna design. The second artwork inserts a specially designed coaxial cable into the twisted loop model, so the complete bifilar model is simulated.

The models are simulated separately so the main antenna structures can be rigorously tested for convergence towards accurate design solutions before the inclusion of a more complex feed system. This involved finding a compromise between accuracy in simulated results and simulation time, so an accurate model was developed that had good resolution but was not memory or time intensive. Once an accurate model has been established and verified, air-loaded models for PCN and Bluetooth applications will be generated for further modification and research.

In section 2.2 the twisted loop structure is introduced and motivations for using air-loaded structures and for using Method of Moments (MoM) software for the modelling are explained. The dimensions for the initial air-loaded design are presented in section 2.3, this topology was a very basic model compared to the more complex quadrifilar helix antenna, to be introduced later in the thesis, particularly with the cable excluded from the simulation for simplicity. The response of the simulated antenna is presented in terms of its return loss.

In order to justify the MoM air-loaded model presented in section 2.3 modelled using MiniNEC Professional for Windows, the antenna is then constructed using a Finite Difference Time Domain (FDTD) method in section 2.4 and the results are compared with those of the MoM design. Impedance curves and return loss plots are used for the comparisons.

To model the antenna accurately using the MoM package, a cable is introduced in section 2.5. The cable was modelled on its own at the outset so that the correct response was obtained without the added complication of the antenna structure. Different designs were attempted in trying to achieve an optimal coaxial cable design, and the method for calculating their characteristic impedance is described. Response
plots of various attempted cable models, as well as the formation of the final cable design can be seen in this section.

This cable is then attached to the air-loaded bifilar developed in section 2.2, and the response of the complete air-loaded design, including impedance and return loss measurements, are assessed. By modelling the cable separate to the antenna the complex geometry of the cable can be accurately modelled before being utilised as the conduit for antenna excitation. The importance of the cable to the antenna structure in its ability to excite resonant modes will be seen.

The MoM bifilar helix model with the cable excluded is then compared with the complete model where the cable was included. The impedance curves of the two models require embedding or de-embedding to make the comparisons possible. Far field radiation patterns of the modes of interest for the twisted loop antenna with cable will be displayed. It will be seen that for the balanced mode the far field pattern resembles that of a simple resonant dipole as suggested by the virtual dipole theory of the bifilar antenna.

Section 2.6 compares the results of the two air-loaded bifilar models with a practical air-loaded measurement. The material properties used to construct the bifilar model to be measured are recorded, and the response of the antenna when measured is plotted with the simulated models for comparison.

Section 2.7 makes practical use of the techniques established in the previous sections to develop an air-loaded Bluetooth model scaled to resonate around the centre frequency of the Bluetooth band 2448.5MHz. The Bluetooth band is 2400MHz – 2483.5MHz; although at the time of the research the upper limit was taken to be slightly higher. The dimensions of the simulation model are once again tabulated and the impedance curves and returns loss plots of the antenna are shown. Radiation patterns at the main modes of resonance will be displayed for analysis and for use later in the thesis. Having completed a twisted loop design for Bluetooth applications, another model will be developed for PCN applications with once again its response plots and far field patterns displayed.
The Bluetooth and PCN design are to be utilised later in the thesis when analysing the properties and theory of twisted loop antennas. Although only GSM, Bluetooth and PCN bifilar designs are covered in this chapter, the techniques developed can be used for other communication bands. It is noticeable that the scaling factors required in developing from dielectric to air-loaded models in the PCN and Bluetooth models are similar and this fact could either be used in future investigations, or in research that studies twisted loop models constructed using intermediate dielectric constants in their core loading.

Having shown that the simulated and measured results have good correlation, theory concerning the resonant modes is exploited as a means of simplifying the coaxial cable feed system further in section 2.8. This is a technique that could be useful in other modelling packages and eliminate some of the modelling issues associated with high resolution modelling. For example, without the elimination of the inner conductor when modelling a coaxial line would have stair-casing effects and very high resolution issues associated with it.

Whether air or dielectric loaded the twisted loop topology poses a modelling challenge for numerical simulation packages. Some of the modelling issues associated with the geometry when using different packages are detailed in section 2.9, this includes the high resolution necessary for accurate modelling of the coaxial feed system to avoid stair casing effects due to the curvature of the thin metallisation in the structure. Finally section 2.10 summarises the essence of chapter 2 as conclusions.

### 2.1 Air-Loaded Geometries

In order to simulate the previously documented Dielectric-Loaded Bifilar Helix Antenna (DLBHA) [1,2,3,4,5] in air-loaded format, the approach taken was to model simplified geometries and then progress on to the full antenna artwork. This strategy enabled an accurate simulation model to be developed in small steps; therefore making sure each step was numerically accurate before progressing to the next stage. This was done by the use of convergence testing; using very high resolution for the
modelling initially and then reducing the number of nodes and wires in the antennas definition in repeated simulations until accuracy is compromised. At this stage it would be known what resolution was required in the antennas structure before accuracy is lost.

Through the use of the simplified geometries for modelling the DLBHA and also by comparing the air-loaded design to the dielectric model to make sure the modelling was comparable, a way of modelling the twisted loop topology in a robust manner to generate fast results was established. It was anticipated that this approach might facilitate designing future generations of the dielectric-loaded model, as computation resources at this stage of this research not sufficiently powerful to achieve the resolution required for numerical accuracy.

As very few packages had the relevant dielectric-loading capabilities at this stage in the research, and those that did were memory intensive and took days to run, air-loaded geometries provided a good medium for modelling the twisted loop structures. It was interesting to assess whether the air-loaded models could provide a channel for fast and accurate modelling which would prove computationally efficient in memory and simulation time.

The main area of concern when modelling the twisted loop geometry is the attachment of the thin metallisation to the dielectric core as well as the modelling of the coaxial feed. Therefore excluding the dielectric core and the cable meant that only two media needed modelling, that of air and metal, making the simulation simplistic. This air-loaded geometry would still be based on the dielectric-loaded bifilar model so the theory for the modelling follows that of the DLBHA. The dielectric-loaded antenna can be broken down into four parts as shown in Figure 2.1.

![Figure 2.1 Dielectric-loaded bifilar helix antenna](image)
The DLBHA is manufactured using a thin metallisation printed onto a high dielectric ceramic core with a permittivity of 36. In essence the antenna is a $180^\circ$ twisted loop antenna balanced fed by a coaxial cable that passes axially through the centre of the ceramic to the top of the structure. The sleeve balun acts as a transition between single-ended and balanced drives in order that the antenna can be fed with balanced currents. This balance serves to provide isolation for the antenna from the handset or device its connected to, therefore when the device is held the antenna will not suffer detuning or far field pattern deformation.

When in operation the currents flow from the inner conductor of the feed point round a helical track, around the rim of the balun and back up the helical track that is geometrically opposite to the outer conductor of the coaxial cable. This resonant loop path is known as the balanced mode. The other dominant mode of the antenna is known as the single-ended mode. This is where the current flows round the helical tracks down the balun, across the balun base and back through the coaxial feed. These two modes can be seen more clearly defined in Figure 2.2.

![Figure 2.2 Balanced (a) and Single-ended Mode (b)](image)

To reduce the complexity of the design when modelling the antenna geometry, the dielectric core was initially excluded. The dielectric core with its high permittivity allows the antenna to have a small size with respect to its wavelength and contains many of the antennas near fields. Therefore modelling the dielectric core and the attachment of the metallisation to the ceramic puck requires much more consideration and software capability.
The other part of the structure that was excluded from the initial simulation runs was the coaxial cable with the ‘fish hook’ feed. This removal was necessary due to the small dimensions of the cable, as this would require high resolution modelling. The fishhook feed system is introduced later once the basic geometry for simulation had been established.

A Method of Moments (MoM) [6] commercial software package called MiniNEC Professional for Windows [7] was used for simulating the metallic structures. The MoM technique is well-documented but is reviewed later in Appendix A. It utilises wires connected together by nodes to build up the antenna geometries. Nodes are used to subdivide each wire into subsections for the MoM computation. Due to the simplicity of the software (without any facility for including dielectric materials) the simulation time would be fast compared to other available packages.

With the exclusion of the dielectric core and the hook feed for the preliminary design the antenna geometry had to be modified. The effect of excluding the dielectric core would be that miniaturisation would not transpire and the antenna would no longer be small compared to its free-space wavelength. For this reason a multiplicative scaling factor of \( \sqrt{\varepsilon} \) needed applying, i.e. a scaling factor of 6 for a relative permittivity of 36. When the cable is excluded from the geometry, the antenna is sourced at the top of the structure by the connection of the top tracks by a voltage gap, which is illustrated later in this chapter. One further consideration in the chapter is that the DLBHA helical tracks is made of thin copper tracks and not round wires as used in the air-loaded MoM simulations. Therefore, a well-documented thin strip to round wire approximation was used to account for this modification [8,9].

### 2.2 Preliminary Air-Loaded GSM Bifilar

The first air-loaded twisted loop model to be addressed was a single band PCN bifilar. The structure is basic and well behaved compared to other cylindrical loop based antennas and well documented in dielectric form therefore results could be readily corroborated for accuracy. When operating effectively in the balanced mode a virtual
dipole should be present between the helical wires halfway up the radiating section, and this would be recognisable in the far field radiation pattern.

To reduce complications in the structure further the dielectric core and the hook feed were removed, and the antenna geometry modified accordingly. The modified dimensions of the antenna can be seen in table 2.1.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Dimension</th>
</tr>
</thead>
<tbody>
<tr>
<td>Total Height</td>
<td>72.6 mm</td>
</tr>
<tr>
<td>Diameter</td>
<td>60.0 mm</td>
</tr>
<tr>
<td>Radiating Section Height</td>
<td>39.0 mm</td>
</tr>
<tr>
<td>Balun Height</td>
<td>33.6 mm</td>
</tr>
<tr>
<td>Track Width</td>
<td>4.0 mm</td>
</tr>
<tr>
<td>Pitch Angle</td>
<td>22.48°</td>
</tr>
</tbody>
</table>

Table 2.1 Dimensions of the air-loaded twisted loop

The helical tracks at the top of the antennas topology were connected together and then the antenna excited by a centralised simple voltage gap source as shown in Figure 2.3.

The bazooka balun of the antenna was constructed using a meshed wire structure to represent the continuous metallisation of the balun seen in the dielectric-loaded bifilar helix antenna. The balun structure was meshed using wire arcs connected together using straight wires. Through convergence testing, the parameters of which are
described later in the chapter, it was found that using two wires arcs to represent the balun rim and the base and then by connecting these wire arcs at every 30° gave a good meshed representation of the balun sleeve. The base of the balun was meshed every 60°, and an extra wire arc at half radii, using 12 straight wires, provided more definition. The balun structure can be observed in Figure 2.4.

![Figure 2.4 Balun Mesh Structure](image)

The wire arcs representing the balun base and rim consisted of 24 wires to give good circular definition to the balun structure. The technique of convergence testing used in the construction of the balun allowed for good definition of the balun without the use of excessive nodes that would add to the computational time. Although the structure was visually basic, convergence testing showed it gave numerically accurate representation of the balun. It was seen that if the structure of the balun was created too complex then errors escalated in the calculation of the antenna response. Therefore this method gave a computationally efficient structure.

The radiating section of the antenna structure was simpler in evaluation. To obtain good initial definition a three wire meshed structure was used for the helical wires in the radiating section to represent the helical tracks. A documented thin strip to cylindrical wire approximation calculated the wire radii [8,9]. Using the high resolution three wire meshed approach utilised large number of nodes so an equivalent single wire design was investigated and compared with the three wire meshed radiating section.

Twenty wires were used for each helix to give structural and visual accuracy. From Best [10] it is seen that only 12 wires needed for computational accuracy though
visually the resolution seems coarse. By using 20 wires for definition, this assured the numerical accuracy of the helices and gave good visualisation with a wire every $9^\circ$. Once again the single wire to strip width approximation was utilised for the radii of the wires. It was found that a single wire could be used in the construction of the radiating section with no change in the antennas response.

The convergence tests used to evaluate the geometries developed for the bifilar structure included the following considerations.

- Increasing the number of nodes on the wires used
- Varying number of wire arcs and straight wires in the balun sleeve
- Varying number of wire arcs and straight wires in the balun base
- Varying the number of wires representing the circular definition of the balun and radiating section
- Observing the effect of using single wires and three wires meshed in the definition of the radiating section

From the use of the convergence testing a basic air-loaded GSM bifilar design was developed that was easy to generate and quick to simulate. The complete initial structure with single wire radiating section can be seen below in Figure 2.5. The geometry gave an accurate response, in the frequency range 800MHz to 3.5GHz, when impedance and return loss plots were compared to further simulated and measured results as will be revealed in the following sections.

![Figure 2.5 Initial Bifilar Structure](image)
2.3 FDTD Method Air-loaded Design and Comparison

As a comparison for the Method of Moments simulation of the preliminary air-loaded design without hook feed, a parallel simulation was performed using the finite difference time domain (FDTD) method. The same air-loaded design was modelled as in the previous section, except that the geometry was built using metallic cells for the metallisation instead of cylindrical wires as in the MoM package.

The FDTD software used was an in-house purpose-built FDTD program with Berengers Perfectly Matched Layer (PMLs) [11] as absorbing boundary conditions defining the computational domain. As the theory for the FDTD method is well-documented [11, 12] a revision of the theory is given in Appendix A. The geometry in the FDTD method is built up using Yee cells [13], which for this initial geometry were $1 \text{mm}^3$ in size. The geometry was built up using a mathematical defined structure, as no set primitives were available. The mathematics firstly defined the balun base as a circular metallic disk, a cylindrical metallic sleeve on top of the disk was utilised to complete the balun structure.

The two helical wires were then defined using the radius, height and pitch angle originating at the balun rim and terminating at the requisite height. The two termination points were then used for the start and finish points of the top track. One of the disadvantages of the in-house FDTD was no visualisation of the geometry being simulated was available, nevertheless the artwork was post-processed for mathematical inaccuracies and the geometry modified so no errors existed in the structure. Once the geometry was complete the structure was sourced using a voltage gap on the top tracks. The course resolution meant that a relatively fast simulation time was obtained. Key parameters used in the FDTD simulation are displayed in Table 2.2.
Although 10,000 time steps were run, convergence in the results had been obtained prior to the completion of the time steps. The extra time steps were performed to be certain that stability had been achieved and that the convergence of the results was not just temporary. The 15 Yee cells thick absorbing boundaries surrounding the antenna structure proved satisfactory for the simulation. Figure 2.6 shows the resistance and reactance curves obtained from the FDTD and the MoM simulations for the air-loaded bifilar without the cable. These impedance values were taken at the top of the antenna geometries where the voltage gap is located, which will be identified in this thesis as $Z_{ant}$. It can be seen in Figure 2.6 that a very good agreement existed between the two air-loaded models in the frequency range 800MHz to 3500MHz. The return loss curve for the MoM simulation can be seen in comparison with other simulation runs later in this chapter in Figure 2.8 and 2.9.
Having obtained good correlation between two different simulation methods for the air-loaded bifilar, the validity of the simulation needed to be proven with an air-loaded measurement. For comparison with an air-loaded measurement the simulated antenna would require a coaxial cable to source the antenna, therefore a cable was introduced into the MoM simulation geometry.

2.4 Cable Study and Insertion into MoM GSM Design

The coaxial cable was not included previously to keep the structure as simple as possible and to keep a fast simulation time. Due to the concentricity of the coaxial cable structure it was to be expected that it would require high resolution modelling for accurate representation in either the FDTD or MoM software. High resolution is required due to the small distance between the inner and outer conductors, and to the length of the cable in comparison to the distance between the conductors. In addition it is necessary to be mindful of the thickness of the conductors that are used in the modelling structure.

When modelling the coaxial cable in MiniNEC Professional for Windows the outer conductor part of the structure was handled similarly to the balun sleeve. Circular wire structures were connected together using straight wires at every 60° to form a hexagonal meshed cylindrical structure representing the outer conductor. A hexagonal structure was used due to the limitations of the number of wires forming the wire circles defining the outer conductor, this is shown in Figure 2.7. The wire circles had an outer diameter of 2.2mm that is comparable to the real dimensions of the coaxial cable. Each wire in the outer conductor had a radius of 0.13mm calculated from the thin film to cylindrical wire approximation given by \( a_c = 0.25a \) (where \( a_c \) is the radius of the cylindrical wire and \( a \) is the width of the thin film).

![Figure 2.7 Coaxial Cable Structure](image-url)
The inner conductor had a radius of 0.26mm and ran axially through the centre of the hexagonal outer conductor to complete the coaxial cable. The inner conductor was connected to the outer hexagonal conductor at the bottom by a single wire centrally sourced with a voltage gap excitation.

The desired outcome for the coaxial cable of the GSM bifilar was to obtain an impedance curve with very little reactance and resistance of 50Ω in the frequency range 800MHz to 3500MHz. The characteristic impedance of the simulated coaxial cable was calculated using two different lengths of simulated coaxial cable by using the general transmission line equation and by solving for the two lengths simultaneously.

The final simulated coaxial cable structure consisted of the inner and outer conductor geometries being sectioned. The sections through the cable took the nodal formation of a diamond format, this is reflected in the wire lengths used in Figure 2.7. This means that the sections at the top and bottom of the cable had few nodes compared to those sections in the middle. By partitioning the cable in this way, the amount of nodes required to make the cable were reduced dramatically, as well as giving a response close to that of a coaxial cable. The reactance was minimal and the resistance was within 3Ω of the desired 50Ω throughout the given frequency range as shown in Figure 2.8.

![Modelled Coaxial Cable Impedance Plot](image)

Figure 2.8 Impedance of Simulated Feed Cable
Different cable structures had been tested in finding this near ideal response for a simulated cable. Some of the structures tested included outer conductors formed using wires placed at $45^\circ$ intervals instead of $60^\circ$, which resulted in a slower run time due to the greater number of wires utilised. Other structures had regular sections to form the cable, i.e. sections of length of 2cm or 1cm, but this gave an inferior response to that of the diamond partitioned cable. The structure of the final coaxial cable design can be seen later in this chapter.

The cable was then attached to the air-loaded bifilar structure using a fishhook feed system as used in the dielectric-loaded bifilar antenna. The inner conductor of the simulated cable was extended out the top of the cable structure and hooked over the outer conductor and connected to one of the helical tracks. The other helical track was then connected to the outer conductor. Thus in most respects the air-loaded bifilar model was a close approximation to the dielectrically loaded case. However an arbitrary 10cm length of cable was used to feed the antenna, as the required $\lambda/4$ length of cable would be too small to protrude through the antenna as required. Normally the cable used would have $10\Omega$ impedance and be $\lambda/4$ in length to function as a quarter wavelength match transformer to allow for the transition from $2\Omega$ impedance at the antenna feed point to the $50\Omega$ input impedance of the coaxial cable input.

The complete air-loaded GSM bifilar structure, as seen in Figure 2.9, consisted of 361 wires and 1273 nodes. The antenna was simulated in the frequency range of 0.5-3.5GHz in 800 steps in order that a high resolution was obtained. In comparison to this, the antenna with the cable omitted had only 259 wires and 574 nodes, thus it can be seen that the insertion of the cable is costly in terms of the simulation time.
Figure 2.9 The complete air-loaded structure (a) wire view (b) Nodal View

Although the inclusion of the cable feed is costly in simulation time, it will be seen that the cable is crucial to the design of the antenna and its modes of excitation. The feed point of the antenna without a cable was located at the top of the simulated antenna with impedance $Z_{\text{ant}}$, which is different to that of the cable fed antenna that was fed at the bottom of the cable with impedance $Z_{\text{in}}$. Therefore in order that a fair comparison is obtained the impedance data needed to be manipulated so that impedance values were read from the same point.

The impedance from the antenna with the cable omitted had its impedance values $Z_{\text{ant}}$ embedded to the position of $Z_{\text{in}}$ before its return loss was calculated. The return loss plot in Figure 2.10, shows that the inclusion of the cable into the bifilar simulation excites two additional modes in the antenna structure. As the balanced mode of the antenna would be present regardless of the presence of the cable, the first resonant mode is identified as the balanced mode, the second as the single ended mode and the third as a hybrid monopole mode.
These two additional modes, the single ended and hybrid modes, will be seen to be critical to the response of the antenna and the nature of each of the modes identified later in the chapter. Electric far field plots for the balanced mode will also be visualised.

2.5 Air-Loaded Measurements and Comparisons

To check the simulated results for accuracy of simulation, air-loaded models of the dielectric-loaded antenna were built for comparison. As far as the air-loaded simulation models, the physical models were scaled up in size by a factor of $\sqrt{\varepsilon_r}$. For example for the case of the PCN bifilar (where $\varepsilon_r=36$), this implies a six fold increase in linear dimensions.

The antenna was made using a metallisation of 10μm on a dielectric PTFE substrate $\varepsilon_r=2.2$ of 0.1mm for rigidity of the structure. For extra rigidity the base of the antenna was made from FR4 of $\varepsilon_r=4.1$. A 10cm length of 50Ω coaxial cable was then inserted axially through the centre of the antenna and the top of the cable used as a fishhook feed to excite the antenna.

![Figure 2.10 Return loss of antenna with and without the coax](image-url)
The antenna was then measured using a 8753D HP network analyser in the frequency range 0.8-3.5GHz using 801 points. The return loss of the air-loaded measurement can be seen below in Figure 2.11 compared to the air-loaded antenna with cable simulation. It can be seen that the measured result like the simulated result has three modes present and that the two plots compare closely with one another.

![Figure 2.11 Measured Vs Simulated Return Loss](image)

Past publications [1,2,3,4] can be used to identify the first two modes as the balanced and single ended modes. The third mode is believed to be a monopole mode brought about by the cable inside the dielectric core. As the first mode is present even without the presence of a cable, as seen in the simulations, it is believed that for the GSM bifilar the first mode is the balanced mode and that the second mode is the single ended mode as reviewed in Figure 2.12.

![Figure 2.12 Balanced (a) and Single-ended Mode (b)](image)
The identification of the balanced mode is supported by looking at the electric far field patterns, as it is known a virtual dipole is present towards the users head, i.e. perpendicular to the top track of the bifilar in a 0.5 helical turn bifilar. The virtual dipole of the bifilar antenna is investigated more thoroughly later in the thesis by mathematical analysis and also by the use of small antenna theory. The currents on the bifilar structure can also be used to identify the balanced mode, as the phase difference in the helical tracks $180^\circ$ half way up the radiating section, typical of a resonant loop antenna. The electric far field radiation pattern of the first mode of the simulated GSM bifilar can be seen in Figure 2.13; where the change in colour in the plots represent the change in the z-axis and not variation in field intensity. It is identical to that of the measured GSM antenna and is a characteristic pattern of the balanced mode.

![Total Electric Fields](image1.png)  ![Total Electric Fields Plan View](image2.png)

Figure 2.13 Electric Far Field Pattern of the balanced mode

The comparison of the measured with the simulated results, gave further confirmation that the air-loaded model had been modelled accurately using the Method of Moments software package. Having established a modelling technique for the accurate simulation of an air-loaded bifilar, it could now be modified and investigated further with confidence. The work performed on the air-loaded bifilar structure was published by Cai et al [14] and contains more details concerning the research that was performed. The author's contribution to this publication was the modelling of the air-
loaded bifilar antenna which showed good agreement with the simulated FDTD geometry as well as the measured results.

2.6 Bluetooth Twisted Loop Antenna

Throughout the time period in which this research was undertaken the focal point of the mobile telephony market was continually changing. Telecommunications companies were requiring multi-banding antennas for the present market, with research into GPS, Bluetooth and low SAR antenna designs for future markets. A number of twisted loop designs were therefore required and built to keep up with the ever-changing market, each design configured for a different communications band with a view to study key areas of the wireless telephony market.

In the previous sections techniques were developed for simulating an accurate air-loaded twisted loop antenna based on a GSM 1800 dielectric-loaded model. From the results it could be seen that an air-loaded geometry could be accurately modelled though the model would require re-scaling in order to obtain the requisite frequency that the model was designed for.

These techniques were applied in generating models to cover different communication bands and investigating into new research areas. Potentially, one of the biggest markets for antenna design is currently the Bluetooth market, and so the next design modelled was the air-loaded Bluetooth twisted loop. The twisted loop antenna was thought to be particularly relevant to the Bluetooth market due to some of the features it possesses compared to its rivals as listed in chapter 1. Consequently there was urgency in modelling a preliminary air-loaded Bluetooth model.

The air-loaded Bluetooth bifilar was initially developed using the dielectric-loaded design and the dimensions scaled by $\sqrt{\varepsilon_r}$ as was confirmed with the PCN design. The techniques developed earlier in the chapter were applied to the Bluetooth design to generate the geometry to be simulated. The response of the antenna would then be analysed for the resonant frequency and the geometry re-scaled accordingly to move the resonant frequency to the centre frequency of the Bluetooth band. This was to
keep all dimensions scaled relative to one another in the preliminary stages. Further modifications are made to the geometry in chapter 3.

The final dimensions to make the air-loaded Bluetooth bifilar resonant at the requisite frequency are presented in Table 2.3. The dielectric-loaded dimensions were increased by a scaling factor of 3.313 to give the air-loaded dimensions with the same resonant frequency or the balanced mode of 2448.5 MHz.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Dimension</th>
</tr>
</thead>
<tbody>
<tr>
<td>Total Height</td>
<td>32.5 mm</td>
</tr>
<tr>
<td>Radiating Section Height</td>
<td>16.1 mm</td>
</tr>
<tr>
<td>Balun Height</td>
<td>16.4 mm</td>
</tr>
<tr>
<td>Radius</td>
<td>13.3 mm</td>
</tr>
<tr>
<td>Pitch Angle</td>
<td>21.1°</td>
</tr>
</tbody>
</table>

Table 2.3 Dimensions of air-loaded Bluetooth twisted loop

The response of the air-loaded Bluetooth bifilar can be seen in chapter 3 where it is used as the standard for the parametric study. For this purpose two geometries were built, with and without the cable.

### 2.7 PCN Twisted Loop Antenna

The third communications band of interest requiring an air-loaded twisted loop geometry for future investigation was the PCN band. Transition dielectric-loaded bifilar antennas designed around the PCN communications band had been manufactured in an attempt to couple the single-ended and balanced mode to make the antenna broadband. Therefore, the antennas manufactured were not designed strictly for the PCN communications band and the resonant frequency of the modes required determining from measurements before the air-loaded geometry was built. These results can be seen below in Figure 2.14, with each of the five antennas displaying
three resonant modes. The first two modes respectively were the balanced and single-ended modes, and the third resonant frequency being a hybrid mode.

![Figure 2.14 Measured PCN Bifilars](image)

When the dimensions of the dielectric-loaded bifilar are re-scaled to give the dimensions for an air-loaded structure, only one of the resonant frequencies remains constant due to the frequency shifting that transpires when the dielectric core is removed. The first resonant frequency located at 1630MHz was the target frequency for the balanced mode of the air-loaded structure. A scaling factor was required to produce the air-loaded geometry from the dielectric loaded dimension; the parameters for the air-loaded structure are displayed in Table 2.4.

An interesting part of this structure compared with the Bluetooth and PCN bifilars is the amount of helical turn. For all former twisted loop antennas in this dissertation a 0.5 helical turn was employed though for this PCN transition design a 0.43 turn is utilised. The effects of this change are examined in a parametric study performed in chapter 3.
### Table 2.4 Transitional PCN Air-loaded Antenna

<table>
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<tr>
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<tr>
<td>Radiating Section Height</td>
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<tr>
<td>Balun Height</td>
<td>22.3 mm</td>
</tr>
<tr>
<td>Radius</td>
<td>21.5 mm</td>
</tr>
<tr>
<td>Pitch Angle</td>
<td>26.7°</td>
</tr>
<tr>
<td>Helical Turn</td>
<td>0.43</td>
</tr>
</tbody>
</table>

**2.8 Simple Coaxial Cable Structure**

In modelling the fundamental twisted loop antennas required for the research carried out in this thesis, it is clear that the construction of the geometry brought about modelling issues that needed addressing. One of the issues addressed prior in the thesis was the modelling of the fish-hook feed and coaxial cable.

The cable structure that was defined for the preliminary models consisted of a meshed wire outer conductor with a sectioned wire inner conductor. The inner conductor was connected to the outer conductor at the bottom of the cable structure by a small single wire used to excite the structure using a voltage gap. The guidelines for MiniNEC states that each wire making up a junction should not be dissimilar in their segment lengths and that for best results the ratio of the segment length on one wire compared to the segment length on an adjoining wire should be less than 2.

Apart from the cable being a complex structure, the small distance between the inner and outer conductor means the wire used for sourcing the structure dictates the computational efficiency of the simulation. For numerical packages with dielectric included, or that use cells to build up its geometry the modelling of the cable becomes more intricate and very high resolution is required. An example of this is seen in the following section dealing with modelling issues. The aim of this section is a proposed
simple model for the coaxial cable, by taking into account the current flow of the different resonant modes.

In the balanced mode the cable plays no part other than to feed the antenna, however in the single-ended mode the current flow down the helices, down the balun, across the balun rim and then back up the outer conductor. This implies that to simplify the model, the antenna could be excited at the top of the artwork on the top tracks, and that the coaxial cable could be replaced with just its outer conductor thereby still allowing the currents to propagate in the single-ended mode.

This method was tested on the PCN air-loaded geometry and the return loss and impedance curves remained consistent with those of the full coaxial cable model artwork. Although the inner conductor consists of only 56 nodes the simulation time is reduced by an hour. This technique could prove more profitable in other numerical software where modelling of the coaxial cable in the twisted loop topology proves computationally inefficient with regards to time and memory required. As the smallest cell often dictates the time taken for the simulation to run. The modelling of the cable in other numerical software packages and other modelling considerations when simulating the twisted loop antenna are summarised in the following section.

2.9 **Modelling packages and considerations**

In recent years a number of simulation packages have become available for antenna modelling. Some of the packages utilised for modelling the twisted loop structure, whether air or dielectrically loaded, can be seen in Table 2.5 categorised by their electromagnetic solvers. Some of the packages below have been analysed in comparative studies in the past [15,16] though not in conjunction with the modelling of a specific artwork.

A number of solvers were included in the study with methodologies which included Transmission Line Modelling (TLM), Method of Moments (MoM) or the Finite
Difference Time Domain (FDTD). The tabulated numerical packages will be analysed on its ability in simulating the twisted loop; and modelling issues that were encountered will be identified. The overall performance of the simulator is also considered.

<table>
<thead>
<tr>
<th>TLM</th>
<th>MoM</th>
<th>FDTD</th>
<th>FI</th>
</tr>
</thead>
<tbody>
<tr>
<td>MicroStripes</td>
<td>Expert MiniNEC</td>
<td>In-house developed FDTD</td>
<td>CST Microwave</td>
</tr>
<tr>
<td></td>
<td>Professional for Windows</td>
<td>method</td>
<td>Studio</td>
</tr>
<tr>
<td>FEKO</td>
<td>LC</td>
<td>XFDTD</td>
<td></td>
</tr>
</tbody>
</table>

Table 2.5 Simulation Packages Evaluated

As previously highlighted, many issues arise when the high dielectric core is introduced into the simulation and therefore an air-loaded structure is used to circumvent modelling difficulties in some of the packages used and in packages with no dielectric capabilities. Further details of the research carried out in this section can be obtained from Jayawardene et al [17].

On the evaluation of the software packages different factors were used to appraise the applications, these factors included:

- Run time
- Resolution issues
- Meshing accuracy
- Macro function availability for fast input of parametric changes to the geometry
- Desired output

The dielectric loaded twisted loop model utilised in the evaluation has been previously documented. It is manufactured using a printed thin metallisation onto a
high dielectric constant ceramic puck with relative permittivity of 36, other dimensions are shown in Table 2.6. It is expressed electrically in terms of its modes earlier in the chapter.

<table>
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<td>36</td>
</tr>
<tr>
<td>Track Width</td>
<td>0.75 mm</td>
</tr>
<tr>
<td>Balun height</td>
<td>5.60 mm</td>
</tr>
<tr>
<td>Helix angle</td>
<td>22.48°</td>
</tr>
<tr>
<td>Metallisation thickness</td>
<td>60 μm</td>
</tr>
</tbody>
</table>

Table 2.6 Dimensions of manufactured DLBHA

When modelling the twisted loop geometry, numerical packages require a recommended number of cells or nodes per wavelength. When a high dielectric media is used a higher cell resolution is needed. This can cause excessive computational time and memory requirements. This is one of the advantages of using air-loaded geometries as it is computationally less demanding, gives a quicker run time and has a increased likelihood of accurate modelling.

The balun and helical tracks are comprised of printed curved thin metallisation that is particularly susceptible to staircasing effects. This means that often a compromise is necessary between computational time and geometry accuracy in order to obtain the required resolution. The compromise in the geometry accuracy takes the form of increased metallisation and cell size. For faster run times the cable was removed from the simulation models and a 1 Volt voltage-gap was utilised sited at the top of the antenna where the two helical arms meet. Although removing the coaxial cable was an inaccurate method to excite the antenna it did provide an efficient alternative to compare the applications.
2.9.0 Transmission line modelling application

MicroStripes. The DLBHA and similar geometries have been modelled extensively using the commercially available TLM application MicroStripes. Detailed description of this work can be found in the literature [18,19,20]. Modelling of the structure took approximately 30 hours using a Sun Sparc20 (200 MHz) using 365,000 cells and about 130 Mbytes of RAM.

MicroStripes resources can be used efficiently in grading the workspace mesh size. Therefore high cell resolution is applied around the model to account for the thin metallisation and a cruder resolution is sufficient further away. This high resolution modelling is also a requirement in the cable structure, if sufficiently high resolution is not applied then short-circuiting may occur between the inner and outer conductors. This is illustrated in Figure 2.15 where the distance between the conductors is less than a cell dimension and highlights the requirement for high resolution. This application could benefit from the simple cable structure previously developed, as short-circuiting would no longer be an issue due to no inner conductor being present.

![Figure 2.15 TLM cable meshing](image)

2.9.1 Method of moments applications

Expert MiniNEC Professional for Windows. The modelling issues for this MoM numerical software has been discussed earlier in detail, and so will be summarised for this comparative study. Other than fast simulation times, MiniNEC has the advantage
of having set primitives for geometry building and provides useful outputs including current distributions, field patterns and impedance plots.

As no dielectric-loading capabilities are available in this software only air-loaded structures were simulated. The thin metallisation was modelled using a thin wire approximation that represents metallic strips as wires, and the cable and balun were represented by convergence tested wire meshed structures. Designs with and without the cable were modelled and when the cable was excluded the run time is seen to reduce from 24 hours to 7 hours for 801 frequencies using a Pentium II 350MHz PC. The importance of modelling the cable and inclusion into the geometry was demonstrated in the duration of this analysis.

**FEKO.** This is not a purely MoM electromagnetic solver package, as it uses a combination of MoM, Unified Theory of Diffraction (UTD) and Physical Optics (PO). It is one of the few MoM based packages in today’s market that has dielectric capabilities. It has other advantages including parallel processing capabilities, conformal meshing, a range of available outputs and a CAD based geometry developer.

In contrast to MiniNEC which uses wires to build up its geometry, FEKO utilises metallic triangles and dielectric cuboids with their size calculated with respect to the nominal wavelength. For the air-loaded design with the cable excluded the metallic triangles could represent the twisted loop structure with a high resolution and gave fast run times. The results compared well with the air-loaded results of MiniNEC as seen in Figure 2.16.
Air-Loaded Bifilar Antenna

When the dielectric capabilities were introduced into the geometry the resolution of the structure was constrained by the amount of RAM required to simulate the model. As the platform used had 330Mbytes of RAM the geometry was restricted to 1812 metallic triangles and 656 dielectric cuboids for the dielectric core and took 24 hours to run on a Pentium II 350MHz PC.

Due to the memory limitations the dielectric core had a coarse resolution that gave rise to staircasing as well as dielectric cuboids protruding through the metallisation. This was problematic as air pockets were sandwiched between the helices and the dielectric core as seen in Figure 2.17. This unfortunately led to incomprehensible results and attention was focussed on a less complicated air-loaded structure which was successfully simulated.
As a result of this work carried out with FEKO, modifications are being made to the software that should improve the attachment of metallisation to dielectric bodies. These changes are necessary in order to enable the modelling of the dielectric-loaded twisted loop topology. Until this modification is made it is clear that both MiniNEC and FEKO are best suited to modelling the simple air-loaded bifilar helix antenna.

2.9.2 Finite difference time domain applications

Solving for Maxwell's equations, the Finite Difference Time Domain (FDTD) method is popular because it suits the requirement of a wide range of electromagnetic models. In conjunction with the in-house FDTD mentioned earlier in the comparison with MiniNEC, three commercially available time domain solvers were investigated to model the antenna geometry. The software models used in the antenna simulations were;

- LC (2.10) which was originally developed by CRAY computers, and now is developed by a number of collaborators
- XFDTD (demo version 5.1) developed by Remcom Incorporated
- Microwave Studio (2.0 RC1) developed by Computer Simulation Technology (CST)

In-house FDTD Code. The in-house developed FDTD code with Berengers split field PMLs and a trigonometric/geometric input file was used to model both the air and dielectric-loaded geometries. The air-loaded model has been introduced earlier and published [14] in conjunction with the MiniNEC modelling and so shall not be discussed further in this section.

The DLBHA model was attempted without a cable with 400,000 mesh cells of size 500 x 500 x 200 microns. The geometry required 2 days run time for 60,000 time steps on a 350MHz Pentium II PC. Unfortunately convergence of the impedance curves was not achieved, even when the time steps were increased to 100,000 time steps to allow for better stability. It was concluded that this was due to the crude resolution that gave rise to staircasing errors.
The advantages of using LC as a simulator is that it has set primitives for modelling, and offers an interface between LC and circuit simulator SPICE. Linux versions are available with parallel processing capabilities and displays a variety of outputs including far field patterns, current distribution and impedance plots.

Unfortunately a helix geometry was not provided amongst the set primitives for modelling. Therefore a MicroStripes input file was modified as the input to LC, with 12 cell thick second order Berengers PMLs being utilised as the absorbing boundary conditions. A resolution of 8,000,000 100 cubic micron cells were used and run on a Cray J90 single processor. The run time for 150,000 time steps was 59 hours and required 953.3Mbyte of RAM. The simulation still had not converged to a plausible result after this time.

**XFDTD Version 5.1.** Similar to other packages, this application has the advantages of set primitives for fast model generation, parallel processing and a good variety of output displays. But in contrast to previously reviewed packages it has SAR computation and a human body biological mesh is available.

A 15 cell thick second order Liao absorbing boundary conditions surrounded the modelled DLBHA. 8,000,000 cells with a size resolution of 100 cubic microns defined the geometry and took 14 days to run 90,000 time steps on a Pentium II 350MHz PC using 239.4Mbytes of RAM. Once again the impedance curves had not converged to a result.

**Microwave Studio (MWS).** The last of the packages to be used in modelling the DLBHA was based on a Finite Integral (FI) solver that is similar to the FDTD therefore categorised here under the FDTD solver applications. It uses a Perfect Boundary Approximation (PBA™) when modelling curved surfaces for increased accuracy. This accuracy of curved surface modelling is also enhanced by adaptive meshing capabilities. Other advantages that add to the versatility of this package include set primitives for modelling, single or dual processor capabilities, a variety of output formats, importation of standardised CAD file for easy model construction and a human mesh model for SAR simulations.
The helices of the DLBHA were precisely modelled by the rotation of a 100-micron thick metallic strip around the dielectric core. The arrangements for signal sourcing of the structure were unlike other packages as it utilised a waveguide port method as its excitation. A 4 cell thick Berengers Perfectly Matched Layer (PML) wall was used for the Absorbing Boundary Conditions (ABC), and 290280 mesh nodes in total were used for the modelling. Total run time was 15 hours on a Pentium III 850MHz PC to obtain the accuracy that is available from a model of this resolution.

As shown in Figure 2.18 there was reasonable agreement between the simulated and measured S11 plot for the DLBHA. As the geometry could be easily manipulated for parametric changes, this gave a good basis for further development of the dielectric-loaded twisted loop, though this is not the main subject of this thesis.

![Figure 2.18 MWS compared with measured Return Loss](image)

**Figure 2.18 MWS compared with measured Return Loss**

### 2.9.3 Modelling applications summary

TLM has been useful in the past in proving the principles regarding the twisted loop topology and has been well documented [18, 19, 20]. It can be a difficult tool for
accurately predicting the antenna response. Careful modelling and compromises are required for the topology as demonstrated by the cable example.

Both of the MoM packages were easy to use and showed good agreement in modelling the air-loaded geometry and with the measured results. However it was found that the cable was critical to the antenna response and required accurate modelling. Although FEKO did exhibit dielectric-loading capabilities it experienced and highlighted the obstacles common to both the TLM and FDTD packages.

The in-house built FDTD code gave good agreement between the air-loaded MoM model and measured results and the results documented. When the DLBHA was attempted the in-house FDTD, LC and XFDTD all failed to converge to a plausible result. Staircasing errors due to curvature of the geometry were one of the principal issues for the failure to converge. Another outstanding issue for the dielectric loaded geometry was interfacing of dielectric, metal and air in a single cell that would occur due to the thin metallisation. Some improvements were observed in these inaccuracies with MWS’s PBA™ (Perfect Boundary Approximation) method based on the Finite Integral solver.

The small antenna that is realised by the dielectric-loading requires that a correspondingly small cell size is essential in the DLBHA, this has a knock-on effect on the memory consumption and the run time. This cell size is dictated by the requirement of 2 cells minimum required for the helical track width and for the dielectric in the coaxial cable. If the metallisation thickness is modelled accurately then very high cell resolution is obligatory, though could be made more memory efficient if the software has the inclusion of adaptive meshing or graded meshing techniques.

2.10 Conclusions

It has been demonstrated that air-loaded geometries can be successfully simulated using industrial specifications for dielectric-loaded twisted loop antennas. Preliminary
models for GSM, Bluetooth and PCN were developed using a MoM software package that utilises straight wires sectioned by nodes to develop the required artwork.

It was observed in the GSM model that using a $\sqrt{\varepsilon_r}$ scaling factor when progressing from the DLBHA to the air-loaded design was not applicable and that the primary modes of resonance were located at higher frequencies. This inaccuracy was only rectified for the Bluetooth and PCN models by re-scaling the air-loaded geometries as the GSM model function was as a development tool for the principal twisted loop structure.

The balun of the antenna was modelled using a meshed structure that had been convergence tested which kept the structure simple for fast simulation time but detailed enough for accurate simulation. The convergence testing comprised of high resolution modelling to define each part of the antenna structure, and then reducing the resolution until the accuracy of the model is compromised. The testing also increased the number of wires used in defining the balun and helical wires, but it was found that increasing the mesh definition of the structure beyond a point made the antenna more difficult for the numerical solver to calculate and inaccuracies were observed. Ultimately a simple mesh structure for the balun and helical wires were utilised which had a short simulation time but good numerical accuracy.

The convergence testing demonstrated the compromise between run time and resolution that is common with many simulation packages. It was also shown that the radiating section, which is critical to the functionality of the antenna, did not need a high-resolution meshed model for the helical arms and that single wires employing the thin wire approximation technique could be utilised with little or no change in the antenna response.

The MiniNEC MoM twisted loop model with the cable excluded was compared with simulated results obtained from an in-house built FDTD method software and then later with a commercially available MoM software package FEKO. All packages showed good agreement for the basic model, even though coarse resolution was utilised with the FDTD method.
When a proposed simulated coaxial cable was introduced as the feed system additional resonant modes were viewed, substantiating that the cable is critical to the antenna’s response and in particular for the single-ended mode. With the cable included into the design good correlation were seen with measured results. Due to the complexity of modelling a coaxial cable in many numerical packages a further feed system was proposed that eliminated the inner conductor and would allow for slightly coarser resolution. This feed system allowed for the propagation of the single-ended modes and hybrid modes and compared well with previous results.

Finally a comparative study was performed on a number of commercially available numerical packages in conjunction with the twisted loop antenna. It was seen that due to the thin metallisation and curvature of the DLBHA stair casing in the geometry and air pockets between the metallisation and dielectric core became an issue for some of the packages utilised. These issues can be reduced by the use of higher resolution modelling by use of smaller cell size. The smaller cell size is also required for a high dielectric as an increased number of cells per wavelength for the geometry is needed. But the consequence of a smaller cell size is increased run time and memory requirements and satisfactory modelling may still not be achieved.

High resolution modelling is a requirement when the coaxial cable is introduced into the geometry. Stair casing issues are seen and compromise between modelling accuracy and memory requirements becomes a necessity. The problems associated with modelling the thin metallisations in the dielectric-loaded twisted loop topology have been somewhat resolved by adaptive and graded meshing techniques, though much care and time is essential in the generation of the model.

The air-loaded design provides a simpler and more robust structure to model that is faster to generate and more computational efficient in terms of run time and memory requirements due to absence of the dielectric core. Parametric changes are easier to implement and fewer issues need consideration, as no metallisation attachment to dielectric is present. For these reasons the analysis utilises air-loaded topologies.
2.11 References


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CHAPTER 3.0

INVESTIGATION INTO BIFILAR ANTENNA THEORY

3.0 Introduction

Simulated air-loaded bifilar helix antenna designs had been developed in chapter 2 and these showed good agreement with measurements and when compared with simulated results from other packages for accuracy. In this chapter investigations are performed focusing more in-depth at the antennas properties. This attempts to show how the air-loaded twisted loop design relates to the published theory by Leisten et al [1-6] of the more complex dielectric-loaded bifilar antenna. It also attempts to advance the existing theory relating to the twisted loop antenna.

A parametric study is performed in section 3.1 on the twisted loop antenna to identify the sensitivity of the primary modes due to subtle changes in the antennas topology. A study of this nature ultimately aims to help identify the changes in antenna response caused by the inaccuracies of the manufacturing process. For example the dielectric core of the dielectric-loaded bifilar helix has a dielectric constant of 36 ± 1, though the shift in resonance caused by the inaccuracy of the dielectric is unknown. A parametric study could identify how much change could be expected from inaccuracies in the dielectric constant and suggest ways in which the artwork could be modified to bring resonance back to the required communications band as modifications to the dielectric core is problematical.

More importantly, and the objective of this investigation, was whether or not the bifilar could be made broadband by the coupling of its primary resonant modes. Amongst the parametric changes observed in the investigation are variance in the balun height, diameter of the antenna and changes in the helical turn. From this study conclusions can also be drawn about whether broad banding of the antenna is possible by the coupling of its primary modes.
The helical turn investigation in the parametric study is monitored for far field pattern deformation in the balanced mode. This is due to structure of the radiating section being critical to the characteristic electromagnetic dipole pattern, which ultimately contributes towards the low Specific Absorption Rate (SAR) properties of the dielectric-loaded bifilar antenna. Therefore any changes that are made in the radiating section are carefully monitored by analysis of the return loss and the far field patterns.

The parametric study will also investigate the primary modes of interest for the twisted loop in the context that they are more complex than present theory indicates. Section 3.2 attempts to further present theoretical understanding by briefly examining the near fields of the air-loaded bifilar antenna. This will help identify the current flow associated with the primary antenna modes. As the balanced mode has been well documented [1,2,3,4] the key interest in this section will be the single-ended mode.

After examining the near fields in more detail, the properties of the balanced mode in the far field are examined. Only the balanced mode is investigated in the far field, as it is the main mode of operation for the bifilar, and displays a characteristic pattern due to a horizontally disposed virtual electromagnetic dipole which is of physical use for low SAR applications. In Section 3.3 small dipole antenna theory is used to try and identify the composition of small electric and magnetic dipoles that could be utilised to form the characteristic pattern of the balanced mode. Theory states an electromagnetic dipole exists due to a magnetic reversal in the radiating section [1,2,3], so modelling of electric and magnetic dipoles are used to prove the existence of the electromagnetic dipole. This would also give an understanding of the electric far field for the antenna in terms of the dominance of either the electric or magnetic dipole, which would in turn be dependent on the topology of the antenna.

Techniques to broadband the air-loaded bifilar antenna are developed and analysed in Section 3.4. The first technique attempted was the flaring of the helical arms in the radiating section to form a Bowtie twisted loop structure. A bowtie formation was considered due to its broadbanding nature, whilst making the metallisation in the radiating section more complementary; i.e. the metallisation and non-metallic parts of
the radiating section would have the same configuration, which is a recognised procedure for broadbanding.

The other technique utilised for broadbanding was introducing parasitic helical arms to create more resonant modes that could be coupled with the primary modes. A study on the effects of the additional arms on the antennas response was performed including changing the length of the parasitics and their positioning around the balun. It will be seen that as the parasitics are positioned closer to the existing helical arms coupling between the arms occurs. This coupling could have adverse effects on the primary modes and therefore the electric far field patterns of the balanced mode were again monitored for variation.

3.1 Parametric Study

During this research, the bandwidth of the dielectric-loaded Bluetooth antenna design needed broadening, as the existing design did not cover the Bluetooth frequency band. The air-loaded design was chosen to aid the advance of the dielectric-loaded model in attempting the coupling of the single-ended and balanced modes. A parametric study would not only resolve if the two primary modes could be coupled but would also demonstrate the changes in antenna response that could expected when different parameters of the antenna are compromised; for example if the dielectric constant of the ceramic core was not precisely 36. It could also reveal what corrections could be made to the antenna if the resonant frequency was found to be too high or too low.

The standard for the studies was assembled by use of an industrial specification for a dielectric-loaded model and modified for the air-loaded geometry. The GSM design earlier in the thesis showed the scaling factor from the dielectric to air-loaded geometry was not a straightforward $\sqrt{e_r}$ factor. Consequently the geometry was modified accordingly to obtain the centre frequency of the Bluetooth band (2448.5 MHz).
To couple the two modes a parametric approach was taken, that allowed an analysis of the modes that are characteristic of different geometric options. This gave insight into how much influence the changes in the geometry had over the antenna response. For efficient design it is advantageous to know what alterations to the antenna’s geometry needed to be made to couple the modes. This could prove to be time efficient in designing future artworks by the modification of past specifications through enhanced understanding of the bifilar theory.

Although the investigation was done on an air-loaded model the results were to be correlated in a future study with the dielectric-loaded model so that a direct correlation between air-loaded simulations and real life ceramic bifilar antennas existed. The proof of correlation was attempted using a metallisation printed on a thin dielectric, which was subsequently wrapped around the ceramic core of dielectric constant 36. The results of the ‘wrap around’ dielectrically loaded antennas did not exactly replicate the manufactured DLBHA, as might be expected due to the complexity of the attachment of the metallisation to the high dielectric.

The air-loaded to dielectric-loaded model correlation was also attempted using a metallisation printed on a thin high dielectric substrate that could be wrapped around to form the cylindrical antenna. The main issue with this technique is the higher the dielectric the more brittle is the substrate. Therefore the correlation could not be attempted with existing technology, although some companies are developing high dielectric flexible substrates that may ultimately be ideal for this type of investigation.

Two designs were utilised for the parametric study, with and without a cable present, so the modes of excitation could be thoroughly examined. The Bluetooth model with the cable excluded was to focus more on the effects of the geometric changes on the balanced mode, as no single-ended mode would be present. The geometry with the cable enabled the examination of both the single-ended and the balanced modes in coalition with each other. Together the studies could thereby give more information about how the antennas perform under contrasting circumstances. Through consideration of the antenna responses in this way it is possible to interpret whether or
not the two independent modes could be coupled in such a way to give rise to broad banding.

The frequency of interest for the balanced mode was the centre frequency of the Bluetooth band 2448.5 MHz. The dimensions of the Bluetooth geometry for the standard can be seen in Table 3.1, a scaling factor of 3.3 was used on the dielectric-loaded model to obtain the air-loaded model.

![Original Dielectric-Loaded Bifilar Antenna used for the Air-Loaded Models](image)

<table>
<thead>
<tr>
<th>ANTENNA PARAMETER</th>
<th>VALUE</th>
</tr>
</thead>
<tbody>
<tr>
<td>Total Height</td>
<td>32.5 mm</td>
</tr>
<tr>
<td>Radius</td>
<td>13.3 mm</td>
</tr>
<tr>
<td>Radiating Section Length</td>
<td>16.1 mm</td>
</tr>
<tr>
<td>Balun Height</td>
<td>16.4 mm</td>
</tr>
<tr>
<td>Helical Turn</td>
<td>0.5</td>
</tr>
</tbody>
</table>

Table 3.1 Parameter of Air-loaded Twisted Loop

The parametric study of the bifilar comprised of four independent geometric changes; each designed to analyse the different effects of the changes in the antenna response. The first two investigations concentrated on the effects due to the change in the puck diameter, this study was relevant due to the tolerances of the ceramic pucks in manufacturing. In the extreme case ceramic pucks with a specified 8.0mm diameter had been measured at 7.8mm. Therefore by varying the diameter of the air-loaded model, the effects on the antennas response could be observed.

The third investigation looked at variance in the antenna height by changing the balun height and maintaining the radiating section. Obviously this study was intended to show the effects that could be expected if the ceramic core had a greater height. To
complete the parametric study, the helical turn in the radiating section of the antenna was varied whilst maintaining the height of the antenna. This exploration was particularly important as the radiating section topology is the key to the antenna’s behaviour, and so both the response of the antenna and the total electric far field patterns were analysed for any changes.

3.1.0 Puck Diameter Variation

The first study involved the Bluetooth geometry without the cable to see what happened to the balanced mode when the puck diameter was varied with respect to the standard. The diameter of the Bluetooth model generated in chapter was increased by 2.0mm in 0.5mm steps. It is reasonable to hypothesise that the effect of increasing the diameter would cause the electrical length of the balanced mode to increase and the resonant frequency to decrease. The opposite could be expected to occur when decreasing the diameter of the air-loaded puck; as was investigated when the diameter was decreased by 2.0mm in 0.5mm steps. These hypotheses were verified by the results of this simulation investigation shown in the return loss plot below in Figure 3.2.

![Figure 3.2 Diameter Variation of Air-Loaded Bluetooth Model with No Cable](image-url)
When the cable was added to the geometry the single-ended mode became enabled in the antennas response. It is to be expected that the changes in the diameter would not affect the single-ended mode of excitation in the same manner as it had the balanced mode. As the modes of excitation have different routes of current flow as shown in Figure 3.3. Both modes have in common that the currents use the top track and the helical arms, but in the single-ended mode the currents flow down the balun, across the base of the balun and the back up the outer conductor of the coaxial cable. This is in contrast to the balanced mode where the currents flow down the helices, around the balun rim, then back up the geometrically opposite helix, along the top track to the cable to from the path loop.

![Figure 3.3 (a) Balanced and (b) Single-Ended Modes of Excitation](image)

The main variance between the electrical lengths of the modes is caused by the difference of the electrical length round the balun rim imposed by the change in puck diameter. The difference in these two lengths would represent any extra shift in resonance seen by one of the modes compared to the other. Again the procedure of the first study is repeated with the diameter of the original twisted loop structure being altered by 2mm in 0.5mm steps. The return loss results of the investigation can be seen in Figure 3.4.
Due to the greater distance of the currents flowing around the balun rim in the balanced mode compared to single-ended mode where the currents pass across the balun base, the balanced mode shifts more in frequency when the diameter of the antenna is altered. For a 0.5mm change in the diameter, a frequency shift of 45MHz was seen in the balanced mode, the same as seen in the first study, compared to only a 15MHz shift in the single-ended mode.

From known current flow theory of the bifilar the extra frequency shift in the balanced mode was expected, although the amount of shift that would occur was unknown. As a guide, the frequency shift could be approximated for the two modes using two geometries from Figure 3.4 whose diameter differed by 0.5mm. MiniNEC specifies the length of each wire used in the simulations, so by adding the lengths of the wires used in each of the paths taken by the currents in the simulations the frequency shifts could be calculated with respect to differential electrical path length. It was found for a diameter reduction of 0.5mm the balanced mode would shift by 36MHz and the single-ended mode would shift by 24MHz.

Although the calculated values using the path length of the simulated models do not accurately predict the simulated results seen in Figure 3.4 they do show that the
balanced mode shifts more than the single-ended mode when the diameter of the antenna is adjusted. This independent movement of the modes with respect to one another could be useful for broadbanding an antenna, as to couple the two primary modes one of the resonant modes would need to shift by more than the other mode.

3.1.1 Balun Height Variance

For the third study the balun height was the variable whilst maintaining the height of the radiating section. This meant that the total height of the antenna would change in accordance with the height modification of the balun. From the theory of the bifilar, if the modes are operating optimally at their frequencies of resonance then the balanced mode should not be affected by this change. Due to the currents flowing around the balun rim changes in the balun height should not affect the balanced mode. The single-ended mode would be affected as the current flows down the balun and across the balun base.

From this hypothesis it was perceived that frequency shifting would occur in the single-ended mode and minimal change would occur in the balanced mode. The shift in the single-ended mode would only be small compared to when the diameter was changed as the diameter of the antenna has a geometrically greater effect on the electrical path length. For example, changing the diameter of the puck changes the lengths of the helices, the top track, balun rim and base.

In performing the investigation, the balun was increased and decreased by 2mm in 1mm steps. The total number of simulations run was reduced, as it was believed that trends would be visible with half the amount of simulations of the two previous studies. The return losses of the investigation can be seen for comparison in Figure 3.5.
It is evident from the plots that both modes are affected by the change in the balun height. For a 1mm change in balun height the balanced mode experienced a frequency shift of 6.5MHz this shift implies that the balun is not working optimally and that some current flows down the balun sides in the case of the balanced mode. The single-ended mode shifted by 15MHz for a 1mm alteration in the balun, unfortunately for an increase of 1mm an increase of 15MHz transpires which does not confirm the bifilar theory.

The balanced mode has been well-reported [4,5] and behaves in accordance with theory. The single-ended mode is more complex in its understanding due to the current flowing on the balun base and on the coaxial cable as seen in the near field plots later in this chapter. This means that the balanced mode will follow present bifilar theory adequately, but for complete understanding the single-ended mode requires a more perceptive analysis; see section 3.2.

3.1.2 Helical Turn Investigation

In the final investigation of this parametric study, the amount of helical turn was analysed. Up to this point the bifilar only had a 0.5 helical turn in the geometry. By performing this geometric change it was believed that the response of the antenna...
would vary greatly as the radiating section is a key aspect of the antennas intricate design.

The total height of the antenna and the pitch angle of the helices were kept constant by adjusting the balun height to meet the helical arms. The amount of turn was reduced to 0.4 and then increased to 0.6 in 0.05 steps. An extra 0.05 turn reduction to 0.35 was considered to help explain some interesting results that arose when the helical turn was reduced. The balanced and single-ended modes could both be expected to be affected as the electrical length of the antenna is altered quite severely for both modes. For a change in helical turn of 0.05, the electrical length of the balanced mode changes by 8.9mm and the single-ended mode by 5.7mm.

As the radiating section is of paramount importance to the total electric far field patterns in the balanced mode of the bifilar geometry, the far field patterns were monitored for deformation in the characteristic electromagnetic dipole pattern. This was of concern, as a positive side of the bifilar is its low SAR properties when operating in the balanced mode as published by Leisten et al [1, 2, 4, 5, 6] and as seen later in the chapter. So any alterations that might be made to the radiating section to improve the antenna response may have an adverse effect on the SAR of the antenna.

From the return losses of this investigation seen in Figure 3.6, it can immediately be seen that the helical turn variance has a greater effect on the antenna response than the other variables in the parametric study.
As the amount of turn was increased the resonant frequency of the balanced mode decreased rapidly due to the increased electrical path length. The electrical length of the single ended mode was compensated for by the increased balun height, so the resonant frequency decreased though not in the same capacity as the balanced mode. This determined that when the helical path length was decreased the resonant frequency of the balanced path increased faster than that of the single-ended mode and that consequently the two modes began to couple. As the balanced mode approached the single ended mode, the frequency shift changed from 180MHz to 280MHz. This implied that the balanced and the single-ended mode could be coupled together, as shown by the response of the 0.35 helical turn simulation.

When the helical turn was reduced to couple the two modes further, the shift in resonant frequency of the single ended mode changed from 60MHz to 5MHz before coupling with the balanced mode. This meant in theory the broad banding by coupling of the bifilar’s primary modes is possible, although to be completely certain of this the return loss response requires monitoring to make sure the match doesn’t deteriorate so as to make the antenna inoperable when the bifilar is changed from a 0.5 helical turn antenna.
The change to the amount of helical turn that was necessary to enable the two modes to couple can be considered to be quite severe in terms of the antenna geometry visualisation. Due to such severe deforming of the natural design of the 0.5 turn bifilar, it was prudent to examine the far field to gauge the effects. The 0.5 helical turn geometry had the expected virtual dipole pattern with the null of the pattern perpendicular to the top tracks of the artwork as seen in Figure 3.7(a). Comparing this with the 0.4 helical turn in Figure 3.7(b), it can be seen that the reduction of the helical turn has altered the natural dipole total far-field pattern characteristics of the bifilar.

![Total Electric Fields](image)

**Figure 3.7 Total Electric far fields for 0.5 helical turn (a) and 0.4 helical turn (b)**

As the amount of helical turn is reduced the pattern of the antenna changes. It can be postulated that there are two interacting reasons that cause the distortion to the pattern. As the balanced and single-ended modes couple together the patterns that are associated with their distinctive modes hybridise into a pattern that is distorted.

As the helical turn is reduced further the coupling of the primary modes increase and the currents flow throughout all metallisation of the antenna structure. For the 0.35 helical turn the two modes had coupled and the antenna was therefore broad banded by the reduction of the helical turn. It was seen that as the virtual electromagnetic dipole became less distinctive as the modes couple, the patterns got more distorted. Figure 3.8 shows the far field pattern of the 0.35 turn Bluetooth geometry, the effects caused by the two modes coupling can clearly be seen.
3.2 Primary Modes' Near Fields

One of the interesting points arising from performing the parametric study on the air-loaded twisted loop antenna was that although the balanced mode responded according to the theory, the single-ended mode was seen to deviate slightly to some of the geometric modifications. For example, an increase in the balun height the single-ended mode caused an increase in resonant frequency; whereas a decrease in resonance would be expected.

This unexpected behaviour was reasoned to be linked with the current flow in the single-ended mode, and that perhaps the path taken is more complex than presently understood. Past publications [6] imply the current flows across the top tracks, down the helical tracks and balun sleeve then across the balun base as shown in Figure 3.9.
If the currents did follow the path as indicated by Figure 3.9, then a single-ended mode could be expected to appear when the antenna has the cable excluded and is excited by a voltage gap where the top tracks meet. Though it has been seen that the single-ended mode only appears when the coaxial cables' outer conductor is present in the geometry. This implies that the outer conductor plays an essential part in the current flow path. Therefore by analysing the near fields of the bifilar antenna in both the balanced and single-ended, it was believed that the paths of the currents could be validated.

![Figure 3.10 Orientation of Bifilar for Near Fields Investigation](image)

For the investigation of the near field of the bifilar was rotated by $90^0$ with respect to its top tracks. This was so the maximum current on the helical tracks that helps induce the electromagnetic dipole could be observed in the form of near fields and not be obscured by potential fields emanating from the cable feed structure. The near fields for the primary modes are shown in Figure 3.11.

![Figure 3.11 (a) Balanced and (b) single-ended near fields](image)
In Figure 3.11(a), the currents for the balanced mode follow the theory closely, with little current flowing down the balun sleeve to the base of the balun. The current seen down the balun sleeve shows the antenna is not quite optimally designed. This slight current flow would account for small changes in the resonant frequency when the balun sleeve height is altered in the parametric study.

Figure 3.11(b) shows when the twisted loop is operating in the single-ended mode, the coaxial cable does have current flowing on the outer conductor as hypothesised. The coaxial cable forms a $\lambda/2$ resonator with the balun, as both have $\lambda/4$ electrical lengths, which when combined with the helical arms and top tracks completes the resonant path loop taken by the current to create the single ended mode. This highlights the importance of the inclusion of the cable into the geometry, and increases the comprehension of the single-ended mode.

### 3.3 Balanced Mode Electromagnetic Dipole

Having viewed the near fields of the primary mode of the air-loaded twisted loop antenna, the resulting far field pattern of the balanced mode was examined more closely. Past publications [1-6] state that the characteristic far field pattern for the twisted loop antenna is caused by an electromagnetic dipole. The publications have previously identified this electromagnetic dipole through the magnetic reversal that occurs due to the bifilar structure and not through the modelling of small electric and magnetic dipoles.

The dielectric loaded bifilar antenna can be classed as a small antenna and therefore its far field patterns should be reproducible from small antenna theory using the modelling of small electric and magnetic dipoles. It is necessary to constrain the analysis to consider small antennas as opposed to consider antennas with significant lengths or volumes to insure that mathematical solutions are obtainable. To perform this analysis the assumption is made that the length of the antenna is very small compared to the wavelength ($l<\lambda$), and also that antenna is made of wires of very small radii so no variation of current is seen across the radius of the wire ($a<\lambda$). This
implies that the current across the antenna is considered to be constant or a virtual point source. As this work is aimed to reproduce the fields produced by the electromagnetic dipole formed by the bifilar helix antenna, these mathematical assumptions are valid as the current is present over a very small distance compared to its wavelength.

This section aims to show that when operating optimally in the balanced mode the far field pattern can be represented by an electromagnetic dipole modelled using small antennas. Defining the far field patterns in terms of the dominance of the electric or magnetic dipoles that are used to represent the twisted loop will extend present understanding of the balanced mode.

If possible this work can give a very simplistic look of the bifilar antenna in terms of its virtual electric and magnetic dipoles. This could be extended to looking at how the dominance of the dipoles are effected in different media; i.e. bifilars with ferrite cores instead of dielectric or varying dielectric constants. By the use of different cores in the bifilar structure, the optimisation of maximising the efficiency of the antenna could be investigated as different field components would be affected by different media. As there is a trade off between efficiency and the specific absorption rate in antennas further analysis could also be performed in this field. Therefore performing analyses of this nature could be considered as a powerful tool for future research.

From past publications and looking at the characteristic near fields in the previous section it was seen that the current flow is well defined by the theory in the balanced mode. The current flows from the top of the antenna down the helical wires, round the balun rim then back up the geometrically opposite helical track. When operating optimally, minimal or no current flow should be seen on the balun sleeve and base.

This implies a rectangular loop antenna twisted by $180^\circ$ could represent the bifilar antenna operating in the balanced mode; this is due to the balun structure being made redundant as a result of the main current propagation being around the loop of the radiating section. When the electric far fields of the twisted rectangular loop are visualised it can be seen that the fields show good agreement with the bifilar structure.
Figure 3.12. The twisted rectangular loop represents an idealistic bifilar pattern if the currents flowed exactly according to theory, and this idealistic behaviour is what is desired when the bifilar twisted loop topology is modelled using small antenna theory.

Co-polar and cross-polar field components were examined for the comparison of the far fields of the twisted loop with the small dipole antenna, using elevation planes at 45° intervals. From looking at the field components for the twisted loop structure in Figure 3.13, it could be seen that the co-polar and cross-polar patterns at 0° were both figure of eight configurations with the other orientations being moderately omni-directional.
Figure 3.13 Theoretical Co and Cross polarisations of the Twisted Loop

When visualising far field patterns in terms of their components, dipoles are normally approached in terms of their principal planes i.e. $0^\circ$ and $90^0$ planes. From this conventional thinking the cross-polar principal planes of the twisted loop can be identified as that of an electric dipole orientated along the x-axis. Unfortunately no single electric dipole configuration could give rise to the patterns seen in the co-polar
principal planes of the twisted loop. Therefore the first simulation to be run was that of a small electric dipole whose patterns are shown in Figure 3.14

![Figure 3.14 Dipole orientated along the X-axis](image-url)
By comparing the far field patterns of the twisted loop with the electric dipole the following points were derived;

- The patterns located on the $45^0$ and $135^0$ planes for the x-axis dipole were cardioid in appearance and didn't have the omni-directional profile as demonstrated by the twisted loop topology.
- The electric dipole displayed co-polar cardioids with their maximum pointing south and cross-polar cardioids with their maximums pointing north.
- Comparison of Figure 3.13 with Figure 3.14 showed only the cross-polar principal planes of the X-axis dipole showed good agreement with the twisted loop far fields as seen

Although only two out the eight planes showed good agreement, it was decided that a pursuit of the twisted loops co-polar principal planes could nevertheless be very beneficial in finding the required small antenna configuration. As no simple electric dipole arrangement would give the required plots modelling of small loops were used to simulate the magnetic dipoles. Past publications [1-6] state the electromagnetic dipole is aligned orthogonal to the top tracks, the magnetic dipole was aligned so it lay orthogonal to the top tracks of the twisted loop along the X-axis shown in Figure 3.15.

Analysis of the co-polar and cross-polar components showed that the required patterns were seen in the co-polar principal planes and minimal field strength was seen in the cross-polar principal planes. The $45^0$ and $135^0$ cuts showed cardioid patterns though opposite in orientation to the electric dipole previously described. The far field patterns of the magnetic dipole are shown in Appendix B.

Now that simulations had been run that individually satisfy some of the principal planes of the twisted loop field patterns, the x-axis magnetic and electric dipoles were combined in a single simulation, though individually fed and excited simultaneously so the two dipoles would be excited at the same time as shown in Figure 3.15.
Figure 3.15 Simultaneously Fed Electric and Magnetic Dipole

Figure 3.16 shows the field strengths of the two dipoles were maintained comparable to each other, so when combined they gave good agreement with the twisted loop in the principal planes.

Figure 3.16 Electric and Magnetic Dipoles simultaneously fed
In addition the cardioid patterns of the electric and magnetic dipole antennas in the 45° and 135° planes, which were oppositely orientated with respect to each other, combined to give omni-directional patterns as exhibited by the twisted loop antenna.

Having found the configuration of small dipole antennas that encapsulates the far field pattern for the balanced mode of the bifilar antenna, the field intensity of the electric and magnetic dipole were altered so that conclusions could be drawn about their relative importance in forming the far field. This might be expected to be useful information when interpreting measurements as the effect of the dielectric core or amount of helical turn could affect the formation of the magnetic reversal and formation of the electromagnetic dipole.
When the magnetic dipole is increased slightly in radii size with respect to the electric dipole the patterns associated with the magnetic dipole dominated the twisted loop far field patterns. The magnetic dipole co-polar patterns become prominent and the electric cross-polar patterns become small as shown in Figure 3.17. In the $45^\circ$ and $135^\circ$ cuts the cardioid curves associated with the magnetic field begin to reappear. The opposite occurs when the electric dipole is larger in size with respect to the magnetic dipole.

The dominance of the virtual magnetic dipole can be visualised when viewing the far field measurements of the 0.43 helical turn dielectric-loaded bifilar antenna for PCN applications. This dominance, highlighted in this section could be attributed to a number of factors or a combination of factors in the PCN antenna. This includes the helical turn having completed under a half turn as investigated in the parametric study earlier in the chapter, or the presence of the dielectric core which is approached in the following section. Only the half turn twisted loop structure provides the required alignment of the electric and magnetic dipoles for the electromagnetic dipole. The measured PCN bifilar fields can be seen in Appendix B along with more comprehensive results that were obtained for this small antenna investigation into the electromagnetic dipole. Appendix B also includes far field patterns for the X-axis electric and magnetic dipole separately, and the fields seen when either of the two dipoles dominates over the other when simulated together.

### 3.4 Broadband Bifilar

The twisted loop topology has shown that it is easily scaled to cater for a variety of communication bands. Unfortunately the requisite bandwidth for these bands is often not covered by the inherent narrow bandwidth of the dielectric-loaded antenna and techniques were required to increase the working spectrum of the antenna. The air-loaded model was chosen to investigate these techniques due to its fast run time and ease of developing new geometries. Earlier in the chapter broadbanding was seen in the parametric study by the reduction of helical turn to induce coupling of the primary modes. This caused deformation in the balanced mode far field pattern and could be
considered as adverse modification of the antennas geometry in terms of the antennas SAR.

Two other techniques were developed in an attempt to increase the operational bandwidth of the antenna. These techniques involved the addition of helical wires to the original topology, thereby keeping the natural look of the twisted loop structure. The first method involved making the helical wires of the radiating section flare out as they reach the balun to form a bowtie type structure. The second method looked at implementing helical wires around the balun as extra resonant modes to couple with the primary modes. The analyses of these techniques were designed primarily as development tools for the dielectric loaded twisted loop antenna and measurements were only to be taken if the simulated results warranted further investigation.

3.4.0 Bowtie Broadbanding

When the twisted loop structure was simulated, the antenna displayed a narrow band response; therefore broadband antenna structures were examined with the concept of implementation into the bifilar antenna structure. A bowtie antenna is a well-documented antenna [7, 8, 9] geometrically approximated to the biconical structure that exhibits broadband characteristics. The bowtie antenna was chosen due to its simplistic form so it could be easily incorporated into the twisted loop structure maintaining the cylindrical shape of the overall design.

The bowtie flare in the radiating section was represented by a varying number of helical wires with varying pitch angles. This is demonstrated more clearly by the $\pm 45^0$ flare bowtie bifilar displayed in Figure 3.18. Each of the additional helices were comprised of 20 wires each and given a pitch angle that would meet the balun rim at a nodal point.
The bowtie’s flare was increased in steps of $15^\circ$ on either side of the original helices ranging the total flaring from $0^\circ$ to $90^\circ$. Convergence testing was performed on the meshing of the bowtie structure for accuracy of modelling, though additional meshing only added to the run time and showed that a simplistic design as in figure 3.18 was acceptable. To maintain simplicity the coaxial feed was excluded from the geometry and the antenna was fed with a simple voltage gap. The coaxial cable was also excluded from the design due to the increased run time that the additional helical wires brought to the structure. As the bowtie was designed to affect the bandwidth of the balanced mode, the exclusion of the coaxial cable was permissible.

The S11 responses of the Bowtie antennas were very similar to the original bifilar antenna with no flaring and so are not displayed. Increasing the flare angle of the bowtie did increase the bandwidth though only slightly. A 10% percentage bandwidth was seen with the ±45$^\circ$ flare bowtie bifilar and was the greatest bandwidth shown by the simulated antennas. This increase in percentage bandwidth from 7.5% to 10% was not sufficient when considering the loss of bandwidth that occurs when going from an air-loaded to dielectric-loaded design, and therefore other possibilities for broadbanding needed considering.

### 3.4.1 Broadbanding by Insertion of Parasitics

The parametric study earlier in the chapter showed the primary modes of the air-loaded twisted loop antenna could be coupled together by making modifications to the
helical turn. This section utilises the concept of mode coupling to achieve the required
greater bandwidth.

The Bluetooth air-loaded bifilar antenna is utilised for this analysis, as the required
bandwidth for the complete Bluetooth band (2400MHZ – 2485MHz) was not covered
by the dielectric loaded model antenna response. Mode coupling was introduced by
creating additional resonant frequencies in the structure by the insertion of parasitics.
The bifilar primary operation mode is that of the balanced mode where the current
flows in a simple loop around the radiating section as previously described. The
parasitic modes were aimed to couple with the balanced mode, so they were
introduced to the structure as additional helical arms placed around the balun rim. The
investigation comprised of two initial studies using the parasitics on the air-loaded
design to probe into the potential of the idea to see if further development would be
advantageous.

The first study concentrated on the variation in length of the parasitic helical arms
analysing for changes in resonant frequency of the parasitic and balanced modes and
for coupling that occurred between the two modes. The second study would examine
changes in the antennas response due to change in location around the balun rim. This
would monitor for direct coupling between the original helical arms and the parasitic
arms. Together the two studies would give a brief but informative insight into a
parasitics role towards broadbanding.

3.4.1.0 Variance in Parasitic Length

The main helices of the radiating section consisted of 20 wires each and gave good
visual and numerical accuracy as shown in chapter 2 and demonstrated by Best [10].
Due to this numerical accuracy, the parasitics were generated from the original
radiating structure. The helical wires of the radiating structure were copied and
rotated 90° around the balun rim and produced two parasitic helices unconnected at
the top of the antenna structure as demonstrated by Figure 3.19. The initial rotation
constant of 90° was chosen to keep the parasitics as far as possible from the main
helices to minimise helix-to-helix coupling at this stage.
The parasitics were then varied as a percentage of their original size from 100% to 0%, where 0% is the standard Bluetooth design. The parasitic elements were reduced in 5% steps as allowed by the helix resolution, and the return loss of the antennas was monitored. The reduction in parasitic height would increase the resonant frequency of the parasitic mode that would then couple with the existing balanced mode. A sample of the simulated results where the parasitics elements were reduced from 100% to 50% is displayed in Figure 3.20. Included in the curves is the plot of the original Bluetooth design with the parasitic elements excluded for the comparison.

![Figure 3.19 Bluetooth twisted loop with parasitics](image)

![Figure 3.20 Bluetooth with parasitic elements from 100% to 50%](image)
In figure 3.20 it can be seen that as the parasitic elements decrease in length their resonant frequency increases, as did the resonant frequency of their corresponding balanced mode. The increase in frequency of the parasitic helices was greater than that of the balanced mode and coupling was apparent. The match of the parasitic mode is seen to improve with the reduction in size whereas the balanced mode match deteriorates.

The greatest amount of coupling was seen when the elements were at approximately 60% of their original size where the bandwidth covered at -3dB was around 500MHz. As Bluetooth antennas are required to meet a match of 9dB return loss this bandwidth may not be useful, although it is recognised that with poor modelling accuracy the true position is not yet clear.

As the parasitic elements were reduced further from 50% to 0%, the resonant frequency of the parasites became too high to influence the antennas response significantly. When the lengths were reduced the match of the balanced mode increased and the resonant frequency of both the antennas original primary modes increased, returning to their natural positions, as displayed in Figure 3.21. This contradicts the explanation for the larger parasitics; this is due to there being uncertainty in the identification of the modes when the parasites are introduced.
When the parasitics are 100% in length three modes are seen to exist, the first is thought to be a parasitic mode, the second is the balanced mode and the third is a single ended mode. As the length of the elements is decreased, the single ended mode is lost, the parasitic mode increases in frequency as expected and the balanced mode shifts towards where single-ended mode is thought to exist.

This implies the parasitic reduced the balanced mode and single-ended modes dramatically, improving the match of the single-ended mode and deteriorating that of the balanced mode. As the parasitics elements were decreased in size the balanced mode increased in frequency and had improved match, and the single-ended mode experienced an increase in resonant frequency and deterioration of match. This would also imply that an unrecognised 'single-ended' mode exists when the parasitics are at their greatest in length.

Although this part of the investigation has shown that broadbanding was possible using a parasitic technique. There is a great need for further investigation into the parasitic behaviour including looking at the match of the simulated model, location of the parasitics and different topologies for the parasitics. One extension of this investigation is presented which focuses at the relocation of the parasitics with respect to coupling the parasitics more directly with the existing radiating section

3.4.1.1 Variance in Balun rim rotation

The second study to be undertaken with the parasitic elements was to analyse the effects of moving the parasitics closer to the original helical wires of the radiating section. Due to the uncertainty in the identification of the modes when the parasitic elements were initially introduced, the design with parasitics of 50% of the original length was utilised. This design only had two modes that could be clearly identified as the single-ended and balanced modes.

These parasitic elements were initially located at 90° rotation around the balun with respect to the original helical wires. The amount of rotation was then reduced in 15° steps so that the parasitic elements became closer to the radiating section, so that
direct coupling with the original helical arms would occur. 15° steps were chosen as then the parasitic helices would integrate easily into the balun mesh at set nodes, causing minimal modification to the original geometry. The return losses of this investigation can be visualised in Figure 3.22.

The introduction of the parasitic element caused a decrease in resonant frequency of the balanced mode by 100MHz. As the degree of rotation was decreased further frequency shifting of the balanced mode was minimal until the parasitics were moved within 45° of the radiating section. The resonant frequency then started to increase to return to its natural frequency of resonance when no parasitics are included into the design.

It can be hypothesised that the parasitic helices behave in a similar fashion to the helices of the original twisted loop in that the parasites helices also generate a virtual dipole. However, as the helices originate from the balun and are not driven from the top of the antenna the virtual dipole generated is weak in comparison to the balanced mode electromagnetic dipole. This parasitic dipole can disrupt the main electric field dipole of the twisted loop if it is placed too close to the twisted loop. Therefore
optimum placement in terms of minimum effect on the original twisted loop is a placement disposed 90°, this hypothesis is supported by the responses shown in Figure 3.22.

The single-ended mode was seen to increase as the rotation decreased until the 15° rotation was simulated. At 15° rotation, maximum coupling would occur between the radiating section and the parasitic elements. Three resonant modes were seen, one balanced mode, one single-ended and one other mode. This extra mode was a product of the coupling of the sets of helical arms.

Due to the small geometry of the dielectric-loaded bifilar the introduction of parasitics close to the original radiating section would require extreme precision in the manufacturing processes. Therefore this second study is more beneficial from a manufacturing tolerance or modelling perspective when progressing from an air-loaded to dielectric-loaded model if parasitics are introduced. It showed that parasitic elements appear to offer an opportunity to increase matchable bandwidth, although matched solutions have not been studied, the results did show that further investigation would be beneficial from the dielectric-loaded viewpoint.

### 3.5 Conclusions

The studies performed in this chapter demonstrated there is much research that can be performed on the twisted loop topology whether air or dielectric loaded. The parametric study performed on the air-loaded Bluetooth bifilar probed into the theory of the twisted loop. It was seen that although the balanced mode reacted to geometric changes as expected, the single-ended mode did not. This implied past publications were not entirely accurate in its description of the single-ended mode.

Clues to the true path were provided when modelling the twisted loop structure with and without the coaxial cable. Only when the cable was present could the single-ended mode be seen in the antennas response. Or as indicated in chapter 2, the single ended mode only occurs when the outer conductor of the cable is present. It was
shown that the outer conductor plays an important role in the current path loop in the single ended mode.

When the near fields of the primary modes of the twisted loop were investigated, once again the balanced mode behaved according to published theory. Some current was seen down the balun sleeves, which showed the antenna was not optimum in design and did explain slight movement of the resonant frequencies when geometric changes were made to the balun sleeve.

The single-ended mode displayed high field strength round the coaxial cable, demonstrating that as suspected current does flow back up the outer conductor. Therefore the true path of the current in the single-ended mode consists of the outer conductor and balun acting as a λ/2 resonator, and together with the helical arms and top tracks the single ended resonant loop is created. This modified current loop is presented in figure 3.23.

![Figure 3.23 Single-ended mode current flow](image)

Although the understanding of the current flow in the single-ended mode is important, it should be remembered the twisted loop antenna primary mode of operation is the balanced mode. When examining the far fields of the balanced mode it was found they could be represented using small electric and magnetic dipoles. In particular they could be represented using an electric and magnetic dipole both aligned in the same orientation as the virtual electromagnetic dipole in the bifilar antenna.
When operating efficiently the bifilar can simply be represented by a rectangular loop with a $180^\circ$ twist as the balun would be unnecessary. Therefore the far field patterns generated by the electric and magnetic dipole would resemble this ideal bifilar case.

Further more the strength of the fields of the small electric and magnetic dipoles dictate the appearance of the far field pattern. So changes in the far field pattern can be understood in terms of the effects either of the small dipoles. For example in the case of the dielectric-loaded twisted loop, the dielectric could affect the field strength of either the magnetic or electric dipole and this could be observed in the far field patterns. The work performed in this section can be used as a tool to fuel future research including looking at how the dominance of the dipoles are effected in different media; i.e. bifilars with ferrite cores instead of dielectric or varying dielectric constants, the optimisation of maximising the efficiency of the antenna. Looking at the trade off between efficiency and the specific absorption rate in antennas could also be performed.

Although the twisted rectangular loop may represent an ideal twisted loop antenna, the dielectric loaded bifilar has the advantages of the balun trap and the dielectric core that are invaluable to the antenna structure. The dielectric allows for the small size and containment of near fields, and the balun trap allows common mode noise rejection.

The twisted loop structure has been shown to have many advantages and very few disadvantages. One recognised disadvantage of the dielectric-loaded twisted loop is that its bandwidth in the balanced mode does not always cover the required modern communications band. To attempt to solve this problem two techniques were evaluated for their potential to broaden the bandwidth of the balanced mode.

The first technique involved flaring the helical wires of the radiating structure to form a bowtie structure, this method increased the bandwidth only slightly when substantial flaring was tried and therefore was deemed not fit for further investigation. The second technique involved the addition of helical wires to form additional resonant modes. The parasitic resonant frequency was expected to couple with the balanced
mode allowing the structure broadband characteristics. Past studies have had varying success by the introduction of additional resonant modes [11,12]. Werner et al. [12] found the use of 9 parallel vertical wires of different length, all fed from the same source was not suitable for wideband communication systems. Fortunately this was not the case when the parasitics were attached to the bifilar structure.

The simulated results indicated that the addition of the parasitics caused the balanced mode and the single-ended mode to decrease dramatically in resonant frequency. When the parasitics were reduced to 60% of their original size the balanced mode had shifted significantly in resonant frequency that it coupled with its single-ended mode giving a -3dB bandwidth of 500MHz, an increase of 250MHz. Although good bandwidth was attained the match of the response had deteriorated.

This did however prove that there was potential for broadbanding and further investigation into this work would be beneficial. Work to complement the air-loaded designs is currently in progress utilising the dielectric-loaded topology at the Centre for Mobile Communications Research, Loughborough University.

3.6 References


CHAPTER 4.0

GPS TWISTED LOOP ANTENNA

4.0 Introduction

In conjunction with the work performed on the bifilar twisted loop antenna in chapter 3, investigations were made into modelling air-loaded quadrifilar helix antennas. The quadrifilar helix antenna is a circularly polarised antenna suitable for GPS applications and can be recognised as a complex form of the twisted loop topology. With the FCC (Federal Communications Commission) regulations requiring half the handsets manufactured in the USA to have GPS capabilities by October 2001, research into the air-loaded quadrifilar helix is highly topical.

Research had been successfully carried in the past with dielectric-loaded quadrifilar helix antennas covering some of the theory and complications that arose in modelling the antenna using a TLM electromagnetic solver [1, 2, 3]. Publications have also highlighted the suitability of quadrifilar helix antennas for their integration into GPS applications [5, 6, 7, 8]. By adopting techniques from chapter 2 and taking into consideration the modelling issues that were addressed in past publications it was reasoned that the quadrifilar helix antenna could simply be modified into an air-loaded design. In section 4.2 the working theory for the quadrifilar helix antenna are reviewed and a preliminary design for an air-loaded Quadrifilar Helix Antenna (QHA) is proposed.

Examination of currents on the helical arms, and of the far field patterns, shows the phase quadrature condition that is essential for circular polarisation in the QHA was not established in the preliminary air-loaded design. In an attempt to achieve phase quadrature modifications to the air QHA structure were undertaken in the form of a crowned rim study also presented in section 4.2. Unfortunately it will be seen the
balun rim variations only accomplished limited success and therefore more in-depth investigation to obtain phase quadrature and circular polarisation was required.

Two studies are presented on basic antennas in section 4.3 to aid in the understanding of why phase quadrature was not realised in the preliminary air-loaded simulation and what modification would be required to rectify this defect. A turnstile antenna is examined with different ratios of arm lengths to slowly generate phase quadrature to analyse what transpires to the far field patterns before and after phase quadrature is accomplished. The turnstile was also used as a tool for generating a more refined feed system for the twisted loop antenna topologies. Essentially the quadrifilar helix antenna working properly in the axial mode can be considered as possessing a virtual turnstile antenna around the mid section of the radiating section due to its phase quadrature in the helical arms. So reduction of the quadrifilar helix to its basic form to further understand its phase quadrature was therefore a logical step.

The second study, which is recorded in section 4.3, was performed on two rectangular loops co-fed by a solitary source point. This functioned as a representation of the QHA in a simpler configuration, as in essence the QHA is two individual twisted loop antennas co-fed 90° out of phase with one another to obtain phase quadrature. The double loop antenna imitates the route propagated by the currents, though with the twist in the helices removed, if the QHA was functioning optimally. Electric far field patterns and impedance curves are used in the development of the characterisation of circular polarisation and utilised as a tool for the development of the QHA.

Having successfully completed the turnstile and dual-loop antennas the problems encountered in section 4.2 with the QHA are re-addressed and resolved in section 4.4. The air-loaded QHA was again attempted but with meander-lines used to generate the requisite phase shifting for phase quadrature. The modified simulated air-loaded QHA is then compared with measured results in section 4.5, right hand circularly polarised far field patterns are utilised for this comparison. It will be seen that the far field patterns correlate very closely with one another, and display the characteristic cardioid pattern displayed by the dielectric-loaded QHA and predicted by Kilgus for a wire QHA [9, 10, 11, 12], who pioneered the initial quadrifilar helix antenna investigations.
In section 4.6 parasitic helical arms, that were introduced in the twisted loop structure in chapter 3, are implemented onto the simulated air-loaded QHA in an attempt to broaden the antennas recognised narrow GPS bandwidth, or alternatively make the antenna multi-functional by allowing dual-band dual-polarised characteristics. This approach allows the antenna to be circularly polarised at the GPS frequency with a cardioid RHCP pattern, and to have linear polarisation in the PCN band with the characteristic virtual dipole pattern. Far field patterns, axial ratios and return loss plots will be used to examine the potential of this innovative design. Section 4.7 then summarises the chapter and provides conclusions.

### 4.1 Preliminary Model of the Quadrifilar Helix Antenna

#### 4.1.0 Theory of the Quadrifilar Helix

Kilgus C.C. originally proposed the quadrifilar helix antenna around the 1970’s [9, 10, 11, 12]. This topology has in recent years has been recognised for its suitability for GPS applications. Although there are now many forms of quadrifilar antennas generically they all employ two quadrature phased bifilar helical loops positioned orthogonally to each other; each individual loop following the same design principles as the twisted loop antenna. The structure of the dielectric-loaded quadrifilar helix antenna to be modelled in an air-loaded format is a more complex design than most published quadrifilars due to its integral balun and fishhook feed as shown in Figure 4.1.

![Figure 4.1 Structure of the dielectric-loaded quadrifilar helix antenna](image-url)
This novel quadrifilar helix antenna topology has been subject to considerable research using electromagnetic modelling and empirical techniques by Leisten et al [1, 2, 3]. It has been recognised that when modelling the antenna with a ceramic dielectric core it is a complex structure to simulate, therefore it would be beneficial to model the antenna using an air configuration. Once the air-loaded model has been established it could be utilised for fast modifications to the dielectric-loaded geometry for optimisation. The air-loaded model could also be used in understanding the issues bought about when manufacturing the antenna or in appraising new techniques to make the antenna multi-tasking.

As mentioned the quadrifilar helix antenna has two quadrature phased bifilar helical loops. This phase quadrature is achieved by the two loops having slightly offset resonant frequencies. One of the bifilar helices is adjusted longer than resonance to produce an input impedance with a phase angle of $+45^0$, and the other bifilar is adjusted shorter than resonance to produce an input impedance with phase angle of $-45^0$. This produces the $90^0$ phase shift between the currents in the first bifilar with respect to the second bifilar.

For the dielectric loaded Quadrifilar Helix Antenna (QHA) topology the line-length difference between the short and the long helix lines are realised by the use of a crowned rim balun. The two peaks of the crowned rim were used as the connections to the short helices and the two troughs utilised as the connections for the long helices. The use of a balun in the structure means a balanced feed signal is projected to the backfire tip of the antenna, and just as importantly the currents on the antenna element are isolated from handset. This isolation from the handset is just one of the advantages of this topology.

When operating in the axial loop mode of resonance, phase quadrature is achieved and the virtual electromagnetic dipoles formed by the two bifilars are $90^0$ out of phase and combine to give a spinning dipole effect. The conditions required for phase quadrature in the quadrifilar helix antenna are summarised in Figure 4.2. The effective spinning dipole means the antenna is circularly polarised and the characteristic cardioid far field pattern should be present.
The current propagation in the loop mode sees the current pass down one helix around the balun rim and back up the geometrically opposed helix to complete the 360° loop. This means at resonance the voltages should be equal in magnitude and opposite in phase approximately half way up the radiating section. Therefore, at this frequency 90° phase breaks in the currents of the individual helical tracks are expected.

Having briefly reviewed some of the geometric principles of the quadrifilar helix antenna, a preliminary air-loaded design can now be presented. The principles covered allow the reader a better insight into the quadrifilar helix antenna and demonstrates the differences and similarities compared to the twisted loop geometry.

4.1.1 Air-loaded QHA

Similar to the design of the twisted loop bifilar antenna presented in chapter 2, the air-loaded quadrifilar helix antenna was initially based on an industrial dielectric-loaded model shown in Figure 4.3. A crowned balun rim is utilised to induce the phase quadrature required for circular polarisation in the dielectric loaded antenna. The small dimensions of the dielectric-loaded quadrifilar helix antenna can be seen in Table 4.1.
Past publications [1, 2] have acknowledged the need for high resolution modelling when simulating this particular topology. This need is due to the small dimensions, the curvature of the surfaces and accuracy required in the geometry to obtain the phase quadrature. Although good agreement was observed with measured results when modelling a simple bifilar antenna, the intricacies that were necessary to attain phase quadrature in the QHA were unknown.

### 4.1.1.0 Air QHA Feed System

Because the quadrifilar helix antenna is complex in its structure, it might have been considered wise to model the air-loaded geometry without the coaxial feed system introduced in chapter 2 to keep the structure relatively straightforward. In similar fashion to that of the twisted loop structure, the removal of the cable could be considered permissible if the axial loop mode was the solitary mode of concern. In
accordance with this simplification the currents can still propagate around the twisted loops in the radiating section, around the balun rim and minimal current propagates down the balun sleeve. This would streamline the antenna structure and significantly reduce the computational time due to the reduced number of nodes required to generate the antenna configuration.

The disadvantage of the cable omission is some potential antenna modes that may prove useful in subsequent studies may not be excited. From chapter 2, it was seen the removal of the coaxial feed compromised the accuracy of the simulation of the twisted loop configuration. Therefore the coaxial fishhook feed through the centre of the QHA structure was retained.

A reasonable degree of confidence in the simulated air coaxial feed structure has been established in chapter 2 and also in chapter 3. The first method of proving the coaxial feed structure was by convergence testing; whereby the cable was repeatedly modelled with varying the number of nodes and wire formations until an accurate impedance plot was obtained. The second way that confidence was obtained in the coaxial structure was to verify its performance when introduced into the geometry, and also its behaviour in the near field analyses on the twisted loop structure. It was obviously to be expected that only in the single ended mode of the twisted loop did the cable feature as a major part of a resonating system.

The primary purpose of modelling the quadrifilar helix antenna was to create a tool to aid in the manufacturing of the dielectric-loaded quadrifilar helix and to put forward proposals for future generations of the design. This meant that accuracy of simulation and excitation of potential modes was important and so the inclusion of the coaxial cable feed system into the air QHA geometry was instigated. Another incentive for the inclusion of the cable feed at this early stage of the design concerned the intent that the antenna should be closely modelled on the dielectric-loaded model. This would ensure that the air-loaded QHA provided the accurate simulation that is desirable for exploitation as a fast predictive tool.
4.1.1.1 Core Antenna Structure

When morphing from the dielectric-loaded design to the air-loaded geometry, the same techniques that were developed when modelling the twisted loop structure were applied. MiniNEC Professional for Windows was used to model the structure using the Method of Moments numerical electromagnetic solver. The balun was constructed using 24 wires for the rim and base of the balun sleeve with 12 vertical wires completing the balun sleeve structure. The base of the balun used radial wires at every 60° meshed with a 12 wire circle located halfway between the balun base’s midpoint and the balun sleeve base. This bazooka balun structure had been realised using convergence testing in the impedance curves when modelling the bifilar antenna. The procedure was to examine the effects of variation of the number of nodes and wires in the structure, until a consistent impedance curve was achieved.

The rim of the balun was altered by its vertical component to obtain the appropriate heights for a crowned rim representation. The helices in the radiating section utilised 20 wires for numerical and visual accuracy as demonstrated in the twisted loop topology. At the top of the antenna structure the helices were paired together, with one helical arm from bifilar 1 being paired with a helical arm from bifilar 2 and the two remaining helical arms being connected together. One pair of the helices were then attached to the outer conductor of the coaxial cable developed in Chapter 1, with the other pair of helices being connected to the fish hook feed. Once again a voltage gap excitation was utilised at the bottom of the coaxial cable to feed the structure. This is illustrated in Figure 4.4, which shows the preliminary air-loaded geometry.
In a similar way to the procedure that was followed with the twisted loop antenna, the exact resonant frequency shift in morphing from a dielectric-loaded to air-loaded design was unknown. As a first approximation the dimensions of the air-loaded QHA were rescaled using the $\sqrt{\varepsilon_r}$ multiplication factor. Scaling to the true GPS frequency was not essential at this stage as attaining phase quadrature and the characteristic Right Hand Circular Polarisation (RHCP) pattern was considered a greater priority. The dimensions for the air-loaded QHA together with their dielectric-loaded counterpart are detailed in Table 4.2.

<table>
<thead>
<tr>
<th>ANTENNA PARAMETER</th>
<th>DIELECTRIC QHA DIMENSIONS</th>
<th>AIR QHA DIMENSIONS</th>
</tr>
</thead>
<tbody>
<tr>
<td>Total height</td>
<td>17.75 mm</td>
<td>106.50 mm</td>
</tr>
<tr>
<td>Diameter</td>
<td>10.00 mm</td>
<td>60.00 mm</td>
</tr>
<tr>
<td>Axial length of long helix</td>
<td>12.40 mm</td>
<td>74.40 mm</td>
</tr>
<tr>
<td>Axial length of short helix</td>
<td>12.30 mm</td>
<td>73.80 mm</td>
</tr>
<tr>
<td>Balun crown height</td>
<td>0.10 mm</td>
<td>0.60 mm</td>
</tr>
<tr>
<td>Balun height</td>
<td>5.35 mm</td>
<td>32.10 mm</td>
</tr>
<tr>
<td>Dielectric constant</td>
<td>36</td>
<td>1</td>
</tr>
</tbody>
</table>

Table 4.2 Dimensions of preliminary air-loaded QHA
The complete structure was built-up of 307 wires and 1393 nodes, with many of the nodes being dedicated to the coaxial cable feed structure. A frequency range of 0.5 GHz to 2.8 GHz was used requiring 575 steps of 4 MHz. This model was crafted so that it could provide adequate accuracy with about 24 hours of run-time. The impedance curves for preliminary design can be seen in Figure 4.5.

From Figure 4.5 it is not immediately apparent as to the location of the frequency at which the antenna would resonate with circular polarisation. It is known the antenna was scaled using a $\sqrt{e}$ factor, and when the same scaling factor was used on the twisted loop topology it resonated at a much lower frequency than the antenna was designed for. From theory, the resonant frequency of the circularly polarised mode should be found between the resonance's of the two bifilar loop modes. Therefore it was considered that the circularly polarised mode was the resonant mode that existed around 930 MHz. By examining the right hand circular far field patterns and the currents on the helical tracks it was anticipated that this could be confirmed, providing the geometry was accurate in its structure.

Closer inspection of the far field RHCP patterns revealed that none of the resonant modes exhibited the expected characteristic cardioid configuration; furthermore at
930 MHz the current phases half way up the radiating section were all within 20° of each other. This meant the conditions for phase quadrature had not been satisfied and the antenna would not exhibit circular polarisation. The far field circular polarised patterns at 930 MHz, given in Figure 4.6, were similar in their formation to the linear polarised pattern of the twisted loop structure in chapter 1.

![Figure 4.6 RHCP Pattern of Air QHA at 930 MHz](image)

The phase quadrature in the dielectric-loaded quadrifilar antenna was obtained by varying the lengths of the two bifilars within the structure with respect to one another by altering the height of the balun crown. As the scaled balun height in the air-loaded quadrifilar helix antenna appeared not to give the requisite phasing on the helical
wires, a parametric study was undertaken analysing the change in phase with respect to the balun crown height.

The balun crown was altered from the scaled 0.6mm height to a maximum height of 10.0mm using steps dependent on the phases obtained from the subsequent simulations. The various balun crown heights investigated with the difference in phases of the helical tracks are shown tabulated in Table 4.3. If phase quadrature was present then the difference between the first and last track should be $270^\circ$ based on a $90^\circ$ phase change between the tracks for phase quadrature, illustrated previously in Figure 4.2.

<table>
<thead>
<tr>
<th>TOTAL BALUN CROWN HEIGHT (mm)</th>
<th>PHASE DIFFERENCE BETWEEN TRACKS (Degrees)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.6</td>
<td>20</td>
</tr>
<tr>
<td>1.8</td>
<td>10</td>
</tr>
<tr>
<td>2.7</td>
<td>10</td>
</tr>
<tr>
<td>3.6</td>
<td>20</td>
</tr>
<tr>
<td>5.0</td>
<td>30</td>
</tr>
<tr>
<td>7.5</td>
<td>40</td>
</tr>
<tr>
<td>10.0</td>
<td>40</td>
</tr>
</tbody>
</table>

Table 4.3 Phase Difference in Radiating Section Due to Varying Crown Height

The Right and left hand circularly polarised far field patterns for each of the models all displayed a linear polarised pattern rather than the characteristic cardioid pattern. Table 4.3 shows that altering the crown height of the balun had little effect on the phases of the helical tracks. Therefore the geometry was reassessed in order to attain the required circular polarisation.
Firstly the cable was removed from the geometry and the antenna fed by a simple voltage gap on a wire combining the pairs of helices. The geometry was therefore in its most simple form employing 183 wires and 474 nodes for the antenna structure. Unfortunately the phase differences on the helical tracks remained similar in value with respect to one another and the far field pattern remained linearly polarised in appearance rather than circularly polarised as desired. This indicated a more extreme method was required for generating the phase quadrature in an air-loaded geometry other than the crowning of the balun rim.

Prior to implementing a method for inducing greater phase differences on the helical tracks, it was considered beneficial running simulations focused on generating circular polarisation in more simplistic and documented circularly polarised antenna models. This would aid in obtaining a more robust feed system and spotlight key aspects requiring consideration when generating circular polarisation.

4.2 Circular Polarisation Characterisation

Identification of the circular polarisation in terms of a characteristic impedance curve was unfamiliar; therefore two studies were performed to aid in the design of the air-loaded quadrifilar helix. The first investigation would look at the generation of circular polarisation by use of turnstile geometries and identify characteristics associated with the design in the form of RHCP patterns and impedance curves.

The second investigation performed was similar to one undertaken in chapter 3 for the twisted loop. The air quadrifilar helix antenna is considered and modelled in terms of its current loop paths to generate the quadrifilar axial mode using current phases, impedance curves, far field RHCP patterns and axial ratio to identify the resonant mode.

4.2.1 Turnstile Antenna

As mentioned earlier in the chapter, the quadrifilar helix antenna utilises two bifilars in phase quadrature to generate circular polarisation. Alternatively the circular
polarisation generated may be considered as two virtual dipoles, created by the two bifilar twisted loops that construct the quadrifilar topology, $90^\circ$ out of phase with one another to produce a virtual spinning dipole. This phasing of two dipoles to engender circular polarisation could be considered as a crude form of a turnstile antenna. Therefore by replacing the virtual dipoles for real dipoles, the first investigation looks at the formation of circular polarisation using a turnstile topology.

Two dipoles of equal lengths $\pm 4.75\text{cm}$ were joined together at their mid points, with the structure lying in the XY-plane. The two dipoles are modified at their mid points so that the structure only requires a single excitation as shown in Figure 4.7. This single feed source was utilised so the turnstile antenna had similar excitation requirements to the air quadrifilar helix. By refining the feed configuration on the turnstile antenna, a similar feed system could be used on the air-loaded quadrifilar helix antenna geometry.

The approach of generating circular polarisation in this fashion rather than simply using a dual feed system for the excitation was also to keep the concepts of the turnstile and air-loaded QHA comparable. That is to say, in both antennas the increase in current path lengths created the conditions required for circular polarisation to exist.

![Figure 4.7 The Turnstile Configuration](image-url)
Equal lengths were initially chosen for the four arms constructing the turnstile antenna to identify the characteristic far field pattern when circular polarisation was not achieved, and to analyse changes that transpired in the development of circular polarisation when the geometry was modified. The pair of lengths creating one of the dipole arms were then increased in steps of 0.25cm with the impedance curve and RHCP patterns monitored at each stage.

The dipole aligned along the x-axis was increased from ±4.75cm to ±5.75cm whilst the dipole aligned along the y-axis was maintained at ±4.75cm in length. When the dipoles are of equal length the RHCP far field pattern was comparable to the pattern displayed by the preliminary air QHA, as shown in Figure 4.8. The LHCP far field patterns of the two antennas were surprisingly very similar in profile also.

Figure 4.8 Comparison of RHCP and LHCP patterns for (a) the Turnstile Antenna comprised of identical length dipoles and (b) the Air QHA
A selection of impedance curves is given in Figure 4.9 to show the trend of the turnstile antenna as the length of its x-axis dipole is increased. An additional impedance curve is also included where the x-axis dipole length was $\pm 5.2\,\text{cm}$ and the y-axis dipole was $\pm 4.75\,\text{cm}$. At these arm lengths the reactance curve flattens briefly to a constant value due to the relative values of the individual dipoles. The values in the legend in Figure 4.9 represents the length extended to by the modified arms from the starting length of 4.75cm.

![Impedance of Turnstile Antennas with Varying Ratio of Arm Lengths](image)

**Figure 4.9 Impedance curves of the Turnstile antenna.**

The RHCP and LHCP patterns, current phases and axial ratio at resonance was monitored for circular polarisation; Figures 4.10 and 4.11 shows the circularly polarised plots obtained when the arms of the turnstile antenna were all the same size and also when the arms were at the length which gave rise to the reactance plateau observed in Figure 4.9. The circularly polarised patterns are not sufficient enough to represent how good the circular polarisation is as a ratio of orthogonal field components is required, therefore axial ratio magnitude and phase was monitored. Also for this application the shape of the far field pattern also needed to be considered as set properties for satellite reception were sort after; i.e. the cardioid far field pattern.
Figure 4.10 RHCP and LHCP patterns of the Turnstile antenna with all arms 4.75cm in length

Figure 4.11 RHCP and LHCP patterns of Turnstile antenna with two arms 5.2cm in length, and two arms 4.75cm in length

Figure 4.11 shows the expected characteristic cardioid pattern for the turnstile antenna in phase quadrature. The quadrature phasing of the currents on the individual arms, an axial ratio magnitude of unity and angle of $90.44^0$ supported the attainment of circular polarisation, compared to the non-circularly polarised turnstile that had axial ratio magnitude of 0.999 and angle of 179.97 and no phase quadrature in the arms. When scrutinised closely the plan views of the circular polarisation plots show the cardioids
are not perfectly rounded, this imperfection was expected and has been acknowledged in turnstile theory by Kraus [13] when recounting the George Brown turnstile antenna [14, 15].

For the turnstile antenna a considerable increase of length in one of the dipoles was required to induce circular polarisation. If such an increase were necessary in the air QHA, an increase in the balun height would be unreasonable and could explain the linear polarisations that were previously seen. This crystallises the need for a more elegant solution to lengthen the current path in one of the bifilar twisted loops in the air QHA.

Before introducing a new strategy to encourage phase quadrature in the air QHA, the success of generating circular polarisation in the turnstile antenna needed to be emulated in an antenna geometry closely related to the quadrifilar helix antenna. This would make the transition from a relatively simplistic design to a complex geometry progressively more straightforward.

4.2.2 Double Loop Antenna

In order to further develop the understanding of the air-loaded quadrifilar helix antenna it was helpful to consider a topology which presents an intermediate step between the turnstile antenna and the full air loaded quadrifilar helix antenna. This topology can be called the double loop antenna and as the name suggests it comprises of two rectangular loop antennas that are orthogonal to each other. These are co-fed by a common voltage gap excitation located at the mid points at the top of the loop topologies as indicated in Figure 4.12.
The structure could be considered an extension of the turnstile structure where the dipoles have been extended to form loop antennas. This structure was employed because the current propagation around loop paths is similar to that of the quadrifilar helix antenna resonating in the axial mode, merely with the exclusion of the half turn twist that is present in the quadrifilar artwork.

In the course of attempting to obtain circular polarisation with the double loops, the widths of the two loops were fixed to preserve the overall diameter of the antenna structure, this was to reflect the existence of the inflexible dimensions of the dielectric core that is present in the case of the dielectric loaded volute. Thus increasing the length in one of the path loops would be responsible for the induction of circular polarisation.

To be scrupulous, the width of the current loops in the DLQHA are not sustained throughout the structure due to the crowned balun rim creating differential size in the widths and lengths of both loops that form the structure. Unfortunately, as seen earlier, phase quadrature was not accomplished in the air-loaded geometry as a consequence of the small paths differences produced by the crowned rim in the air QHA. Therefore by using the simplistic double loops structure and modifying just one
parameter, i.e. the length of one loop, the ratio of the path lengths required to produce circular polarisation could be determined.

Utilising this type of antenna and procedure to obtain circular polarisation would demonstrate how the quadrifilar helix antenna, operating in the axial mode, could be modelled as an air-loaded structure using a single feed system. This technique would eliminate the requirement for the crowned balun rim that showed limited success in obtaining phase quadrature. Past publications have modelled simplistic quadrifilar helix antennas and took advantage of a double feed to obtain phase quadrature [16], which although beneficial from a modelling prospective does not address the application of phasing when implementing this more simple feed design.

The double loop antenna was initially simulated as two rectangular loops with 6.0cm width and 7.41cm sides to give the same height and width of the radiating section as the air QHA. The double loops were co-fed with a single voltage gap excitation. Steps of 0.5cm were then used to increase the height of one of the loops, to give increments of 1.0cm path difference between the loops with each step until the characteristics of circular polarisation were identified. The lengths of the modified loop were then finely adjusted to optimise the circular polarisation by monitoring the currents, axial ratio, impedance curves and far field patterns of each simulation.

As the path lengths of the one loop increased with respect to the other, similar trends in impedance to that of the turnstile antenna were observed. A plateau in the reactance once again emerged as the antenna developed from being linearly to circularly polarised. The circularly polarised condition was identified from the axial ratio becoming unity and the currents on the loops exhibiting phase quadrature with respect to each other. Additionally the far field radiation patterns under went a metamorphosis from the ‘linearly polarised’ pattern to the characteristic circularly polarised cardioid pattern. The far field RHCP patterns of the initial double loops antenna and the final double loop antenna design are shown in Figure 4.13.
Figure 4.13 RHCP far field patterns of (a) initial and (b) final double loop antennas

The increase in length required to generate the circular polarisation was 1.0cm in each of the vertical sides of the modified loop. This value presumably reflects the amount
of crown that would be required if circular polarisation were to be accomplished using the crowned balun rim in the air QHA. The axial ratio magnitude was 1.002 with an angle of 82.92 for the circular polarised double loops. Whilst the loops of equal length that did not display the characteristic RHCP far field pattern had an axial ratio of 0.999 and angle of 180.00. Although the dimensions and structure of the double loop antenna has not replicated the structure of the air QHA, it has shown the crowned rim to be less than ideal for inducing circular polarisation and demonstrated that circular polarisation could be generated with the twisted loop structure. Therefore a different geometric modification is demanded to bring about the relevant path differentials in the air QHA design.

4.3 Meander-line Air-loaded QHA

Circular polarisation had been successfully acquired using two different intermediate antenna structures and a better understanding of the requisite conditions had been obtained. The simulated double loop antenna resembled the current path for the axial mode of the air quadrifilar helix antenna, and illustrated why the use of a crowned balun rim was unsuitable for this type of air QHA owing to the relative current path lengths of the two loops.

Consideration of the double loop antenna has made researching the air-loaded QHA for its appropriate functionality more fulfilling. As the crowned balun rim only had limited success it was rejected as a method for phasing the air loaded antenna and therefore the balun rim was maintained to be flat at a constant height. Similar to the course of action taken with the double loops, the length of one of the loops in the air QHA was modified to create the phase difference to generate circular polarisation. Unlike the double loop antenna the height of the radiating section in the air QHA is a constant parameter, therefore the additional length was introduced into the air QHA structure by meandering the sides of one of the twisted loops.

At this stage of the research, characteristic of the present changing market, the dielectric-loaded quadrifilar helix antennas specification had been slightly modified.
As the research performed was designed to aid the industrial model, the target research model was changed slightly to adopt the modified dimensions that are given in Table 4.4.

<table>
<thead>
<tr>
<th>ANTENNA PARAMETER</th>
<th>DIELECTRIC QHA DIMENSIONS</th>
</tr>
</thead>
<tbody>
<tr>
<td>Total height</td>
<td>17.750 mm</td>
</tr>
<tr>
<td>Diameter</td>
<td>10.000 mm</td>
</tr>
<tr>
<td>Axial length of long helix</td>
<td>12.335 mm</td>
</tr>
<tr>
<td>Axial length of short helix</td>
<td>12.305 mm</td>
</tr>
<tr>
<td>Balun crown height</td>
<td>0.030 mm</td>
</tr>
<tr>
<td>Dielectric constant</td>
<td>36</td>
</tr>
</tbody>
</table>

Table 4.4 Dimensions of modified dielectric-loaded QHA

The preliminary air-loaded quadrifilar helix utilising the original specifications was adopted for the initial implementation of the meander-line helical wires. The crowned rim was maintained at a height of 0.36cm, after consultation with Dr. Agboraw revealed that dielectric-loaded modelling utilised a crown height of 0.06cm and not the 0.01cm height as given by the specified dimensions for the actual antenna.

The helical wires in the original air-loaded design were comprised of 20 wires using 21 nodes to define their location. The nodal points could be generated mathematically assuming the helical wire is subdivided in 20 equivalent parts and by use of the total length and pitch angle. The helical wire is first represented vertically in the Z-axis in a planar view, and then rotated to the specified pitch angle. Finally the nodal coordinates are converted so they are positioned around the air-loaded core cylinder.

The development of this mathematical procedure meant the generation of the nodal points for the meander-line helical wires became straightforward. Once again 21 nodal points were employed defining the start, finish and central points of the meanders. The use of 21 points ensured starting and finishing on the central point of a
segment as well as defining the helix accurately. The calculated nodal points could then be entered into the simulator, or modified if the model necessitated rescaling to acquire the requisite GPS resonant frequency.

To generate the meander-line, the start and finish of each meander would undergo a horizontal displacement of ±x in the initial stage of the mathematical procedure. Thus giving a regular meander to the helical wire before inclination to the appropriate pitch angle and rotation around the cylindrical core. The generation of the planar view meander-line is illustrated in Figure 4.14.

![Figure 4.14 Generation of the meander-line before rotation around air cylinder](image)

If the quadrifilar were thought of as two bifilars, with one of the bifilars having meander-line helical wires, then increasing the x-axis displacement of the meander-lines in the mathematical procedure just described would increase the loop path of the one bifilar with respect to the other. From the knowledge attained from the turnstile and double loop antenna, at a specific meander-line displacement the currents on the helical wires would exhibit phase quadrature and the air QHA antenna would be circularly polarised.
Using the preliminary air-loaded model with a 0.36cm crowned rim, the meander line was implemented into the quadrifilar on two geometrically opposing helices, as demonstrated in Figure 4.15. The initial displacement of the meander line was adjusted in small steps to induce the circular polarisation characteristics. The trend seen in the impedance curves for the different meander-line displacements are shown in Figure 4.16 and Figure 4.17. The characteristic formation of the reactance plateau, similar to those observed in the turnstile and double loop antennas, were seen for a meander-line displacement of ±0.16cm.
Figure 4.16 Resistance Curves for Air QHA with Varying Meander Displacements

Figure 4.17 Reactance Curves for Air QHA with Varying Meander Displacements

When the air-QHA, with the 0.16cm meander-line helical wires, was scrutinised at the frequencies at which the reactance plateau existed, the currents on the helical wires were in phase quadrature and a well formed cardioid profile was observed as the right hand circular far field pattern with the boresight of the pattern in the zenith.
In contrast to the previously modelled circularly polarised antennas, the left hand circular far field pattern was negligible in the air-QHA. This diminutive left hand circular polarised pattern is a recognised attribute of this class of antenna, and comes about as a result of its integral balun and clockwise twisted helical wires whose structure emulates a travelling wave antenna with right hand circular polarisation. In this manor the left hand circular pattern is suppressed with respect to the right hand polarisation. The left and right hand circular polarised patterns can be seen in Figure 4.18. Comparisons with measured results can be seen later in the chapter after the air QHA antenna has been modified to the new specifications.

Figure 4.18 Meander-Line QHA (a) Right Hand and (b) Left Hand Circular Far Field Patterns
4.3.1 Air QHA with Modified Specification

Having successfully attained the characteristic right hand circular far field pattern modelling the air QHA utilising meander helical arms and a $\sqrt{e_r}$ scaling factor, attention was turned to the new specifications. It is interesting to see whether these successful simulation exercises could be extended to improve the performance or manufacturability of the antenna.

After consultation with Dr. Leisten it was discovered the resonant frequency of the dielectric-loaded antenna was set to be 18MHz above the GPS centre frequency of 1575.42MHz. The higher frequency of 1593.42MHz was to allow for a weather seal cap to be placed over the antenna, as when the cap is placed over the dielectric-loaded antenna the resonant frequency drops by 18MHz to the requisite GPS frequency. To keep a close connection between the research and the industrial designs this new resonant frequency was also adopted for the air-loaded model.

The modelling of the air-loaded QHA earlier in the chapter showed the crowned rim to have negligible effects on the phase quadrature. The crowned balun rim was therefore not applicable for the air-loaded experiments and phase quadrature was generated solely using the meander line helical wires in the new design. The radiating section length was taken to be the height of the longest axial radiating height (12.335mm before rescaling), so all path length changes would occur above the balun rim and not include the balun rim. Therefore for both the simulation artwork and the original design no path changes occur below this axial height. Consequently the balun rim was maintained at a constant dielectrically loaded equivalent height of 5.415mm.

Once again the displacement was increased in the zigzag helical wires to obtain the correct meander in the helical arms to generate the circular polarisation. Then similar to the procedure with the twisted loop antenna in chapter 2, the air QHA was rescaled so the reactance plateau, seen in the previous impedance curves, was centred on the requisite resonant frequency. The final dimensions used in the simulated artwork are
shown in Table 4.5 together with the new specifications of the dielectric-loaded model.

<table>
<thead>
<tr>
<th>ANTENNA PARAMETER</th>
<th>DIELECTRIC QHA DIMENSIONS</th>
<th>AIR QHA DIMENSIONS</th>
</tr>
</thead>
<tbody>
<tr>
<td>Total height</td>
<td>17.750 mm</td>
<td>57.5008 mm</td>
</tr>
<tr>
<td>Diameter</td>
<td>10.000 mm</td>
<td>32.3948 mm</td>
</tr>
<tr>
<td>Axial length of long helix</td>
<td>12.335 mm</td>
<td>39.9590 mm</td>
</tr>
<tr>
<td>Axial length of short helix</td>
<td>12.305mm</td>
<td>39.9590mm</td>
</tr>
<tr>
<td>Balun crown height</td>
<td>0.030 mm</td>
<td>0.0000 mm</td>
</tr>
<tr>
<td>Meander-line Displacement</td>
<td>Only Straight Helical wires</td>
<td>1.9000 mm</td>
</tr>
<tr>
<td>Dielectric constant</td>
<td>36</td>
<td>1</td>
</tr>
</tbody>
</table>

Table 4.5 Dimensions of Dielectric-loaded model and Final Air-loaded design

From the scaled dimensions in Table 4.5, it is seen that an ultimate scaling factor of 3.23948 was required when developing the air-loaded artwork from the dielectric-loaded model specification. It is also of interest that a slightly coarser meander-line is required to account for the exclusion of the balun crown. The final air-QHA model is presented in Figure 4.19 and was comprised of 812 nodes and 305 wires. The reduction in size of the model meant the number of nodes used in the definition of the model declined from 1393 nodes to 812 nodes, which meant simulation runs were more computationally efficient in terms of memory required and simulation time.
The approach of using rescaling and the generation and implementation of meander lines was found to be time consuming in the construction of the simulation geometry. Therefore the approach taken by the author was to simulate the geometry with meander lines, though not tailored to induce the circular polarisation. From the impedance curves of the simulation and the knowledge of the impedance behaviour with varying meander-line displacement from Figure 4.16 and Figure 4.17, an approximate resonant frequency for the circular polarisation could be determined when the meander lines are modified further. The model is then rescaled to obtain the approximate requisite frequency, and then the meander line modified to induce circular polarisation.

This procedure was adopted because when the model is rescaled after obtaining the circular polarisation it was found the reactance lost its characteristic plateau, and the right hand circular cardioid pattern lost some definition in the form of front-to-back ratio. Consequently further meander line manipulation was required to regain the definition of the far field pattern and the ideal axial ratio magnitude of 1.

For the given dimensions in Table 4.5 the reactance plateau, shown in the impedance curves, covered the frequency range 1585MHz to 1615MHz. At 1593MHz the axial ratio magnitude was calculated to be 1.012 with phase of 92.59. A very good right hand circular far field cardioid pattern was observed with $-40\text{dB}$ front-to-back ratio.
The 2-dimensional principal plane cuts of the far field patterns are shown in Figure 4.20 and 4.21. Once again, good right hand circular patterns were attained alongside expected suppressed fields for the left hand circular components. As expected these far field patterns showed good resemblance to the patterns attained from the Air-QHA built with the old specifications. The additional electrical length brought about by the meander line to induce the circular polarisation was found to be 1.95cm.

Figure 4.20 Right Hand Circular Field Components for the Simulated Air-QHA based on New Specifications
4.4 Comparison of simulation with measured results

Having developed an air quadrifilar helix antenna that utilises meander line helical wires in the radiating section to generate circular polarisation, a practical model was constructed to verify the right hand circular patterns observed in the simulation. The hand built air loaded model utilised thin film copper laminate on PTFE to realise rigidity that is necessary to support the model structure. This material was used in the main cylindrical part of the antenna and the top of the antenna.

The base of the balun was constructed of single sided copper plated FR4; this allowed a solid base for the antenna to be wrapped around and impart cylindrical form and further rigidity in the antenna structure. The FR4 and PTFE both had low dielectric constants to have minimal influence on the air loading of the antenna. A 50Ω cable was then inserted axially through the centre of the antenna, with a fishhook feed connection at the top of the antenna topology. The air-loaded model is pictured in

![Figure 4.21 Left Hand Circular Field Components for the Simulated Air-QHA based on New Specifications](image-url)
Figure 4.22 alongside its dielectric-loaded counterpart; this gives a good visual representation of the miniaturisation that occurs due to the dielectric core.

![Image of Air and Dielectric Loaded QHA](image)

**Figure 4.22 Air and Dielectric loaded QHA**

The right hand circular patterns were measured for both the air and dielectric loaded antennas and then compared with the predicted patterns from the simulation. The far field patterns for the two antennas are shown in Figure 4.23 and 4.24 together with the frequencies at which the patterns were acquired. In both cases well-formed cardioid patterns were observed, showing good agreement with the simulated model. The difference in the resonant frequency between the air and dielectric loaded models are mainly due to the dielectric model having a weather seal cap that causes a reduction in resonant frequency of 18MHz to the exact requisite GPS frequency.

![Image of RHCP Far Field Patterns of the air QHA](image)

**Figure 4.23 RHCP Far Field Patterns of the air QHA (1602.35MHz)**
The zenith gain of the air-loaded quadrifilar helix antenna was seen to have a superior gain by 5.22 dB, which is attributed to the reduction in size of the antenna and the dielectric loading. The front-to-back ratio was approximately 15 dB in the case of the air-loaded antenna and over 60 dB in the case of the dielectric-loaded antenna. The improved front-to-back ratio seen in the dielectric QHA is probably due to laser tuning that optimises the phase quadrature in the helical arms.

Antenna engineers familiar with GPS antennas for mobile communications may correctly perceive the zenith gain for the dielectric-loaded antenna in Figure 4.24 is lower than the gain of a patch antenna measured in free space. Although this is true in free space, the dielectric-loaded QHA shows superior performance to the patch antenna in a handheld environment. It also displays a number of geometrical and mechanical advantages over the patch antenna and other existing antennas for mobile GPS devices as demonstrated by the author [6, 17], this included the performance of a patch antenna and a QHA antenna in handheld environments. Other publications that incorporate human proximity effects on Handset antennas are becoming more commonplace. Human interaction with circularly polarised antennas for satellites communications have also been investigated [16, 18], which see deterioration in the far field patterns due to a human head in close proximity to the antenna.
General advantages and disadvantages of using quadrifilar helix antennas, compared to other antennas for GPS applications have been published by James, J.R. [19, 20] and Kraus, J.D. [21]. Aside from the QHA for GPS applications Padros, et al. [22] has performed a comparative study on other antenna designs available for receiving GPS and shows a conical spiral with absorbing material out performing the industrially well-utilised patch antenna.

### 4.5 Insertion of Parasitics into Air-QHA

During this period of the research, one of the problems encountered with the dielectric-loaded QHA was the narrow bandwidth covered at resonance. Although it covered the GPS bandwidth, it did not allow the manufacturer much margin for error in the production of the antenna. Therefore if the artwork was slightly incorrect and the resonant frequency was slightly higher or lower than required, then the GPS bandwidth would not be covered by the response of the antenna.

If the antennas response could be improved by increasing the operational bandwidth, the antenna could still cover the GPS band if the artwork on the ceramic core was not precise. This would make mass production of the dielectric-loaded QHA more simplistic and a lower percentage of the antennas produced would be inoperable or require tuning. Therefore in an attempt to aid the manufacturing of the dielectric-loaded antenna by trying to increase the operational bandwidth, the effects of introducing parasitics to the air-QHA were investigated.

Although more sophisticated software was available to model the dielectric-loaded antenna if desired at this time in the research, MiniNEC was once again used for the preliminary investigations of the study. The use of MiniNEC would quickly determine if the insertion of parasitics warranted further and more intense investigations by implementation into software with dielectric loading capabilities (CST’s Microwave Studio showing realism and therefore scoring as the most promising of the packages tested). This would also allow time for a thorough examination of CST’s Microwave Studio as well as other electromagnetic simulators in their ability to model the
dielectric-loaded twisted loop antennas, and for the refinement of the dielectric geometries that were already modelled.

When the parasitics were included in one of the earlier Bluetooth twisted loop designs in chapter 3, they were positioned half way between the two helical wires forming the radiating section. The air-QHA has a radiating section comprised of four helical wires, two straight and two meander helical wires, therefore four parasitic helical wires were utilised in the GPS investigation.

The four parasitics were placed around the balun rim, half way between the existing helical wires of the radiating section. Each of the parasitics had the same pitch angle as the existing helices so they remained equidistant with the helices as they progressed up the radiating section. This procedure was adopted to maintain a degree of symmetry about the structure, and to keep the influence the existing radiating helical wires have from the parasitics relatively consistent with respect to one another. A visualisation of the GPS air-loaded quadrifilar helix antenna with parasitics can be seen in Figure 4.25.

![Figure 4.25 Air Quadrifilar Helix with 50% Parasitics](image)

Figure 4.25 shows the parasitic helices originating from the balun and finishing open-ended pointing towards the top of the antenna structure. Twenty wires were used to
provide good definition and numerical accuracy to the initial full-length parasitics. This definition in the parasitic helices would mean the same resolution is used for the parasitics and the original helices in the radiating section.

The parasitic helices were then gradually reduced in size by removing a wire from each of the parasitics at their open end. Therefore for each wire removed from a parasitic helix a reduction in its size of 5% would be experienced. This is similar to the procedure performed in the parasitic study with the twisted loop bluetooth design in chapter 3.

The complete structure with the parasitics at their maximum height required 385 wires and 933 nodes. This equates to over 80 supplementary nodes being inserted into the structure for the additional four helices, this was mandatory so the parasitics could be incorporated into the design with no simulation guidelines broken. From the bluetooth study with the parasitics, it was found the parasitics had maximum influence on the resonant frequency of the antenna when they were at around 50% to 60% of their original height. Therefore the parasitic helices on the air-loaded GPS antenna were initiated at 70% of their full height and then reduced in 5% steps. For each new simulation the S11 response and far field patterns were recorded and analysed.

The S11 for a sample of the simulations carried out, with parasitics helices ranging from 70% to 50% of their original height, can be seen in Figure 4.26 and 4.27. The S11 of the air QHA with no parasitics is also included for the comparison. The parasitic helices were of constant height with respect to one another for each simulation. The constant height was so the resonant modes brought about by the parasitics were all at the same frequency, and so fewer variables were present in the simulations performed for this preliminary investigation.
The insertion of the 70% parasitics into the GPS structure causes the resonance of the axial mode to increase by over 50MHz in frequency as seen in Figure 4.27. As the parasitic helices decrease in length, its resonant frequency increases and forces the axial mode even higher in frequency. As the parasitic helices were decreased to
around 55% to 60% of their original size, maximum coupling with the axial mode was observed. Further simulations with parasitics around this length would be recommended to attain optimum coupling for broadbanding.

As the parasitic helices are reduced further in length, the resonant frequency of the parasitic mode increased further and the two resonant modes became distant to each other once again. When the parasitics were reduced in length by 50% or more, the GPS axial mode began returning to the resonant frequency 1593.42MHz for which the air QHA was designed.

One of the concerns about introducing parasitics into the quadrifilar helix design was the influence they would have on the right hand circular far field pattern. If the characteristic cardioid pattern was distorted by the introduction of the parasitics, then beamwidth coverage and gain of the RHCP far field radiation pattern of the antenna could be compromised. Thereby frustrating the advantages a quadrifilar helix generally possesses over other antennas for GPS applications.

Another concern arising from the introduction of the parasitics was the magnitude of the axial ratio. Supposing the characteristic cardioid pattern remained relatively unchanged, if the parasitics presence affected the axial ratio considerably, then the antenna would no longer be circularly polarised and be less optimal for GPS. Therefore the far field patterns and axial ratio at the resonant frequency of the axial mode required monitoring.

To analyse the effects of the parasitics, the far field patterns are observed for two case scenarios. Firstly, when the parasitics are 60% of their original length and very close to coupling with the axial mode and secondly when the parasitics are at 50% of their original length and have moved further away from the axial mode. Although the second case study does not permit broadbanding, it does highlight the effects of the parasitic helices on the axial mode and introduces the possibility of making the antenna dual band due to the second resonant mode being in the PCS communication band.
The S11 response of the 60\% parasitics in Figure 4.27 showed two resonant frequencies in close proximity to one another, making the identification of the axial loop mode problematical from analysis of the S11 alone. Scrutiny of the far field patterns of both modes made recognition of the GPS mode relatively straightforward, as demonstrated in Figure 4.28 and 4.29.

![Figure 4.28 Circular far field patterns for the first resonant mode of the air QHA with 60\% Parasitics (1530MHz)]
It can be inferred from the far field patterns in Figure 4.28 and 4.29 that the second resonant mode is the axial GPS mode of the antenna and the addition of parasitic helices have instigated deformations in the cardioid pattern as well as major frequency shifting. Inspection of the axial ratio at the bore sight of the antenna revealed a value of 0.757 and an angle of 116.53 at the first resonance mode, and a value of 1.265 and 94.53 at the second. Neither of the values recorded gave additional evidence to suggest which emanated from the circularly polarised mode, although the trend in the S11 curves and the appearance of the far field pattern would suggest the second resonance mode originated from the axial loop mode. The shape of the far field pattern of the first resonant mode, induced by the parasitics, resembled the linear polarised modes seen in chapter 2 and chapter 3 for the twisted loop structures.

The pattern deformation of the axial mode induced by the parasitics meant broadbanding by this method for GPS application was not recommended. However, the linear appearance of the parasitic mode suggested that the idea of dual-banding was not an unreasonable proposition, providing the parasitic mode lay in the correct
communication band and was sufficiently distant in resonant frequency from the GPS band (centre frequency of 1575.42MHz) that it had negligible effects on the axial mode. This assumes that the axial loop mode returns to its cardioid shape if the parasitic mode resonates sufficiently far away from the axial mode.

The far field patterns of the modes for the air QHA with the parasitics 50% of their original length in Figure 4.30 and 4.31 do provide some evidence to the assumption of the axial mode returning to its original formation when the parasitic mode is sufficiently higher or lower in frequency. The first mode of resonance for the 50% parasitic QHA is the axial mode, and as predicted its far field pattern had notably returned to its characteristic cardioid pattern. The second mode of resonance was identified as the parasitic mode and its linear far field patterns appeared similar to the bifilar far field patterns seen previously in chapter 2 and 3, complete with a null in the horizontal plane that could be directed towards the users head. It is perceived that through further research this null towards the users head could be optimised.

Figure 4.30 Circular far field patterns for the first resonant mode of the air QHA with 50% Parasitics (1570MHz)
When the two modes were scrutinised further utilising the axial ratios of the circularly polarised components, the axial mode was found to be circularly polarised exhibiting an axial ratio magnitude of 0.989 and angle of 94.51°. This demonstrated that when the parasitics were of a resonant length close to the resonance of the axial mode, the circular polarisation of the antenna is compromised.

The axial ratio magnitude and phase displayed by the parasitic mode were 0.540 and 108.30° respectively at the zenith of the far field pattern. It is clear the axial ratios of the two resonant modes for the 50% parasitics QHA are not as closely related as the two modes for the 60% parasitics QHA, and by making the parasitic and axial mode sufficiently far apart in resonant frequency allows independence in their operation in terms of far field pattern and axial ratio.

The co and cross polar far field patterns of the second resonance did provide evidence to suggest a linear mode suitable for PCS communications band (1850-1990MHz) could be developed with further research of the parasitics. This research could involve...
the number and location of parasitics to align the null of the far field pattern in the direction of the users head.

It is recognised by the author that different signal strengths would be required for GPS, which involves the reception of weak signals, and PCS or other communication bands that use comparably higher signal strength. The different signal strengths for the different operations would be controllable. It is also accepted different polarisations would be required for the different operations, but as the GPS information would be obtained as a one-off measurement and not continually calculated, the function of the antenna in this capacity is not an unreasonable proposition. However the original motivation to use the parasitic resonators as a means to increase the bandwidth of the GPS antenna was seen to be limited.

Once again with the parasitic work, only a brief study had been completed to see the effects they have on the antennas response and to see if they warranted additional and more intense analysis. For this reason no measurements had been performed for comparison with the simulated results. As yet no further investigations have been undertaken by any party to the author’s knowledge.

### 4.6 Conclusions

A quadrifilar helix antenna designed for GPS L1 band (frequency 1575.42 MHz) was developed and modelled using the Method of Moments (MoM) electromagnetic solver MiniNEC Professional for Windows. Although at this stage in the research available computational resources had improved significantly, from 330Mbytes of RAM Pentium II 350MHz PC to 512Mbytes Pentium IV 800MHz and better, a simplistic approach to modelling the QHA without its dielectric-loading was undertaken.

The air-loaded design was chosen because the additional complications of modelling an antenna with two twisted loops, that were quadrature phased to generate circular polarisation, was unknown. Furthermore the assessment of the dielectric-loading
capabilities of present day software had not been meticulously investigated. Only a limited number of simulations had been performed on a simple twisted loop bifilar structure and a variety of results had been obtained. Therefore further investigations into the software packages with dielectric-loading capabilities were required, with modelling techniques and confidence in the packages ability needing to be developed.

Past publications [16] have shown air quadrifilar helices, without an integral balun or cable feed, modelled with dual feeds at 90° to one another. This simplifies the design considerably as well as the generation of phase quadrature, as the differential in path lengths would no longer be required. Unfortunately this design would not have the advantages of isolation brought about by the novel integral balun and would require a balun trap to isolate and maintain the feed. These types of quadrifilar helices would also exhibit comparable left-hand circular components to right-hand circular polarisation.

When modelling the air QHA with integral balun using a single sourced coaxial feed with the method of MoM package (MiniNEC) it was found the balun rim had only limited influence on the inducement of phase quadrature essential for this type antenna. Even with a much increased crown height from 0.6mm to 10.0mm on the rim of the balun, phase quadrature was not realized in the currents halfway up the radiating section.

The use of turnstile and double loop antenna proved useful tools in understanding the generation of circular polarisation and the awareness of certain characteristics when examining the impedance plots, axial ratios as well as the development of the far field patterns as the path lengths of the arms or loops are modified.

When modelling the air QHA with a single feed excitation it was found that one way of generating the path differential required for phase quadrature was to meander one pair of the twisted loops in the radiating section. The procedure for generating the points for the meander-line helices have been outlined earlier in the chapter, as well as a procedure for efficiently rescaling the model to acquire circular polarisation at a particular frequency when using meander lines.
The circularly polarised far field patterns showed well defined characteristic cardioid pattern for the right hand circular plots, with a front-to-back ratio of -40dB or better, and very little left hand circular polarised patterns. When the boresight of the far field pattern was scrutinized, its axial ratio was found to be 1.01 with an angle of 92.59, which is close for optimal circular polarisation (perfect CP having axial ratio = 1.0 and angle of 90.00). The simulated right-hand circular patterns showed good agreement with measured results of the air and dielectric-loaded antennas. This was for the final design which required scaling up by a 3.24 multiplication factor from its dielectric counterpart. This factor gives a reflection of the metallisation being on the surface of the dielectric core and not being embedded into the ceramic.

By the insertion of parasitics into the quadrifilar helix structure there lays an opportunity to make the antenna dual banding (GPS/PCS) and dual polarised. This is provided the resonant mode induced by the parasitics is not too close to the existing axial mode, otherwise deformation of the axial loop mode will be observed. From the investigations performed, the introduction of parasitics to make the quadrifilar helix antennas narrow band broader would not be advisable due to the distortion of the far field pattern, the difficult controlling of the frequency shift caused by the parasitic mode and the change in axial ratio.

It is clear from the work performed in this chapter there are opportunities for developing the work with the parasitics with the air QHA, including the optimisation of the far field patterns at the requisite resonant modes, the effect of introducing more parasitics into the structure or the removal of some of the existing parasitics. Work could also be performed on the relocation of some of the parasitics to other areas on the air-loaded surface, and the use of additional parasitics to increase the efficiency in the linear polarised modes. The investigation to increase the efficiency of the linear mode by the introduction of parasitics can also be applied to the bifilar twisted loop structures in chapter 2 and 3. These are just a few proposals that could be investigated and have been inspired by the work with the parasitics helices.
4.7 References


CHAPTER 5.0

MEASURED RESULTS OF THE QUADRIFILAR HELIX ANTENNA

5.0 Introduction

In this Chapter the properties of a dielectric-loaded quadrifilar helix antenna will be compared with another antenna suited to Global Positioning System (GPS) applications. The other antenna used in this evaluation is a patch antenna that is widely used for existing GPS applications. The reason for this comparison is to justify and illustrate, using measured results, some of the properties of the dielectric loaded twisted loop topology that have been mentioned throughout this thesis.

The use of a comparative study will emphasise some of the advantages and disadvantages of using the dielectric loaded quadrifilar helix for mobile GPS applications. It will also highlight some of the issues encountered by antenna designers of GPS antennas for handheld GPS devices. Section 5.1 analyses the change in antenna response that occurs in free space when varying ground planes are used on the patch antenna and then on the quadrifilar helix. Although a ground plane is not required in the case of the quadrifilar helix, the patch antenna does require a ground plane because the patch is fed in a single-ended way and effectively resonates against the ground system.

Section 5.2 observes what happens to the response of both antennas when subjected to typical handheld positions. For this investigation the patch was mounted on a 50mm x 50mm ground plane as this was representative of the size of ground plane used for patch antennas in many GPS devices. The quadrifilar helix antenna was measured without a ground plane as it was not required for this type of antenna to function. The responses of the two antennas were compared by observing their far field radiation patterns and S11 plots.
For this chapter the main parameters used for the comparisons were zenith gain, beamwidth and front-to-back ratio. The chapter will be summarised in section 5.3 as conclusion and suggestions put forward for possible future investigations.

5.1 Free Space Measurements

The patch antenna and Dielectric Loaded Quadrifilar Helix Antenna (DLQHA) were measured in free space for their far field radiation pattern. It is widely known that patch antennas require a ground plane to operate and therefore the patch was mounted on varying size ground planes to observe any change in response. This procedure was also repeated for the quadrifilar helix and the response was analysed.

Each antenna was placed within an anechoic chamber 4.7 metres away from a right hand circularly polarised reference antenna. In all cases the boresight of the antenna under test was directed towards the reference antenna. The antennas under test were rotated 360° in steps of 5° about the elevation plane for a fixed azimuth angle for good resolution. All the far field radiation patterns were measured at the GPS frequency 1575.42MHz.

5.1.0 Patch Free Space Measurement

When the patch antenna was mounted in the anechoic chamber its ground plane was increased in size from 25mm x 25mm to 50mm x 50mm and finally to 75mm x 75mm. The S11 response of the patch antenna with varying ground planes is shown in Figure 5.1. It can be seen that as the ground plane was increased the resonant frequency of the patch shift higher in frequency and the match at resonance improves. This is characteristic of patch antennas and is widely known. In this measurement it was seen that the resonant frequency of 1575.42MHz was not achieved for any of the ground plane variants.

The main shift in resonant frequency was seen when the ground plane dimensions were changed from 25mm x 25mm to 50mm x 50mm. The shift in resonance became
less when the ground plane was increased further to the 75mm x 75mm ground plane, but fewer GPS devices have the capacity to incorporate a 75mm x 75mm ground plane into their design, and with modern handheld devices getting smaller the shift in resonant frequency of the patch antenna could become an issue.

![Diagram of S11 of GPS Patch with Varying Ground Planes](image)

**Figure 5.1 S11 of Patch Antenna with Varying Ground Planes**

When the relative far field patterns are observed for the patch antenna with varying ground planes in Figure 5.2, the importance of the ground plane to the patch antenna is observed. It can be seen the initial increase in ground plane size, from 25mm x 25mm to 50mm x 50mm, improved the zenith gain from -4.0dBi to 1.9dBi and also improved the general shape of the far field pattern. i.e. the front-to-back ratio was seen to increase from 2.0dB to 10.7dB

Ideally for GPS applications a high zenith gain is required with a broad 3dB beamwidth and high front-to-back ratio. This would ensure the majority of the far field pattern is above the horizon which is ideal for satellite reception. Therefore in Figure 5.2 the patch with a 50mm x 50mm ground plane was preferable. Due to the resonant frequency of the patch with the 75mm x 75mm ground plane being further
from the GPS frequency than the patch with the 50mm x 50mm ground plane, its far field pattern was less optimal.

![Patch at GPS Frequency](image)

**Figure 5.2 Far Field Patterns of Patch with Varying Ground Plane**

Although in Figure 5.2 the patch on a 50mm x 50mm ground plane was seen to be more optimal at the GPS frequency, it is recognised that the patch on the 75mm x 75mm ground plane does have a more optimal pattern at its resonant frequency. Therefore the patch could be tuned so the antenna on the 75mm x 75mm ground plane resonated at the GPS frequency. As patch antennas are seldom used in free space for GPS applications the main issues of concern are changes in resonant frequency of the patch antenna with varying ground plane and the characteristic pattern if the patch antenna is not optimally tuned to the GPS frequency.

### 5.1.1 DLQHA Free Space Measurement

The same procedure for the patch antenna was then applied to the dielectrically loaded quadrifilar helix antenna. As the quadrifilar helix initially has no ground plane, only a 25mm x 25mm and 50mm x 50mm ground plane was utilised in the far field measurements. From these measurements trends for the far field patterns with increasing ground planes for the quadrifilar helix could be observed.
Figure 5.3 S11 of DLQHA in Free Space with Varying Ground Plane

Figure 5.3 shows the S11 antenna response of the DLQHA with the varying ground planes. The antenna responses show negligible frequency shifting was experienced when additional ground planes, other than the coppering on the reverse side of the patch element itself, were added and then increased in size. This response comes about due to the integral balun and feed system in the DLQHA structure that gives the antenna a balanced feed and isolation. Therefore very little frequency shift was expected and seen in the antenna response when the ground planes were introduced.

The isolation created by the balanced feed system and the integral balun would also mean that the currents remain in the radiating section of the twisted loop topology, and therefore negligible change should be seen in the shape of the far field radiation plots. The right hand circular far field plots for the DLQHA with varying ground planes are shown in Figure 5.4. The characteristic cardioid pattern as predicted by Kilgus [1-3], and in other publications [4-7], was prominent and apart from an increase in zenith gain from −2.7dBi to 0.3dBi very little change was observed in the far field pattern with increasing ground plane.
Without the ground plane present a front-to-back ratio of 15.8dB and a -3dB beamwidth of 125° was observed. This was compared to the patch with a 50mm x 50mm ground plane that had a front-to-back ratio of 10.7dB and a –3dB beamwidth of 100° as these would be the relative sizes of the antennas allowed in existing GPS products.

Although the free space measurements do show the characteristic responses of the antennas measured, they do not show the antennas behaviour on an applications platform where the characteristic response of the antenna could change dramatically.

5.2 Working Environment Measurements

To measure the GPS antennas in a typical working environment the topologies of the antennas were considered. The dielectric loaded quadrifilar helix antenna was used without a ground plane and utilised in a typical handheld position for a protruding antenna. In contrast, the patch antenna was mounted in a similar fashion to current embedded antennas with comparable flat topologies. For the two handheld positions
in the measurements the relative positions of protruding antennas and patch antennas were considered in current handsets. The two positions of the antennas are presented in Figure 5.5 and will be described in more detail.

![Figure 5.5 Handheld Measurement Positions of the (a) DLQHA and the (b) Patch Antenna](image_url)

In Figure 5.5 the turntable and hand used for the measurements can be seen. The turntable was required to hold the biological hand during the measurements. To make sure the turntable had little effect on the measurements, the antennas were measured in free space and then in 'free space' in the presence of the turntable.

The hand was then introduced into the relevant handheld position. The hand produced by SARTest, is made to be representative of a normal human hand so energy absorption should be similar to real-life scenarios. The experimental set up is then rotated $360^\circ$ scanning the elevation plane for the far field measurement; this is equivalent to a $360^\circ$ rotation of the horizontal plane in the anechoic chamber. Once again the right had circular far field pattern and S11 measurements were taken.

### 5.2.0 Handheld Quadrifilar Helix Antenna

When holding a handheld phones with a protruding antenna the user normally holds the handset of the phone and not the antenna. Therefore when deciding on an appropriate position in which to take a far field measurement it was considered that the worst-case scenario for the DLQHA would be when the user held the matching
box of the antenna. This was seen in Figure 5.5(a) that showed the matching box being held by the finger and thumb of the model hand.

In reality this would portray a worse scenario than reality, as the matching box would be embedded in the phone housing, and the casing would provide a small distance a distance between the hand and the matching circuit box. The S11 results of the DLQHA scenarios can be seen in Figure 5.6. It can be seen that the neither the presence of the turntable nor the presence of a hand has much effect on the performance of the quadrifilar helix antenna with integral balun.

![Figure 5.6 S11 Proximity Effects of Hand on DLQHA](image)

The unyielding behaviour seen in the S11 curve in Figure 5.6 was also seen in the far field patterns presented in Figure 5.7. Above the horizon negligible change was seen in the far field radiation pattern. Below the horizon some changes could be seen, but in general the far field radiation pattern of the DLQHA held its characteristic profile.

In Figure 5.7 the zenith gain decreased from $-2.7\text{dBi}$ to $-3.2\text{dBi}$ by the introduction of the hand. The beamwidth also decreased from $125^0$ to $120^0$ and the front-to-back ratio decreased from $15.8\text{dB}$ to $11.1\text{dB}$. Therefore demonstrating that a protruding antenna
of this topology does not suffer greatly from close proximity effects from the phantom hand when used in a typical handheld position.

![Effects of Hand on Sarantel Antenna with No Ground Plane](image)

Figure 5.7 RHCP Far Field Pattern of Proximity Effects of a Hand on DLQHA

### 5.2.1 Handheld Patch Antenna

In Figure 5.5 it was seen that the patch antenna was measured in a different position to the quadrifilar helix antenna. Due to its flat topology the patch antenna would be embedded in the casing of the mobile handset. This would mean that the hand would be placed directly over the antenna in the position seen in Figure 5.5(b). Furthermore as most current GPS devices utilise the patch antenna on a 50mm x 50mm ground plane the measurements also utilised the same size ground plane.

When the model hand was incorporated into the measurement it was positioned 5cm away from the antenna, with only the fingers of the hand obscuring the direct line of sight of the boresight to the reference antenna. In reality with a handset the palm of the hand could be positioned over the antenna less than 1cm away making any proximity effects more intense.
Figure 5.8 S11 of Hand Proximity Effects on a Patch Antenna with 50mm x 50mm Ground Plane

Figure 5.8 shows how the response of the antenna changed when in the typical handheld position. The resonant frequency of the antenna shifted up in frequency and its match deteriorated. The change in the antennas characteristics were also seen in its far field pattern as shown in Figure 5.9. The zenith gain was seen to drop from 1.9dBi to -14.5dBi, and the shape of the far field pattern showed immense deformation that would be unsuitable for the platform the antenna was designed for.
The majority of the far field pattern shown in Figure 5.9 for the patch antenna in the applications position was pointing in a southerly direction. Unfortunately for GPS the far field radiation pattern should be directing the reception towards the satellites, i.e. above the horizon. The characteristics of both of the measured antennas before and during the handheld positions are summarised below in Table 5.1.

<table>
<thead>
<tr>
<th></th>
<th>Zenith Gain (dBi)</th>
<th>Beamwidth (Degrees)</th>
<th>Front-to-Back Ratio (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>DLQHA</td>
<td>-2.72</td>
<td>125</td>
<td>15.82</td>
</tr>
<tr>
<td>Handheld DLQHA</td>
<td>-3.24</td>
<td>120</td>
<td>11.12</td>
</tr>
<tr>
<td>Patch with 50mm x 50mm Ground Plane</td>
<td>1.91</td>
<td>100</td>
<td>10.71</td>
</tr>
<tr>
<td>Handheld Patch with 50mm x 50mm Ground Plane</td>
<td>-14.47</td>
<td>N/A</td>
<td>-0.94</td>
</tr>
</tbody>
</table>

Table 5.1 Summary of the Close Proximity Effects on DLQHA and a Patch Antenna

5.3 Conclusions

Measurements were performed on a dielectric loaded quadrifilar helix antenna with integral balun and a patch antenna to evaluate their performance in free space and in a suitable handheld position. For the measurements in the relevant handheld position a phantom hand was utilised which had the same density as a normal human hand.

When the free space measurements were executed, the resonant frequency of the patch antenna changed depending on the size of its ground plane. The greatest shift in resonant frequency for the patch antenna was when the ground plane was increased from 25mm x 25mm to 50mm x 50mm. When the trend in modern products of becoming smaller is considered this shift in resonant frequency becomes particularly relevant.

Furthermore, for the patch antenna with a 25mm x 25mm ground plane the far field pattern was least optimal for GPS reception compared to the increased ground plane measurements. Therefore if the size of GPS devices continues to decrease in future markets then the patch antennas performance would be expected to decline.
In comparison the DLQHA had negligible frequency shift when a ground plane was included in the measurement and the far field radiation pattern showed a persistent cardioid pattern. Therefore the antenna would not need redesigning for a decrease in handset size. This is characteristic of the antennas isolation properties due to its integral balun and containment of its reactive near fields due to the high dielectric core are publicised by Leisten et al [5,8]. General characteristic properties about the DLQHA have been published in the form of simulated results, by [5-6], and measured results by McEvoy et al and Leisten et al [7,9,10].

The isolation of the dielectric loaded quadrifilar helix antenna was highlighted further when mounted in the applications position, with the thumb and the finger of the hand touching the matching circuit box. It was seen that even with the inclusion of the hand, the far field radiation pattern and S11 still maintained their shape. Very negligible close proximity effects were observed in the antennas response.

The patch antenna with a 50mm x 50mm ground plane had shown a good free space far field pattern. However when mounted in an applications position with the hand 50mm away from the antenna a significant frequency shift as well as significant far field pattern deformation was observed. This showed that if a patch were introduced, as a solution for a GPS handheld device a number of issues would have to be addressed.

1. What resonant frequency should the antenna have before being mounted on the phone
2. To reduce the deformation of the far field pattern how should the patch be positioned on the handset given its size and shape
3. When the antenna is held, are different antenna designs required for different size hands due to the frequency shifting and pattern deformation?

For the dielectric loaded quadrifilar helix antenna these issues do not need to be addressed due to its predictable and repeatable response. Although it had lower zenith gain than the patch antenna in free space it did have superior beamwidth and front-to-
back ratio. In the applications positions the quadrifilar helix antenna out performed the patch on all three criteria.

Given the narrow bandwidth of the GPS frequency band, a predictable response would be required which does not alter when the GPS device is in use. Therefore making the DLQHA a more enabling technology for current GPS applications over the patch antenna in the experiments performed.

5.4 References


CHAPTER 6.0

CONCLUSIONS AND FUTURE WORK

The design and analysis of cylindrical twisted loop structures for use in mobile telecommunications was the main theme in this thesis. In particular two primary structures are used in the analysis; the basic twisted loop structure which is used for GSM, PCN and Bluetooth applications and secondly the quadrifilar helix antenna for GPS applications. Past publications by Leisten et al [1-11] have seen the two structures exploited as dielectric-loaded antenna models and their benefits over conventional antennas for mobile telecommunications highlighted.

The research performed in this thesis was to support work being carried out by Sarantel Ltd, which has begun to produce miniature dielectric-loaded handset antennas. In particular it was motivated to facilitate rapid geometry construction and simulation of the twisted loop and quadrifilar helix structures in the analysis, design and modification investigations. By building air-loaded wire geometries using a commercially available Method of Moments electromagnetic simulation package called MiniNEC, the demand for relatively rapid model construction and simulation was met.

Original work that was achieved in the duration of the research included;

1. The successful modelling of an air loaded bifilar antenna with integral balun using both MiniNEC and FDTD which agreed with measured results.
2. Comparative study of numerical electromagnetic simulation software in their ability to simulate the bifilar antenna.
4. Characteristic far field pattern analysis of the Bifilar antenna utilising small electric and magnetic dipoles.
5. Modelling of a bifilar antenna with a helical bowtie radiating section
6. A new coaxial cable based feed system that avoids the inclusion of the inner conductor, which consequently speeds up run time and avoids some of the stair-casing issues associated with the modelling of a coaxial cable.
7. Increasing the operational bandwidth of the bifilar by the insertion of parasitic resonators.
8. Modelling an air-loaded GPS quadrifilar helix with integral balun, utilising a single feed system.
9. Insertion of parasitic resonators into the air GPS quadrifilar helix structure.
10. A dual-band dual-polarised structure that radiates with circular polarisation in the GPS communications band and linear polarisation at the PCS communications band.
11. Comparison of a dielectric loaded quadrifilar helix antenna with a conventional patch antenna for application in a handheld environment.

The initial air-loaded twisted loop structure was developed from an existing dielectric-loaded antenna model using a direct scaling factor of $\sqrt{\varepsilon_r}$. Although this did not generate a simulation model that accurately replicated the dielectric-loaded antennas response, the GSM model allowed for the primary modes of operation of the twisted loop structure to be accurately electrically modelled. The radiating section of the antenna structure was constructed using 20 wires per helix to allow good visual and numerical accuracy. The integral balun sleeve was composed using convergence tested meshed structure which kept the structure simple for quick simulation time but still permitted accurate modelling. By using convergence testing for the initial design, the compromise between accuracy and high resolution modelling was demonstrated and this also highlighted the need for high resolution in the radiating structure was not necessary.

The thin film that is associated with the helical wires on the dielectric-loaded antenna structure was realised in the simulation model using a thin strip to wire approximation [12]. The complete twisted loop structure was successfully modelled with and without its fishhook feed using MiniNEC. The simulated model without the cable showed
good comparison with simulated results of the air-loaded model using an in-house finite difference time domain package. The simulated model with the cable showed good agreement with measured results. Furthermore, the inclusion of the specifically designed simulated coaxial cable into the antenna structure excited more of the antennas primary modes. This showed the importance of the coaxial cable to the antenna structure and ultimately response of the antenna.

Further analysis and modelling of the coaxial feed system showed that the coaxial cable could be simplified with the inner conductor removed. The technique involving the removal of the inner conductor meant a coarser resolution of the feed system was available, but crucially still allowed for the propagation of the single-ended and hybrid modes without loss in accuracy of the overall antenna response. Due to the substantial amount of nodes required in the construction of the coaxial feed, a coarser resolution of the feed system would reduce the overall of the simulation time significantly.

This technique could also be applied to other electromagnetic packages when modelling the twisted loop, for example; highlighted in the comparative study of the available software packages was the issue of stair-casing effects and simulation time based on the smallest cell. These effects become particularly crucial in the case of the cable feed in a dielectric-loaded model, as the smallest cell and most severe stair-casing effects would be located at the outer conductor. Application of the technique developed in chapter 2 would eliminate the need for the inner conductor and PTFE, and simplify the geometry significantly. This would resolve the stair-casing effects of the feed system and decrease simulation time, as the smallest cell would no longer be that of the outer conductor.

The use of MoM air-loaded wire geometries to simulate the twisted loop structure significantly reduced the modelling issues of those faced when modelling the dielectric-loaded model. Importantly, it allowed parametric changes to be implemented in the twisted loop model quickly with fewer modelling considerations. One such parametric study was performed on a bluetooth twisted loop topology, with
the balanced and single-ended modes were analysed for frequency shift under varying geometric changes. The frequency shifting displayed by the balanced mode for the different conditions, could be justified by the theoretical changes in current propagation of the path loop lengths. The same, however, could not be said for the single-ended mode, which implied the existing theoretical path loop propagated by the current in the single-ended mode needed reviewing.

Simulations with and without the feed cable indicated the true identity of the current path propagation of the single-ended mode, as the mode was only present when the feed cable was also in attendance. Analysis of the near fields demonstrated the actual path taken by the currents, where high field strength was seen around the coaxial cable. This showed the current loop path consisted of the current flowing down the helices, down the balun side, across the balun base to the outer conductor and then back up the outer conductor to complete its loop path. Examination of the near fields for the balanced mode showed the wire model was not an optimal design, as some currents still propagated down the balun. These currents would account for the slight resonant frequency shifting seen in the parametric studies where it was initially considered no shifting would occur.

Further analysis was performed on the balanced mode focusing on its characteristic far field radiation pattern. The modelling of a small electric and magnetic dipole provided a novel method of generating the far field pattern for the balanced mode and understanding any deformation that may be occur through modifications. The ability to generate the far field pattern using the small dipoles was confirmed using a numerical analysis approach. This technique could be applied to looking at how the dominance of the dipoles are effected in different media; i.e. bifilars with ferrite cores instead of dielectric or varying dielectric constants, the optimisation of maximising the efficiency of the antenna. Looking at the trade off between efficiency and the specific absorption rate in antennas could also be performed.

Although the twisted loop antenna has been highlighted throughout this research as having many benefits, one of the main disadvantages of the dielectric-loaded antenna was the narrow bandwidth covered at the resonant modes. Although preliminary
designs failed to have a significant impact on the available bandwidth, the addition of parasitics to the structure proved encouraging. The use of parasitics in the structure introduced supplementary resonant modes, which by varying adjusting the length of the parasitics, could be made to couple with the existing balanced mode. Consequently this would have the required effect of increasing the operational bandwidth of the antenna. Unfortunately the match of the antenna response seemed to be compromised by the coupling parasitic modes, however from a positive prospective an outlet for future research in the broadbanding or multi-banding of the twisted loop structure was initiated.

In conjunction with the latter part of the research with the twisted loop structure, research was also performed on the circularly polarised quadrifilar helix with integral balun. Once again, a wire geometry was constructed using the MoM MiniNEC electromagnetic software package but it was found that extra consideration was required for the quadrifilar helix structure. In the dielectric-loaded model a crowned balun rim was used to induce the phase quadrature that is vital for the generation of circular polarisation. Regrettably it was found a crowned rim could not induce quadrature phasing in the helical wires of the radiating section of the wire geometry generated.

Simplistic antenna configurations were utilised to ascertain antenna response trends in the generation of circular polarisation to aid in the development of the wire quadrifilar helix. A meander line approach was ultimately undertaken in one of the pairs of orthogonal loops that make up the quadrifilar helix structure. These meander lines were located in the helical wires of the radiating section and were accurately defined in Cartesian coordinates using a mathematically defined spreadsheet. The additional electrical length provided by the meanders permitted the structure to radiate with circular polarisation. The right hand circular far field patterns showed good agreement with measured results of both air and dielectric loaded models; all exhibited the characteristic cardioid patterns, as demonstrated for the quadrifilar helix by Kilgus [13-15], as well as near ideal axial ratio of unity.
As with the dielectric-loaded twisted loop antenna, the dielectric-loaded quadrifilar helix also possesses a narrow bandwidth. Parasitic helices were added to the air GPS antenna geometry in an attempt to increase the operational bandwidth of the antenna. An investigation into the addition of parasitics to the structure revealed that deformation of the characteristic cardioid pattern occurred when the parasitics resonant mode was situated in the vicinity of the existing axial loop mode.

Consequently, the quadrifilar helix antenna was investigated with the parasitics resonant mode designed around the PCS communications band. This allowed the axial loop mode to propagate with circular polarisation with negligible deformation to the volute pattern. The parasitic mode showed high co and cross polarisation similar in composition to that of the twisted loop. This means an opportunity has been demonstrated to exist for a novel dual-band dual polarised antenna structure.

The introduction of parasitics to the twisted loop and quadrifilar helix structures created many avenues for future research. The dual-band dual-polarised GPS/PCN antenna is arguably the most important novel design to surface from the research carried out for todays market. Markets for the US E-911 and the European E-112 emergency positioning location applications has meant a demand for a dual-polarised dual-band antenna has emerged. Future work can therefore be aimed at the optimisation and realisation of the dual-band design. More research could also be performed in a full analysis of the effects of the parasitics in both the bifilar and quadrifilar structures, as well as utilising the parasitics to increase the efficiency of the bifilar twisted loop antenna.

Finally, performing measurements on an applications platform highlighted the advantages of the cylindrical antenna structure used in this thesis. The relative positions of the patch antenna and quadrifilar helix structure were equated to current positions of antennas in mobile phones of embedded and protruding antennas. Little or no pattern deformation and negligible resonant frequency shifting was experienced by the quadrifilar helix antenna with integral balun when measured in the different arrangements. The patch antenna, as well as requiring a sizeable ground plane, exhibited pattern deformation and resonant frequency shifting when measured. The
measurements performed were based on theoretical positioning of the relevant antennas, however, recent publications by the author have shown a commercially available handsets with GPS capabilities to be non-optimal for satellite reception when compared to the quadrifilar helix antenna [11]. Future work is currently being undertaken in this area by the author, analysing the GPS performance and development of handheld devices for US E-911 and European E-112 applications.

6.0 References


Conclusions and Future Work


APPENDIX A

ELECTROMAGNETIC SIMULATORS THEORY

Introduction

In the duration of the research a number of electromagnetic simulator packages have been investigated and used for modelling the twisted loop antennas. The two techniques that were used in the electromagnetic solvers within the simulation packages by the author consisted of the Method of Moments (MoM) and the Finite Difference Time Domain (FDTD) techniques. These two techniques are summarised in brief in this appendix to give the reader a greater appreciation of the modelling techniques.

Method of Moments (MoM)

MiniNEC uses thin wires to build an antenna geometry which is then solved using the MoM technique which is a well-documented technique [1-4]. The method of moments process begins with a number of important assumptions which are valid for thin wires. These include that the wire radius, a, is small with respect to the wire length and also very small with respect to the wavelength. Because each wire within the antenna geometry needs to be subdivided into short segments, the radius of the wires must be small compared to the segment length. This means that the current can be assumed to be axially directed (i.e. there is no azimuthal component of the current).

The electric field for the wire geometry is formulated in terms of its vector magnetic potential and the scalar electric potential. These two potentials can be calculated from potential integrals which are solutions of the Helmholtz vector and scalar wave equations.
The integrands from the potential integrals are the wire current and the wire charge distributions. The current and charge are linked by the equation of continuity. The boundary condition that tangential electric fields at the surface of a perfect conductor is used in MiniNEC, so that the electric field must be zero. As previously said the wires are assumed to be thin therefore forcing the total axial electric field on the wire to zero. The three sources of the tangential electric field on the wire are:

- Currents and charges on the wires and on nearby wires
- Incoming waves from distance and nearby radiators
- Local sources of electric field on the wire, usually in the form of voltage sources or transmission lines that connect to the wires

By summing the tangential electric field components at each segment on the wire antenna and enforcing the zero total value an integral representation for the currents and charges is obtained.

In essence, a solution for the induced current density in the form of an integral equation where the unknown induced current density is part of the integrand. MoM is then used to solve the current density. To illustrate the method, the electrostatic charge distribution on a finite straight wire of length \( l \) and radius \( a \) is firstly considered. It is placed along the \( y \) axis as shown in figure 1.

![Figure 1 Straight Wire aligned along the Y-axis](image)

The wire has a normalised constant electric potential of 1V. A linear charge distribution \( \rho(r') \) creates an electric potential \( V(r) \) given by
\[ V(r) = \frac{1}{4\pi\varepsilon_r} \int_{\text{source}}^{} \frac{\rho(r')}{R} dl' \]

(1)

This is valid everywhere including on the wire itself. Thus choosing the observation along the wire axis \((x = z = 0)\) and representing the charge density on the surface of the wire, equation 1 can be expressed as

\[ 1 = \frac{1}{4\pi\varepsilon_0} \int_0^I \frac{\rho(y')}{R(y, y')} \, dy', (0 \leq y \leq l) \]

(2)

where

\[ R(y, y') = R(r, r') \big|_{x=z=0} = \sqrt{(y - y')^2 + [x']^2 + [z']^2} = \sqrt{(y - y')^2 + a^2} \]

R is the distance from the any one point to the observation point. The observation point is chosen along the wire axis and the charge density is represented along the surface of the wire to avoid \(R(y, y') = 0\), as this would introduce a singularity in the integrand of equation 2. Equation 2 which is an integral equation is then solved for the unknown charge density \(\rho(y')\) based on the 1V potential. The solution can be arrived at by numerically reducing equation 2 to a series of linear algebraic equations that may be solved by conventional matrix equation techniques. This is achieved by approximating the unknown charge distribution \(\rho(y')\) by an expansion of N known terms with constant but unknown coefficients, i.e.

\[ \rho(y') = \sum_{n=1}^{N} a_n g_n (y') \]

(3)

Substituting this into equation 2

\[ 4\pi\varepsilon_0 = \int_0^I \frac{1}{R(y, y')} \left[ \sum_{n=1}^{N} a_n g_n (y') \right] \, dy' \]

(4)

The wire is next divided up into N uniform segments, each with a length of \(\Delta = l/N\), as shown in figure 2. The \(g_n(y')\) functions in the expansion (3) are basis (or expansion) functions and are chosen to accurately model the unknown quantity, whilst minimising computation. To avoid complexity subdomain piecewise constant functions will be used. These functions are of constant value over one segment and zero elsewhere.
In equation 4 $y$ is replaced by a fixed point on the surface of the wire, such as $y_m$, this results in an integrand that is a function of $y'$, so the integral can be evaluated. This leads to one equation with $N$ unknowns $a_n$. To solve for these $N$ amplitude constants, $N$ linearly independent equations are required. To obtain these equations $N$ observation point’s $y_m$ are chosen on the surface of the wire at the centre of each $\Delta$ length element as demonstrated in figure 2. This results in one equation corresponding to each observation point; for $N$ such points the equation becomes

$$4\pi\varepsilon_0 = a_1 \int_0^\Delta \frac{g_1(y')}{R(y_1, y')} \, dy' + \cdots + a_N \int_{(N-1)\Delta}^\Delta \frac{g_N(y')}{R(y_N, y')} \, dy'$$

This can be written more concisely with matrix notation as

$$[V_m] = [Z_{mn}] [U_n]$$

where each $Z_{mn}$ term is equal to

$$Z_{mn} = \int_0^\Delta \frac{g_m(y')}{\sqrt{(y_m - y')^2 + a^2}} \, dy' = \int_{(n-1)\Delta}^{n\Delta} \frac{1}{\sqrt{(y_m - y')^2 + a^2}} \, dy'$$
and

\[
[I_n] = [a_n] \quad (7b)
\]

\[
[V_m] = [4\pi \varepsilon_0] \quad (7b)
\]

The \( V_m \) column matrix has all terms equal to \( 4\pi \varepsilon_0 \), and the \( I_n = a_n \) values are the unknown charge distribution coefficients. Solving equation 7 for \([I_n]\) gives

\[
[I_n] = [a_n] = [Z_{mn}]^{-1} [V_m] \quad (8)
\]

Either equation 7 or equation 8 can be solved easily using a computer by the use of matrix inversion or equation solving routines.

In summary equation 2 is an integral equation that can be used to solve for the charge distribution. This is achieved using a method which is referred to as Method of Moments. To solve equation 2 the unknown charge density \( \rho(y') \) is represented by \( N \) terms as given in equation 3. For equation 3 \( g_n(y') \) were the set of \( N \) known functions referred to as basis or expansion functions, \( a_n \) represents a set of constant, yet unknown, coefficients. These basis or expansion functions are chosen to represent the unknown charge distribution as best as possible.

Equation 2 is applicable at every point on the wire and by enforcing (2) at \( N \) discrete but different points on the wire equation 2 is reduced to a set of \( N \) linearly independent algebraic equations shown in equation 6. In equation 6 a system of \( N \) linear equations and \( N \) unknowns was derived by applying a constant 1V potential at \( N \) discrete points on the wire which is referred to as point matching. This set of equations is generalised and solved for the unknown coefficients \( a_n \) in equation 8 using inversion matrix techniques. By calculating the coefficients \( a_n \) of the \([I]\) matrix the charge distributions can be approximated using equation 3. Once the charge distribution and current distribution is known other antenna parameters such as input impedance and radiation patterns can be readily obtained.
Finite Difference Time Domain (FDTD)

The Finite Difference Time Domain (FDTD) technique, which is a well-documented technique [5-7] solves Maxwell’s curl equations in derivative form in the time Domain. These equations are expressed in a linear form by utilising central finite differencing. Nearest neighbour interactions are considered in the computation and are advanced temporally in discrete time steps over rectangular spatial cells called Yee cells [8]. These cells are what are required to build up the antenna geometry. Although these cells mean that the antenna geometries are volumetric, thin plates and thin wire can be successfully modelled.

As mentioned the FDTD technique solves Maxwell’s curl equations, in a linear medium, as in equations 9 - 12 in the time domain

\[
\nabla \times \mathbf{E} = -\frac{\partial \mathbf{B}}{\partial t} \quad (9) \\
\n\nabla \times \mathbf{H} = \frac{\partial \mathbf{D}}{\partial t} + \mathbf{J} \quad (10) \\
\n\n\nabla \cdot \mathbf{D} = \rho \quad (11) \\
\n\n\n\nabla \cdot \mathbf{B} = 0 \quad (12)
\]

Where \( \mathbf{D} = \varepsilon \mathbf{E} \) and \( \mathbf{B} = \mu \mathbf{H} \)

These equations are sufficient to represent the field behaviour over time for linear isotropic materials, providing the initial field distribution is specified and satisfies Maxwell’s equations. Often the fields and sources are initialised to zero at time zero. The two divergence equations stated above are made redundant as they are contained within the curl equations and the initial boundary conditions. Therefore the starting point for the FDTD formulation can be rewritten as
\[
\frac{\partial H}{\partial t} = -\frac{1}{\mu} (\nabla \times E) - \frac{\sigma^*}{\mu} H
\]

(13)

\[
\frac{\partial E}{\partial t} = -\frac{\sigma}{\varepsilon} E + \frac{1}{\varepsilon} (\nabla \times H)
\]

(14)

Where \( J = \sigma E \) to allow for any lossy dielectric material and \( \sigma^* \) for magnetic conductivity in the case of any magnetic loss. This formulation only looks at the electromagnetic fields \( E \) and \( H \) and not the fluxes \( D \) and \( B \). Any linear isotropic material can be specified if required as all constitutive parameters \( \varepsilon, \mu, \sigma \) and \( \sigma^* \) are present. The two equations 13 and 14 can be discretised if required to obtain total \( E \) and \( H \) field techniques by separating the fields into scattered and incident \( E \) fields.

To keep this summary of the FDTD technique simplistic a perfect conductor FDTD formulation is considered. If a perfect conductor is being used then the scattered fields must satisfy the free space conditions \( \sigma^* = \sigma = 0, \mu = \mu_0, \varepsilon = \varepsilon_0 \), therefore reducing equations 13 and 14 to

\[
\frac{\partial H}{\partial t} = -\frac{1}{\mu_0} (\nabla \times E)
\]

(15)

\[
\frac{\partial E}{\partial t} = \frac{1}{\varepsilon_0} (\nabla \times H)
\]

(16)

The derivatives within these free space field equations are then replaced with finite differencing using

\[
\frac{\partial f}{\partial t} \approx \lim_{\Delta t \to 0} \frac{f(x,t_2) - f(x,t_1)}{\Delta t}
\]

(17)

\[
\frac{\partial f}{\partial x} \approx \lim_{\Delta x \to 0} \frac{f(x,t) - f(x_1,t)}{\Delta x}
\]

(18)

The finite differencing used is finite rather than infinitesimal. A central difference scheme is utilised that only retains its first order terms. The \( E \) and \( H \) fields are interleaved spatially and temporally because of this differencing; this is illustrated later with the Yee cell which is used to build up the antenna geometry and workspace. \( \Delta t \) within the
equation is given by the Courant Stability condition, $\Delta t \leq (\Delta x)/c\sqrt{3}$ for cubical cells. By decomposing the vector Maxwell’s curl equations further into their scalar parts and by introducing the finite differencing from (18) the $E_x$ and $H_y$ components become

$$\frac{E^n_x - E^{n-1}_x}{\Delta t} = \frac{1}{\varepsilon_0} \left[ \frac{\Delta H^{n+1/2}_z}{\Delta y} - \frac{\Delta H^{n-1/2}_y}{\Delta z} \right]$$

(19)

$$\frac{H^{n+1/2}_y - H^{n-1/2}_y}{\Delta t} = \frac{1}{\mu_0} \left[ \frac{\Delta E^n_z}{\Delta x} - \frac{\Delta E^n_x}{\Delta z} \right]$$

(20)

Only two components are taken in order to show the process, all other components would follow naturally. This completes the perfect conductor separate field formulation. Six field components are used in total which all follow the same format as equation 19 and 20. These equations are then implemented into FORTRAN language. For a geometry space is quantized letting $x = l\Delta x$, $y = J\Delta y$ and $z = K\Delta z$. Similarly time is also quantized therefore $t = n\Delta t$. Uniform cells, known as ‘Yee cells’, are defined within the workspace and are located by indices I, J and K. Yee cells position the field components at offsets which results in spatially centred differencing. In Yee notation $E^\text{n}(I,J,K)$ represents the $z$ component of the electric field at time $t = n\Delta t$ at spatial location $x = I\Delta x$, $y = J\Delta y$ and $z = (K+1/2)\Delta z$. This is demonstrated in the Yee cell in figure 3.

![Figure 3 Field Evaluation Points on a Yee Cell for (a) Electric Fields (b) Magnetic Fields](image)
By using this Yee cells and the equations 19 and 20, the E field can be calculated at a time corresponding to \( n = N \) from its prior value at time \( n = N-1 \) and the curl of H at time \( n = N-1/2 \), where \( t = \Delta t \). H is then evaluated at \( n = N+1/2 \) from its earlier value at \( n = N-1/2 \) and the curl of E at \( n = N \). This gives a centred difference or ‘leapfrog in time’ approach for the E and H fields components. This process is continually repeated; with each time the index \( N \) updated by 1 until a given number of time steps is reached.

For lossy material and lossy dielectric slightly more complex equations to those in 19 and 20 are derived, however the basic principal of temporal and spatial stepping of the E and H fields remains the same. Other considerations for the FDTD implementation are

- Yee cell size
- Time step size
- Absorbing Boundary Conditions (ABC’s)

Each of these factors has an important role in the accuracy of the results that are obtained. The cell size must be small enough to permit accurate results at the highest frequency of interest but large enough to keep resources manageable. It is directly affected by the materials present; for example if dielectric is being used then the cell size would need to be smaller. The greater the permittivity or conductivity, the shorter the wavelength at a set frequency and therefore a small cell is clearly required.

Once the cell size is selected, the time step can be determined using the Courant stability condition. If time steps are too large then instability becomes an issue. Absorbing Boundary Conditions are required as no computer can store an unlimited amount of data and therefore the workspace over which the geometry is simulated must be limited in size. The computation domain must be big enough to enclose the geometry and a suitable boundary on the outer perimeter is used to simulate the domains extension to infinity. This avoids reflected waves that would undoubtedly have detrimental effects on the accuracy of the final results.
References


APPENDIX B

SIMULATED AND MEASURED FAR FIELDS OF DIPOLES AND TWISTED LOOPS

Introduction

The following plots show the far field radiation patterns for different configurations of small electric and magnetic dipoles. These were simulated using MiniNEC to try and reproduce the far field patterns that are characteristic of the bifilar antenna operating in the balanced mode. Before presenting the different electric and magnetic dipole configurations the required characteristic balanced mode pattern is presented.

The topology is then simplified into its twisted loop structure; this represents the path the currents would propagate around on the antenna structure if the balun on the bifilar was functioning optimally in the balanced mode. Next the small dipoles were simulated using different configurations including the electric and magnetic dipoles simulated on their own in free space, then together, and then finally with one of the dipoles dominating.

The final plot is a measured dielectric loaded PCN antenna which had only a 0.43 helical turn in its radiating section. By following the plots it can be seen how the dominance of the magnetic or electric dipole dominates the electromagnetic dipole far field pattern of the bifilar depending on the helical turn and media used in the radiating section.
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Simulated and Measured Dipole Fields

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Figure 2 Simulated Twisted Loop Antenna
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Simulated and Measured Dipole Fields

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Figure 4 Electric Dipole aligned along the X-axis
Simulated and Measured Dipole Fields

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Figure 5 Electric and Magnetic Dipoles Independently Fed

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