Shape Optimization of Low-Profile UWB Body-of-Revolution Monopole Antennas

Dissertation

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Abstract

There is great interest in developing low-profile ultra-wideband (UWB) antennas that operate from low VHF up to several GHz to support high bandwidth communication systems. A concurrent need exists for small size antennas that generate omnidirectional vertically polarized radiation over the horizon. This continued demand for UWB and small size antennas implies significant challenges in radio frequency aperture design. Not surprisingly, this topic has been of strong interest in the research community over the last decade.

This dissertation focuses on the development of novel approaches to systematically design and optimize the shape of a body-of-revolution (BoR) monopole antenna subject to radiation objectives and size constraints. To this end, two types of antennas are proposed following different design philosophies. Specifically, a frequency-scaled inverted-hat antenna (IHA) is first proposed with its outer surface composed of multiple ellipses that follow the growth rate of an exponential spiral. It is demonstrated that controllable input impedance and uniform radiation patterns can be maintained by this design. For the latter property, increasing diffraction from the elliptical curvatures is exploited to achieve a broader radiation pattern at higher frequencies.

A second UWB monopole is proposed by instead following a design optimization approach to construct the outer surface of the antenna. The employed techniques include a random walk for antenna outer profile generation, a genetic algorithm for
optimization, and a BoR moment method analysis—all combined in a single design package. The weighted global criterion method is adopted to construct the multi-objective cost function. The minimization of this cost function amounts to minimizing the difference between a potential optimal point and a utopia point that combines separate optimal designs for all key objectives within the solution space. Example designs are presented to demonstrate performance improvements due to antenna shape optimization. In comparison to the original IHAs of the same size, the optimized designs are shown to deliver larger bandwidth, higher gain at low VHF band, and more stable pattern at higher frequencies. Importantly, except for the cost function construction, the design optimization approach requires little prior knowledge in the theory of antennas, making it well suited for exploring a variety of unknown topics in the future. In addition, the matching condition and pattern stability of the optimized monopole can be further improved by means of resistive loading.

Several prototypes of different sizes were fabricated using standard spinning technology and measured from low VHF band up to 2 GHz. These wideband apertures demonstrate the efficacy and accuracy of the novel and low-profile UWB designs.
To my parents
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# Table of Contents

<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>Abstract</td>
<td>ii</td>
</tr>
<tr>
<td>Dedication</td>
<td>iv</td>
</tr>
<tr>
<td>Acknowledgments</td>
<td>v</td>
</tr>
<tr>
<td>Vita</td>
<td>vi</td>
</tr>
<tr>
<td>List of Tables</td>
<td>xii</td>
</tr>
<tr>
<td>List of Figures</td>
<td>xiii</td>
</tr>
<tr>
<td>1. Introduction</td>
<td>1</td>
</tr>
<tr>
<td>1.1 Motivation and Objective</td>
<td>1</td>
</tr>
<tr>
<td>1.2 Dissertation Organization</td>
<td>7</td>
</tr>
<tr>
<td>2. Survey of VHF/UHF UWB Antennas</td>
<td>9</td>
</tr>
<tr>
<td>2.1 Introduction</td>
<td>9</td>
</tr>
<tr>
<td>2.2 UWB Monopole/Dipole Antennas</td>
<td>10</td>
</tr>
<tr>
<td>2.3 UWB Spiral Antennas</td>
<td>13</td>
</tr>
<tr>
<td>2.4 Summary</td>
<td>15</td>
</tr>
<tr>
<td>3. Frequency-Scaled UWB Inverted-Hat Antenna</td>
<td>16</td>
</tr>
<tr>
<td>3.1 Introduction</td>
<td>16</td>
</tr>
<tr>
<td>3.2 Inverted-Hat Antenna Operating Principle</td>
<td>18</td>
</tr>
<tr>
<td>3.3 Initial Shaping</td>
<td>19</td>
</tr>
<tr>
<td>3.4 Multi-ellipse Design for Frequency-Independent Operation</td>
<td>26</td>
</tr>
<tr>
<td>3.4.1 Antenna Profile Description</td>
<td>27</td>
</tr>
</tbody>
</table>
List of Tables

<table>
<thead>
<tr>
<th>Table</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>4.1</td>
<td>49</td>
</tr>
<tr>
<td>4.2</td>
<td>52</td>
</tr>
<tr>
<td>5.1</td>
<td>68</td>
</tr>
<tr>
<td>5.2</td>
<td>72</td>
</tr>
<tr>
<td>5.3</td>
<td>73</td>
</tr>
</tbody>
</table>

4.1 Computational statistics of the 24” wide and 20” tall BoR dome-dipole antenna in Figure 4.4 at 6 GHz using FEKO and BoR analysis.

4.2 Codification of a 6” tall IHA parameters for GA operation.

5.1 Utopia point: minimum outputs for the objective function $F_1$, $F_2$, and $F_3$.

5.2 Objective function value $F^p_i$ of the final optimized design and the associated fraction deviation ($F^p_{i,\text{trans}} - 1$) from the utopia point.

5.3 Coordinates of the control vertices defining the final optimal profile.
## List of Figures

<table>
<thead>
<tr>
<th>Figure</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.1 Configuration of a conventional conical antenna in the form of body-of-revolution (BoR): (a) cross-sectional diagram and (b) 3-D model.</td>
<td>2</td>
</tr>
<tr>
<td>1.2 The lowest achievable cutoff frequency for minimum $Q$ antennas of different sizes in terms of realized gain.</td>
<td>3</td>
</tr>
<tr>
<td>1.3 Irregular $E$-plane radiation pattern of a $12\lambda$ wide and $3\lambda$ tall conical antenna on an infinite ground plane.</td>
<td>4</td>
</tr>
<tr>
<td>1.4 Compromise design vs. optimal shapes associated with all three key objectives.</td>
<td>6</td>
</tr>
<tr>
<td>3.1 Configuration of the UWB inverted-hat antenna (IHA): (a) cross-sectional diagram and (b) 3-D model.</td>
<td>18</td>
</tr>
<tr>
<td>3.2 Geometry of the $15'' \times 6''$ (a) double-ellipse and (b) triple-ellipse IHA.</td>
<td>20</td>
</tr>
<tr>
<td>3.3 Fabricated $15'' \times 6''$ double-ellipse IHA prototype: (a) side view and (b) antenna with a $21''$ diameter ground plane and radome.</td>
<td>20</td>
</tr>
<tr>
<td>3.4 (a) Outdoor and (b) indoor measurement setup for the fabricated $15'' \times 6''$ double-ellipse IHA prototype in Figure 3.3.</td>
<td>21</td>
</tr>
<tr>
<td>3.5 Measured and simulated horizontal realized gain of the $15'' \times 6''$ IHAs in Figure 3.2.</td>
<td>22</td>
</tr>
<tr>
<td>3.6 Simulated 3-D gain patterns of the $15'' \times 6''$ double-ellipse IHA: (a) $f = 50$ MHz,(b) $f = 500$ MHz,(c) $f = 1500$ MHz, and (d) $f = 2000$ MHz.</td>
<td>23</td>
</tr>
<tr>
<td>3.7 Impedance comparison of the $15'' \times 6''$ double-ellipse and triple-ellipse IHA (on an infinite ground plane): (a) real part and (b) imaginary part.</td>
<td>24</td>
</tr>
</tbody>
</table>
3.8 Calculated reflection coefficients of the 15” × 6” IHAs in Figure 3.2.  

3.9 Design curves relating aperture size, height and frequency in achieving a pre-specified gain of -10 dBi, -15 dBi and -20 dBi for the double-ellipse IHA (infinite ground plane): (a) antenna aperture size fixed at 15” and (b) antenna height fixed at 6”.  

3.10 (a) Spiral pattern for the determination of the major radius of each elliptical segment. (b) Configuration (side view) of a frequency-scaled triple-ellipse inverted-hat antenna (IHA) having an aperture width of \( w \) and height of \( h \); for each elliptical segment, \( x_i \) is the major radius and \( y_i \) is the minor radius. (c) 3-D view of a frequency-scaled triple-ellipse linear IHA.  

3.11 (a) 15” wide and 6” tall IHA geometry for several multi-ellipse outer surfaces, (b) realized gain in the horizontal plane (\( \theta = 90^\circ \)), (c) input resistance, and (d) input reactance.  

3.12 Impedance calculation for different outer surface profiles of the eleven-ellipse design using 15” wide and 6” tall IHA: (a) geometry, (b) resistance, and (c) reactance.  

3.13 Impedance performance for two IHA designs vs. a conical antenna all being 24” wide and 6” tall: (a) geometry, (b) resistance, and (c) reactance.  

3.14 \( E \)-plane radiation pattern comparison between a 15” × 6” conical antenna and the 11-ellipse convex IHA in Figure 3.12 at 2 GHz.  

3.15 Measurements vs. calculation for a modified seven-ellipse IHA: (a) geometry, (b) fabricated prototype with radome, (c) reflection coefficient, and (d) realized gain in the horizontal plane (\( \theta = 90^\circ \)).  

3.16 Measured and calculated normalized gain patterns for a modified seven-ellipse IHA (in the elevation plane).  

3.17 A 15” × 6” double-ellipse IHA was mounted on top of a ground vehicle to test its radiation properties.
4.1 Configuration of an UWB inverted-hat antenna: (a) cross-sectional diagram and (b) photograph of a fabricated 24” wide and 6” tall prototype.

4.2 Discretization and triangular basis functions for a body-of-revolution (BoR) dipole antenna analysis.

4.3 (a) Triangular function $T_n(t)$ with four samples. (b) Derivative of triangular function $\dot{T}_n(t)$ with four samples.

4.4 Fabricated 24” wide and 20” tall BoR dome-dipole antenna.

4.5 Simulated and measured VSWR for the 24” wide and 20” tall BoR dome-dipole antenna in Figure 4.4.

4.6 Simulated and measured realized gain in the elevation plane ($\theta = 90^\circ$) for the 24” wide and 20” tall BoR dome-dipole antenna in Figure 4.4.

4.7 $E$-plane radiation patterns for the 24” wide and 20” tall BoR dome-dipole using BoR analysis: (a) $f = 100$ MHz, (b) $f = 2$ GHz, (c) $f = 4$ GHz, (d) $f = 6$ GHz, (e) $f = 8$ GHz, and (f) $f = 10$ GHz.

4.8 Profile (side view) of the GA optimized and non-optimized 6” tall IHA.

4.9 (a) Directivity (at $\theta = 90^\circ$) and (b) impedance of the optimized and non-optimized 6” tall IHA from 1 GHz to 6 GHz.

4.10 Radiation patterns ($E$-plane) of the (a) optimized and (b) non-optimized 6” tall IHA at 1 GHz, 3 GHz, and 6 GHz.

5.1 The proposed RGB method combining a random-walk (RW) scheme, a genetic algorithm (GA), and a body-of-revolution (BoR) solver.

5.2 A flowchart of the optimization procedure.

5.3 Generation of an antenna 2-D profile via random walk.

5.4 Coordinate statistics from a Monte Carlo simulation: antenna 2-D profiles with 5 control vertices are generated 50,000 times using the RW ($x_i, y_i$) and the RNG ($p_i, q_i$) scheme, respectively.
5.5 Illustration of the weighted global criterion method. The goal is to minimize the distance \( N(x) \) between the solution point \( F(x) \) and the utopia point \( F^* \).

5.6 Side view of the optimum monopole antenna profiles (24” wide and 6” tall) associated with objective function \( F_1, F_2, \) and \( F_3 \), respectively.

5.7 (a) Side view and (b) 3-D view of the final optimized 24” wide and 6” tall BoR monopole antenna.

5.8 Horizontal realized gain (\( \theta = 90^\circ \)) of the 24” \( \times \) 6” final design via shape optimization (inset figure displays the low frequency performance).

5.9 Input impedance of the 24” \( \times \) 6” final design via shape optimization on the Smith Chart (normalized to 50 \( \Omega \)).

5.10 Low frequency gain comparison between the final design and optimum antenna #1 (inset figure displays the associated antenna configurations).

5.11 VSWR comparison between the final design and optimum antenna #2 (inset figure displays the associated antenna configurations).

5.12 Directivity (\( \theta = 90^\circ \)) comparison between the final design and optimum antenna #3 (inset figure displays the associated antenna configurations).

5.13 Costs associated with the best individual of each generation through the GA optimization.

5.14 Performance comparison of the 24” wide and 6” tall shape-optimized antenna vs. 11-ellipse IHA: (a) VSWR and (b) realized gain at horizon (\( \theta = 90^\circ \)).

5.15 The onset frequency point at -15 dBi gain of a 6” tall body-of-revolution monopole antenna over infinite ground plane for different aperture widths.

5.16 The onset frequency point at -15 dBi gain of a 24” wide body-of-revolution monopole antenna over infinite ground plane for different heights.
5.17 Side view of (a) 12" wide and 6" tall and (b) 24" wide and 2" tall BoR monopole antenna achieving lowest frequency at -15 dBi realized gain via shape optimization. ................................................................. 78

5.18 Side view of the antennas considered for a resistive loading treatment: (a) the shape-optimized antenna found in section 4.4.1, (b) a smaller PEC aperture, (c) a smaller PEC aperture incorporated with an R-card having a sheet impedance of 100 Ω/□, and (d) a smaller PEC aperture incorporated with an R-card having a sheet impedance of 500 Ω/□. .......................................................... 80

5.19 (a) Reflection coefficient and (b) realized gain in the horizontal plane (θ = 90°) of the untreated and resistively-loaded antennas (30 MHz–200 MHz). ................................................................. 81

5.20 (a) Reflection coefficient and (b) realized gain in the horizontal plane (θ = 90°) of the untreated and resistively-loaded antennas (30 MHz–2 GHz). ................................................................. 82

5.21 Radiation patterns (E-plane) of the untreated and resistively-loaded antennas: (a) 500 MHz, (b) 1200 MHz, (c) 1500 MHz, and (d) 2000 MHz. ................................................................. 83

5.22 A fabricated 24" wide and 6" tall (a) aluminum and (b) resistively-loaded (120 Ω/□) prototype assembled with a 26" diameter ground plane. ................................................................. 84

5.23 (a) Outdoor and (b) indoor measurement setup for the 24" wide and 6" tall aluminum and resistively-loaded (120 Ω/□) prototypes in Figure 5.22. 85

5.24 Measured and simulated (a) VSWR and (b) realized gain at horizon (θ = 90°) of the 24" wide and 6" tall aluminum and resistively-loaded prototype shown in Figure 5.22. ................................................................. 86

5.25 Measured and simulated radiation patterns (E-plane) of the 24" wide and 6" tall aluminum and resistively-loaded prototype shown in Figure 5.22: (a) 1000 MHz, (b) 1500 MHz, (c) 1800 MHz, and (d) 2000 MHz. ................................................................. 87
6.1 (a) 15” wide and 6” tall double-ellipse inverted-hat antenna (IHA), (b) 12” spiral antenna, and (c) the hybrid design placed on a 22” ground plane. ................................................................. 95

6.2 Preliminary gain measurement results of the monopole-spiral hybrid design in Figure 6.1: (a) comparison of the measured horizontal gain (with and without spiral on top of the double-ellipse IHA) and (b) boresight gain of the spiral (IHA was terminated). ........................................ 95

A.1 Program structure of the Body-of-Revolution Monopole Antenna Shape Optimization Code (BoRMASOC). ............................................ 112
Chapter 1: Introduction

1.1 Motivation and Objective

There is great interest in developing low-profile ultra-wideband (UWB) antennas that operate from low VHF up to several GHz to support high bandwidth communication systems. For instance, a challenge for platform such as unmanned aerial vehicles (UAV) is the integration of small UWB antennas that cover the frequency span from 2 to 2000 MHz [1]. This 1000:1 bandwidth serves a variety of communication needs, including HF voice (2–30 MHz), SINGARS (30–88 MHz), SATCOM (240–320 MHz), EPLRS (450–470 MHz), JTRS GMR (up to 2 GHz), to mention a few. The conventional approach used to cover such a large bandwidth is to employ a separate antenna for each system. However, it gives rise to problems (e.g. space, weight, cost and electromagnetic compatibility/interference, etc.) when multiple communication systems are required at the same time. Software defined radio (SDR) is a promising technology to overcome this difficulty. Specifically, using the large operating bandwidth along with tunable receiver hardwares, a single broadband radiator can support different functions with adaptable data rates. Nevertheless, in designing such antennas for airframe or ground vehicles, aperture size and profile should be reduced in size by a factor of 10 or more from currently available wavelength size antennas (e.g. 5 m or ~15 ft at 30 MHz for a typical λ/2 dipole). Unfortunately, bandwidths of
electrically-small antennas decrease inversely with the third power of their electrical length \( ka \) [2]. Thus, there is a concurrent need for large bandwidth and small size antennas, implying significant challenges in radio frequency (RF) aperture design.

Being low-profile, spiral antennas have been extensively studied as frequency-independent radiators [3–6]. That is, their impedance, gain, and pattern vary little with frequency, making it an ideal candidate for replacing a number of antennas on small platforms. However, spirals radiate a patch-like pattern, not desirable for over the horizon low frequency operation. Body-of-revolution (BoR) monopole antennas generate omnidirectional vertically polarized (VP) radiation in the horizontal plane, which is critical for communication systems using SINGARS and UHF frequencies. Among this class of axisymmetric structures, the volcano smoke [7] and conical antenna [8, 9] (see Figure 1.1) are well-known broadband radiators sharing the feature of tapering their surface away from the feed to maintain nearly constant impedance over a large bandwidth. However, they are protruding and non-conformal for low frequency operation. Besides, their highest operational frequency is limited by the radiation patterns. That is, as the frequency increases, side lobes become dominant.

![Figure 1.1](image)

Figure 1.1: Configuration of a conventional conical antenna in the form of body-of-revolution (BoR): (a) cross-sectional diagram and (b) 3-D model.
due to the increased electrical separation between the aperture and ground plane. Irregular patterns in the frequency domain also indicate a signal distortion in the time domain for an UWB pulse.

Miniaturization, broadband matching, and pattern uniformity are therefore three key design objectives for a BoR monopole antenna considered in this dissertation. But, these objectives are often in conflict with each other and trade-offs have to be made to find out a compromise design.

Size reduction of an antenna often leads to the issue of deteriorating performance such as lower gain and narrower bandwidth. Electrically small UWB antennas typically have a high-pass response. In [10], a theoretical limit was established for the cutoff frequency (cutoff electrical size) subject to a maximum tolerable reflection coefficient using the fundamental limit on the radiation $Q$ [11] in conjunction with the

![Figure 1.2: The lowest achievable cutoff frequency for minimum $Q$ antennas of different sizes in terms of realized gain.](image)
Fano-Bode matching theory [12]. This relation was then combined with the directivity limitations [13] to define a limit on the maximum achievable realized gain as a function of antenna cutoff size. Figure 1.2 displays this limit for two BoR monopole antennas (on an infinite ground plane) subject to size constraints. These antennas are $w$ inches in width and $h$ inches in height. Clearly, the goal of miniaturization is to achieve higher gain at low frequencies as the antenna is electrically small without increasing its physical size. It is important to point out that the aforementioned limit is for a minimum $Q$ antenna of which the response is high-pass. Higher gain at a given frequency can be reached at the price of reduced bandwidth. The conical configuration exhibits broadband characteristics as its input impedance is primarily determined by the cone angle when the radiator becomes sufficiently electrically large [8]. However, the radiation resistance is typically very low for low-profile designs of which the footprint (width) to height ratio is high. Caution must be exercised.

![Figure 1.3: Irregular E-plane radiation pattern of a 12\(\lambda\) wide and 3\(\lambda\) tall conical antenna on an infinite ground plane.](image)

Figure 1.3: Irregular E-plane radiation pattern of a 12\(\lambda\) wide and 3\(\lambda\) tall conical antenna on an infinite ground plane.
when shaping the antenna outer surface from the feed section to maintain a wide-band matching. Additionally, pattern distortion at high frequencies severely limits the application of conical antenna of which the bandwidth might be in excess of 10:1. Figure 1.3 indicates a significant side lobe issue for a $12\lambda$ wide and $3\lambda$ tall conical antenna, implying plenty room to improve upon its pattern uniformity.

Given the aforementioned challenges, this dissertation focuses on the development of novel approaches to systematically design and optimize the shape of a BoR monopole antenna subject to radiation objectives and size constraints. To this end, two types of antennas are proposed following different design philosophies.

Referring to [14], an UWB antenna should be formed by assembling structures with similar geometry scaled with distance from a reference feed point. It is therefore favorable to create a series of “frequency-scaled” scattering structure, which maintain an identical equivalent radiation aperture at different frequencies. Inspired by the spiral antenna operating mechanism, a frequency-scaled inverted-hat antenna (IHA) is first proposed with its outer surface composed of multiple ellipses that follow the growth rate of an exponential spiral. The curved surface profile reduces field variations as compared to the linear profile, resulting in stronger radiation at low frequencies. It is also demonstrated that controllable input impedance and uniform radiation patterns can be maintained by this design. For the latter property, increasing diffraction from the elliptical curvatures is exploited to achieve a broader radiation pattern at higher frequencies.

A second UWB monopole is proposed by instead following a design optimization approach to construct the outer surface of the antenna. The employed techniques include a random walk for antenna outer profile generation, a genetic algorithm for
optimization, and a BoR moment method analysis—all combined in a single design package. The weighted global criterion method is adopted to construct the multi-objective cost function. The minimization of this cost function is essentially minimizing the difference between a compromise solution point and a utopia point within the criterion space. The utopia point combines separate optimum designs for all three key objectives, as shown in Figure 1.4. Examples are included to demonstrate performance improvements due to antenna shape optimization. In comparison to the original IHAs of the same size, the optimized designs are shown to deliver larger bandwidth, higher gain at low VHF band, and more stable pattern at higher frequencies. Importantly, except for the cost function construction, the design optimization approach requires little prior knowledge in the theory of antennas, making it well suited for exploring a variety of unknown topics in the future. In addition, the matching condition and pattern stability of the optimized monopole can be further improved by means of resistive loading.

The key contributions of this dissertation are summarized below.
1. Invention of the frequency-scaled inverted-hat antenna (IHA) that maintains a similar scattering geometry in terms of wavelength, providing controllable input impedance and uniform radiation patterns.

2. Development of a novel multi-objective antenna shape optimization method amounted to minimizing the difference between an achievable design point and a utopia point that combines separate optimal designs for all key objectives within the solution space.

3. Design, fabrication and measurement of several low-profile BoR monopole antennas operating from low VHF up to 2 GHz. These compact wideband structures can be readily applied across a variety of airborne and vehicular platforms.

1.2 Dissertation Organization

This dissertation is organized as follows.

In Chapter 2, a survey is conducted on UWB antennas that operate at VHF and UHF band. Based on their radiation patterns, these wideband radiators are categorized into monopole/dipole antennas and spiral antennas. Emphasis is particularly placed on the broadbanding and miniaturization techniques.

In Chapter 3, a novel low-profile inverted-hat antenna (IHA) that can operate not only down to low VHF frequencies but also retain good performances at higher frequencies (up to 2 GHz) is proposed. A systematic approach to design the outer surface of the IHA subject to aperture size and height is developed. The resulting multi-ellipse IHAs achieve controllable input impedance and uniform radiation pattern by maintaining a similar scattering geometry in terms of wavelength. Several prototypes are fabricated and measured to verify the design concept.
To facilitate the shape optimization of IHAs, in Chapter 4, a body-of-revolution method of moment (BoR-MoM) analysis is carried out, along with presenting a validation via the measurement of a dome-dipole antenna over a 100:1 bandwidth. The reasonably good agreement between calculations and measurements demonstrates the feasibility of integrating this BoR-MoM analysis with a formal optimization technique to design the outer surface of low-profile UWB BoR monopole antennas.

Chapter 5 starts with the shape optimization methodology, which consists of a random walk profile representation, a genetic algorithm, and a BoR-MoM analysis. The cost function construction strategy based upon the weighted global criterion method is discussed next. Two design examples are provided further to illustrate the optimization procedure as well as to demonstrate its effectiveness in comparison to the original IHAs. A resistive loading approach is also considered as the post-optimization treatment to improve antenna matching and pattern stability concurrently. In the end, measurements are presented to validate the design efforts made on a 24” wide and 6” tall aperture operating from 40 MHz up to 2 GHz.

This dissertation concludes with a summary of major contributions. Future work include extending the scope of application of the existing shape optimization method and combining the wideband monopole with a planar spiral antenna for pattern diversity.
Chapter 2: Survey of VHF/UHF UWB Antennas

2.1 Introduction

UWB antennas are more than 100 years old [15]. In 2002, the Federal Communications Commission (FCC) issued a ruling that allowed intentional UWB emission in the frequency range between 3.1 GHz and 10.6 GHz, which greatly drew the public’s attention to UWB technology—particularly for commercial applications [16]. By definition, an UWB antenna operates over a fractional bandwidth that is greater than 0.2, or occupies 500 MHz or more of the frequency spectrum, regardless of the fractional bandwidth [17]. It is worth mentioning that, in this dissertation, the continuous frequency coverage is not limited to the FCC ruling. Instead, the primary band of interest is from 30 MHz to 2 GHz—in order to accommodate the needs of future military communication systems such as joint tactical radio systems (JTRS) [1] and future combat system (FCS) [18]. It is natural that consideration be given to conformal UWB antennas since radiators are required to be placed as close to a surface (e.g. a ground plane) because of many practical purposes. The objective of this chapter is to provide an insight into the state-of-the-art designs in this area. There are many ways to classify UWB antennas. For instance, they can be categorized as directional or non-directional; they can also be classified as either electric or magnetic antennas. In the remainder of this chapter, UWB antennas are arranged
in classes according to their radiation pattern shapes: monopole/dipole or spiral. A monopole/dipole antenna radiates vertically polarized (VP) waves over the horizon. In contrast, a spiral antenna is circularly polarized (CP), offering the maximum radiation toward the zenith—a patch-like pattern. Some of the conventional wideband radiators, such as horns and tapered slot antennas (TSA), fall in neither of the two categories, and they will not be considered here. However, this classification covers most low-profile UWB antennas that have been extensively studied over the years for VHF/UHF communication systems. More importantly, it reasonably relates radiation patterns to design techniques thanks to the distinct radiation mechanisms associated with these two types of broadband antennas.

2.2 UWB Monopole/Dipole Antennas

UWB antennas actually date back to the “spark gap days”. In the great invention of electric telegraphy in 1898 [19], Oliver Lodge claimed his device was able to “transmit messages across space to any one or more of a number of different individuals in various localities . . . ” One of his drawings showed that biconical antennas were used in a transmit-receive link. In the late 1930s, Carter revisited Lodge’s design and achieved a broadband impedance matching by connecting the radiating element and the feed line with a tapered section [20]. This work aimed to provide a wideband structure for television transmission or reception. Within the same time span, Lindenblad designed “a practical radiator having a substantially flat characteristic over a wide range of frequencies making it thereby especially suited for television transmission.” This coaxial horn element [21] was similar to the well-known “volcano smoke
antenna” developed by Kraus [7]. Another great invention brought by Brillouin represented not only the state-of-the-art design but also profound UWB antenna concept at that time [22]. He claimed his invention was “... for ultra high frequencies, radiates vertically polarized waves only, has a terminal impedance close to the impedance of free space, maintains constant directivity over a broad band of frequency, and has substantially no upper limiting frequency ...” This might be the first time in history that a frequency-independent antenna offering a monopole/dipole-like pattern was proposed. Further advances include Master’s diamond dipole [23], Stohr’s ellipsoidal monopole/dipole [24], Harmuth’s large current radiator [25], to name a few.

Among numerous wideband antennas possessing different shapes, the simplest configuration is the biconical antenna, which is formed by placing two infinite cones together. Its radiation characteristic is determined by the flare angle only; in other words, it is frequency-independent. However, in practice, the lengths of the cones must be finite. The performance degradation stems from the formation of standing waves due to the discontinuity of the structure. For practical purpose, a monopole version—the conical antenna—has been widely adopted since the antenna height is reduced by one half comparing to the biconical configuration. Papas and King calculated the input impedance along with radiation patterns of spherically capped conical antennas fed by a coaxial line [8, 9], which pioneered the research in this area. Brown and Woodward experimentally examined the radiation characteristics of conical and triangular antennas [26]. Adachi et al. provided a theoretical and experimental study of a finite conical antenna excited at its tip by a linear element [27]. Time-domain characteristics of a conical antenna excited by a dc pulse were investigated by Harrison and Williams [28]. Pridmore-Brown employed diffraction theory to predict the
radiation fields generated by a slot-excited conical antenna [29]. To reduce the weight and wind resistance, biconical antennas consisting of wire bow-ties were examined by several scholars [30, 31]. It has been shown that a BoR antenna can be reasonably approximated by wire-constructed structures with eight or more elements. The V-conical antenna was proposed in [32] to improve the energy efficiency toward a particular direction by incorporating some directionality in the simulator design. A conical antenna with a section of continuous resistive loading was presented in [33] to resolve the issue of pulse distortion on radiation. As the antenna size becomes a major concern in real-world applications, effort has been made to miniaturize the conical antenna in recent years. Among them, capacitive loading by introducing a top plate and adding parasitic shorted conducting pings are two most common approaches [34, 35]. Besides, it was reported in [36] that the height of a conventional biconical antenna was reduced by over 60% while maintaining electrical performances, as a result of adding several geometric features and applying dielectric loadings properly.

A number of other low-profile configurations that differ from a cone shape but deliver UWB performance are available in the literature as well. Goubau antenna, which consists of four electrically small vertical conductors with each one terminated in a conductive plate, significantly broadens the bandwidth of traditional top loaded monopoles [37]. It was reported in [38] that a mode-0 spiral-mode microstrip (SMM) antenna has a 10:1 gain bandwidth but only one half of the height of a conical monopole. This low-profile conformal design is substantially a traveling wave antenna. A broadband roll monopole, which is compact compared to the conical antenna and symmetric compared to the planar monopole, was presented in [39] for the operation from 1.2 GHz to 2.2 GHz. Using the genetic algorithm (GA), Altshuler and Linden
developed an UWB impedance-loaded monopole having a 50:1 impedance bandwidth (VSWR < 4.5) and a near hemispherical coverage while maintaining the size less than 0.1\( \lambda \) at the lowest frequency [40]. Broadband planar monopoles were also considered in recent years for portable wireless devices where low-profile radiators are typically required. Among them, circular and elliptical discs [41], inverted-cones [42], staircase printed monopoles [43], and shape-optimized flare dipoles [44] were shown to achieve large impedance bandwidth and stable patterns in the vertical plane. But, their patterns are not omindirectional in the horizontal plane. This drawback sets off the advantage of volume reduction, especially at high frequencies.

2.3 UWB Spiral Antennas

Spiral antennas deliver constant impedance, almost perfect circularly-polarized (CP) radiation, and patch-like patterns over a wide range of frequencies. Since they were first introduced in the 1950s by Edwin Turner, many variations have been considered [3–6]. In this survey, a focus is particularly placed on the recent work of conformal spiral installations.

In [45], Nurnberger and Volakis presented a design that integrates the balun and feed structure into the planar radiating structure, which greatly simplifies the construction, making it highly attractive for conformal applications. As a continuation work, they discussed the physical characteristics of a cavity-backed slot spiral, as well as the associated infinite balun and termination designs in [46]. By introducing an arm termination in conjunction with the reflecting cavity and planar balun, a very thin yet extremely broadband slot spiral design was achieved in [47]. Several types of
novel spiral antennas were developed to accommodate the needs of integrating multiple frequencies and functionalities under a single aperture for automotive wireless communication systems [48–50]. Spirals can be viewed as consisting of many one-wavelength loops connected in a serial fashion (i.e. a lossy transmission line) [51]. Using a model based on such a lossy transmission line, impedance and wave velocity control was accomplished in a spiral antenna, which incorporates reactive distributed loading into its arms. This approach reduced size, particularly at low frequencies, and the model’s equivalent circuit was shown to be identical to that proposed for negative media transmission lines [52]. However, the necessity to employ many reactive elements made this loading less attractive in practice. Thus, a geometrical approach was pursued to realize inductive and capacitive loadings, necessary for wave slow down and therefore miniaturization [53]. The latter led to nearly optimal miniaturization as compared to recently developed UWB antenna theoretical limits in [10]. As is well known, metallic ground planes or cavity backings reduce antenna gain particularly for broadband configurations at low frequencies. Artificial magnetic conductors (AMC) or some other high impedance surfaces were employed to retain performance in conformal installations by utilizing their reflection phase feature [54–56]. For instance, a low-profile equiangular spiral antenna backed by an electromagnetic band gap (EBG) reflector was proposed in [57]. However, these approaches have bandwidth limitations, and they desire a large surface to be at or near resonance in the VHF/UHF band, making them impractical. Magneto-dielectric materials were investigated to increase the in-phase reflection bandwidth [58] and achieve miniaturization effect [59], respectively. In [60], a novel reflective surface consisting of a ferrite tile with variable conductive coating as a backing for a conformal UHF spiral antenna was proposed. A
remarkable low frequency gain improvement using a low cost ferrite loading was also presented in [61]. This design delivered a gain of 0 dB at 200 MHz for a 14” aperture (∼\(\lambda/4\) in size at 200 MHz) and -13 dB gain at 150 MHz (0.175\(\lambda\) in size at 150 MHz).

2.4 Summary

To summarize, an UWB antenna operating at VHF and UHF band that offers either monopole-like pattern or patch-like pattern has been continuously studied over the past few decades. To accommodate the lasting needs for compact, light-weight, and cost-effective broadband RF apertures, it is natural that consideration is given to systematically designing antenna shape for optimum radiation performances. The remainder of this dissertation is thereby dedicated to addressing this issue.
Chapter 3: Frequency-Scaled UWB Inverted-Hat Antenna

3.1 Introduction

A strong need exists for small size ultra-wideband (UWB) antennas. The volcano smoke antenna, originally proposed by Kraus [7], the conical configuration [8, 9] and the vivaldi-type antennas [62] are examples of wideband antennas. All these designs have the feature of tapering their surface away from the feed in an effort to maintain nearly constant impedance over a large bandwidth. Their low frequency limit is determined by the maximum antenna length and associated diffraction of unradiated fields that eventually reach the ends of the structure (viz. the discone caps or the vivaldi tips). To decrease their operational frequencies, their height must also be increased making them protruding and non-conformal. Broadband planar monopoles were considered in recent years for portable wireless devices where low-profile radiators are typically required. Among them, planar circular and elliptical discs [41], inverted-cones [42] and staircase printed monopoles [43] were shown to achieve large impedance bandwidth and stable patterns in the vertical plane. But, their patterns were not omnidirectional in the horizontal plane. Three-dimensional (3-D) antenna structures (constructed of planar elements) were also reported in [39, 63]. However, the latter were not low-profile as their height was more than $\lambda/6$ at the
lowest frequency of operation. Parasitic elements were introduced to the body-of-revolution (BoR) structure to increase the impedance bandwidth while maintaining a low-profile [34, 35]. Nevertheless, the resulting azimuth plane pattern was not omnidirectional at high frequencies. Being low profile, spiral antennas have been extensively studied as alternative frequency-independent radiators [3–6]. Much effort has been expanded in recent years to reduce the size of conformal spirals via volumetric inductive loading, circuit loading and material loading [51, 53]. Based on these efforts, a 6.25” aperture (1.5” in height) spiral can deliver -15 dBi gain down to 170 MHz. However, the above mentioned miniaturization is typically lost when the coiled spiral is placed conformally on vehicle platforms. Besides, spirals radiate a patch-like pattern, not desirable for over the horizon low frequency operation.

In this chapter, a novel low-profile inverted-hat antenna (IHA) that can operate not only down to low VHF frequencies but also retain good performance at higher frequencies (up to 2 GHz) is proposed. In contrast with the spiral, this design maintains excellent vertical polarization (VP) down to grazing incidences. Specifically, as an initial effort, a double-ellipse IHA (see Figure 3.2(a)), which is 15” (λ/10) in width and 6” (λ/25) in height, is shown to have a gain of -15 dBi down to 75 MHz in the presence of a 21” diameter circular ground plane. A broadband operation (50 MHz–2 GHz) is also achieved by properly controlling the diffracted fields at each local elliptical segment forming the IHA. However, the presented design is not optimal as its impedance exhibits large variations. A systematic approach to design the outer profile of the IHA subject to aperture size and height is then developed. The resulting multi-ellipse IHAs achieve controllable input impedance and uniform radiation pattern by maintaining a similar scattering geometry in terms of wavelength. A
seven-ellipse IHA prototype is designed and fabricated, showing good agreement between measurements and calculations. The proposed UWB configuration is targeted for Single-Channel Ground-Air Radio System (SINGARS) as well as VHF and UHF band.

### 3.2 Inverted-Hat Antenna Operating Principle

The proposed broadband antenna concept is depicted in Figure 3.1. This design can be considered an evolution of the volcano smoke structure. However, although its feed is of the same type, its rise (height) above the ground comprises several curved segments that extend laterally to maintain a low profile. The mono-blade antenna described by Wicks et al. [64] presents an analogous modification to lower the height of the vivaldi-type. In contrast, the proposed antenna is more associated with blade-type antennas with its lower surface formed by multiple, smoothly connected concave and convex elliptical segments whose curvature is logarithmically related to the distance.

![Figure 3.1: Configuration of the UWB inverted-hat antenna (IHA): (a) cross-sectional diagram and (b) 3-D model.](image)

Figure 3.1: Configuration of the UWB inverted-hat antenna (IHA): (a) cross-sectional diagram and (b) 3-D model.
from the feed point. Such a surface profile enhances frequency independence. That is, higher frequencies diffract from the region closer to the feed, whereas the lower frequencies diffract from sections farther away from the feed. As a result, the proposed antenna produces frequency-invariant impedance and patterns, necessary for UWB performance.

Referring to Figure 3.1, the entire structure with its pointed feed section resembles that of an inverted hat. It is therefore referred to as the inverted-hat antenna (IHA). Its operation is based on the excitation of a traveling wave between the ground plane and the top portion of the IHA. Thus, the higher frequency operational limit is influenced by the presence of higher order traveling modes. That is, as the distance between the ground plane and the inverted-hat surface becomes electrically larger, higher order modes will be eventually supported, causing mismatches and pattern distortions. However, the lower operational frequency of the IHA is a function of its overall size and curvature.

### 3.3 Initial Shaping

A parametric study was carried out in order to find the minor and major axes of the ellipses (see Figure 3.2) subject to achieving best performance at the lowest possible frequency. The major and minor radii of the ellipse closest to the feed were varied from 1” to 3” with a step size of 1”. Final values for these axes are given in Figure 3.2, completely prescribing the double-ellipse and triple-ellipse IHA geometry.
Figure 3.2: Geometry of the 15” × 6” (a) double-ellipse and (b) triple-ellipse IHA.

Figure 3.3: Fabricated 15” × 6” double-ellipse IHA prototype: (a) side view and (b) antenna with a 21” diameter ground plane and radome.
The fabricated double-ellipse IHA, shown in Figure 3.3(a), extending to 15” in aperture width and 6” in total height (see Figure 3.2(a)) was built via standard metal spinning technology. This metalworking process is typically used to form a disc or tube of metal into an axially symmetric part when keeping it rotating at high speed. The IHA was then mounted on a 21” diameter metallic ground plane (see Figure 3.3(b)) for testing its radiation properties from 30 MHz up to 2 GHz. As can be understood, the low frequency (below 300 MHz) measurements were done outdoors, which is displayed in Figure 3.4(a). Specifically, a log-periodic dipole array (LPDA) was used for transmission and a standard horn (AEL1734) was employed for reference down to 100 MHz. Below 100 MHz, a set of dipoles were used for calibration. The high frequency measurement (above 300 MHz) was performed in the anechoic chamber as seen in Figure 3.4(b). Figure 3.5 shows a comparison of the measured and simulated realized gain when the IHA radiates on a 21” diameter metallic ground plane. In simulation, electric field density was first computed over a radiation sphere,

![Figure 3.4: (a) Outdoor and (b) indoor measurement setup for the fabricated 15” × 6” double-ellipse IHA prototype in Figure 3.3.](image)

21
followed by comparing it with that of an isotropic radiator to obtain the antenna directivity $D$ at a certain angle. The input impedance $Z_{in}$ was calculated via dividing a unit voltage by the current on the feed element. For a lossless case, the realized gain $G$ is given by

$$G = (1 - |\Gamma|^2)D,$$  \hspace{1cm} (3.1)

where

$$|\Gamma| = \frac{Z_{in} - Z_0}{Z_{in} + Z_0}.$$  \hspace{1cm} (3.2)

$Z_0$ is the system impedance.

Figure 3.5: Measured and simulated horizontal realized gain of the 15” × 6” IHAs in Figure 3.2.
As seen, the agreement between measurement and calculation is reasonably good, and this holds even at lower frequencies. More remarkable is the realized gain performance showing -15dBi at 75 MHz, where the antenna electrical size is only $\lambda/10$ (15") in width and $\lambda/25$ (6") in height. Figure 3.5 also shows the gain curve in presence of an infinite ground plane. In this case, the -15 dBi gain point has reached down to 50 MHz, roughly 25 MHz below that of the 21” ground plane. Also, the gain at higher frequencies reaches +5 dBi, a value more in line with printed/planar antennas. Note that radiation is only allowed in the upper half-sphere when a monopole being placed over an infinite ground plane, resulting in a $2 \sim 3$ dB higher directivity as compared to the finite ground plane case when radiation exists in the full space. Edge diffracted currents around finite ground plane also lead to beam tilting, reducing the horizontal

Figure 3.6: Simulated 3-D gain patterns of the 15” × 6” double-ellipse IHA: (a) $f = 50$ MHz, (b) $f = 500$ MHz, (c) $f = 1500$ MHz, and (d) $f = 2000$ MHz.
realized gain. A ferrite ground plane is expected to provide additional gain improvements at low frequencies [65]. This is because a “material mode” might be excited within an impedance matched layer \((\epsilon_r = \mu_r)\) of the ferrite ground plane, providing better matching at the low frequencies.

Calculated 3-D radiation patterns of the 15” \(\times\) 6” double-ellipse IHA are shown in Figure 3.6. As can be seen, below 1 GHz, the patterns are pretty much of similar shapes. However, the distortion at higher frequencies is due to higher order modes. Of particular interest is certainly the input impedance down to low frequencies. From Figure 3.7(a), it is clear that the input resistance of the double-ellipse IHA does not level off at 50 \(\Omega\) (but rather at around 150 \(\Omega\)). However, the triple-ellipse design, as depicted in Figure 3.2(b), exhibits a resistance that levels off at 10 \(\Omega\) towards higher frequencies. More importantly, this design has a reactance much closer to zero (see Figure 3.7(b)). This is primarily due to the inductive component introduced by the 1” radius tip at the feed. However, since the design objective is to achieve the maximum

![Figure 3.7: Impedance comparison of the 15” \(\times\) 6” double-ellipse and triple-ellipse IHA (on an infinite ground plane): (a) real part and (b) imaginary part.](image-url)
Figure 3.8: Calculated reflection coefficients of the 15” × 6” IHAs in Figure 3.2.

low frequency VP gain, no emphasis has been placed on the impedance bandwidth. As a result, these two antennas are poorly matched to 50 Ω (see Figure 3.8). The multi-ellipse IHAs introduced in the next section will resolve this matching issue and improve existing designs.

Having validated the IHA, a study was then conducted to evaluate its performance for different heights and aperture sizes. To this end, Figure 3.9(a) provides curves showing the frequency at which the antenna reaches a gain of -20 dBi, -15 dBi and -10 dBi for different apertures when the height is fixed at 6”. Correspondingly, Figure 3.9(b) shows similar curves for different heights with the aperture size fixed at 15”. Specifically, from Figure 3.9(a) and Figure 3.9(b), it is observed that a 24” × 6” or 15” × 10” double-ellipse IHA can achieve -15 dBi down to 40 MHz.
Figure 3.9: Design curves relating aperture size, height and frequency in achieving a pre-specified gain of -10 dBi, -15 dBi and -20 dBi for the double-ellipse IHA (infinite ground plane): (a) antenna aperture size fixed at 15” and (b) antenna height fixed at 6”.

3.4 Multi-ellipse Design for Frequency-Independent Operation

In the previous section, the proposed double-ellipse and triple-ellipse IHA were shown to achieve good performance at both low and high frequencies. However, the presented design was not optimal as its impedance exhibited large variations. As a continuation work, this section provides a systematic approach to generate a frequency-scaled IHA structure subject to aperture width and height constraints. An experimental verification is also given. The proposed compact wideband antenna exhibits frequency-independent behavior by introducing a moderate number of appropriately scaled elliptical segments and by controlling the outer surface growth profile. A modified seven-ellipse IHA is demonstrated to provide a 10:1 impedance
bandwidth with satisfactory radiation properties. The simulation and measurement results are in reasonably good agreement.

### 3.4.1 Antenna Profile Description

The geometry of the proposed IHA is shown in Figure 3.10. Its feed generates outward travelling waves giving vertically polarized radiation over the horizon. Referring to [14], an UWB antenna can be formed by assembling structures with similar geometry scaled with distance from a reference feed point. To achieve a nearly invariant radiation over a large bandwidth, the IHA’s surface was constructed by many elliptically contoured (concave and convex) surfaces as depicted in Figure 3.10(b).

The arrangement of theses elliptical surfaces was inspired by the exponential spiral’s growth. Specifically, the major radii of the ellipses forming the IHA were chosen to be related with the $\phi = 0$ spiral crossing points. Referring to Figure 3.10(a), $X_n$—the $x$-axis crossings of the spiral are given by

$$X_n = e^{a\phi_n},$$

where $\phi_n = \pi, 3\pi, 5\pi, 7\pi, \ldots$, and $a$ is the spiral growth rate. It is surmised that if the IHA’s outer surface consists of $N$ ellipses, then

$$X_n = e^{a(2n-1)\pi} \quad \text{for} \quad n = 1, 2, \ldots, N. \quad (3.4)$$

As dictated, the frequency-scaled operation requires that the radiating structure sections be similar, but scaled proportionally to the wavelength. A constant antenna impedance, gain, and pattern can therefore be maintained as frequency varies. Different frequencies are designed to diffract at appropriately scaled local elliptical segments. The major radius ratio between adjacent ellipses is therefore a constant, implying
Figure 3.10: (a) Spiral pattern for the determination of the major radius of each elliptical segment. (b) Configuration (side view) of a frequency-scaled triple-ellipse inverted-hat antenna (IHA) having an aperture width of $w$ and height of $h$; for each elliptical segment, $x_i$ is the major radius and $y_i$ is the minor radius. (c) 3-D view of a frequency-scaled triple-ellipse linear IHA.
\[
\frac{X_{n+1}}{X_n} = e^{a\pi}.
\] (3.5)

The antenna is fed from the high frequencies region, and the edge discontinuity is the primary diffraction structure for the low frequencies. Therefore, the \(N\)th ellipse (uppermost one) corresponds to the lowest frequency. From Figure 3.10(b), if the total IHA width is \(w\), \(X_N\) was chosen to be \(w/2\) to account for the traveling distance of the lowest frequency before diffraction, viz

\[
X_N = e^{a(2N-1)\pi} = \frac{w}{2}.
\] (3.6)

Thus,

\[
a = \frac{1}{(2N-1)\pi} \ln \frac{w}{2}.
\] (3.7)

Consequently, the major radius of other elliptical segments \(X_n\) (\(n = 1, 2, \ldots N - 1\)) can be determined by inserting (3.7) into (3.4).

As seen later, the outer surface growth profile of the IHA provides input impedance control. This is achieved by adjusting the ratio of major radius \(X_n\) to minor radius \(Y_n\) of the elliptical segments. Specifically, (see Figure 3.12(a))

\[
Y_n = \begin{cases} 
\gamma_1 \cdot X_n & : \text{linear} \\
M \cdot (e^{\gamma_2 \cdot X_n} - 1) & : \text{convex} \\
M \cdot \ln(\gamma_3 \cdot X_n + 1) & : \text{concave}
\end{cases}
\] (3.8)

Herein, the exponential and logarithmic functions are used to represent the convex and concave mean profiles of different IHAs. Substituting \(X_N = w/2\) and \(Y_N = h\) into equation (3.8), the indicated \(\gamma_i\) factors are given by

\[
\begin{align*}
\gamma_1 &= \frac{h}{(w/2)} \\
\gamma_2 &= \frac{1}{(w/2)} \cdot \ln\left(\frac{h}{M} + 1\right) \\
\gamma_3 &= \frac{1}{(w/2)} \cdot (e^{\frac{h}{M}} - 1).
\end{align*}
\] (3.9)
Also, $M$ is a parameter used to adjust the curvature of the convex or concave curve to improve impedance matching. For the following design, $M = 0.5$ is found to be a good choice. Note that the major and minor radius reach $w/2$ and $h$, respectively, for the uppermost ellipse. These values do, of course, control the lowest frequency of operation. This is true for all types of IHA antennas (linear, convex, and concave) as depicted in Figure 3.12(a).

As seen from Figure 3.10(b), the $X_n$ and $Y_n$ values must be scaled down to satisfy the pre-specified aperture width $w$ and height $h$. Doing so, the revised $x_n$ and $y_n$ become

$$x_n = f_x \cdot X_n$$
$$y_n = f_y \cdot Y_n.$$  \hspace{1cm} (3.10)

The scaling parameters $f_x$ and $f_y$ are appropriately given by

$$f_x = \frac{(w/2)}{\sum_{n=1}^{N} X_n}$$
$$f_y = \frac{h}{\sum_{n=1}^{N} Y_n}.$$  \hspace{1cm} (3.11)

For the linear profile multi-ellipse IHA, $f_x$ and $f_y$ are identical. An explicit illustration of the linear triple-ellipse IHA geometry is shown in Figure 3.10(b), accompanied by a 3-D view in Figure 3.10(c).

### 3.4.2 Outer Surface Control for Frequency-Independent Operation

Following the geometrical prescription given above, now consider several different IHA outer surfaces (up to eleven ellipses). In all cases, the IHA is 15” (381 mm) wide and 6” (152 mm) tall. Full-wave simulations were carried out using commercial computational electromagnetics software FEKO. The geometry and corresponding electrical performance for various multi-ellipse IHA are given in Figure 3.11 (The
Figure 3.11: (a) 15” wide and 6” tall IHA geometry for several multi-ellipse outer surfaces, (b) realized gain in the horizontal plane ($\theta = 90^\circ$), (c) input resistance, and (d) input reactance.

double-ellipse IHA is referred to the one given in section 3.3). In particular, an infinite ground plane was assumed for all calculations below. As seen, when the number of ellipses describing the outer surface increases, the resistance approaches 50 $\Omega$ and the reactance levels off at 0 $\Omega$, indicating a very desirable matching condition.

Furthermore, the realized gain over horizon achieves a fairly constant level of about 5 dB at high frequencies. This is a result of the multiple diffraction points forming the outer surface of the multi-ellipse IHA design. Indeed, smooth transitions between
each segment allow for continuous diffraction along the whole curved antenna profile, resulting in UWB performance. It is remarked that precisely locating the diffraction point (analogous to the active region of a spiral antenna) for a specific frequency is part of future work.

As can be seen from Figure 3.11(a), the points interconnecting the ellipses lie on a linear curve. However, for better impedance control, it is preferred to consider

![Figure 3.11(a)](image)

Figure 3.12: Impedance calculation for different outer surface profiles of the eleven-ellipse design using 15” wide and 6” tall IHA: (a) geometry, (b) resistance, and (c) reactance.
designs where these points lie on a non-linear curve (say, convex or concave). Indeed, Figure 3.12 shows that the concave profile exhibits higher resistance than the linear one, whereas the convex profile gives lower input resistance value. This observation is similar to that for the conical antenna of which the input impedance is largely determined by the bottom opening angle.

To better illustrate impedance flexibility, a 24” (610 mm) wide and 6” (152 mm) tall IHA with different outer surface profiles was also investigated. Enlargement of the aperture size is, of course, needed to achieve lower operational frequency. However, a larger aperture size leads to lower input resistances, causing mismatch and gain reduction. In particular, calculations were carried out for a conical antenna, single-ellipse IHA and eleven-ellipse concave IHA. As shown in Figure 3.13, the input

![Diagram of IHA designs](image)

Figure 3.13: Impedance performance for two IHA designs vs. a conical antenna all being 24” wide and 6” tall: (a) geometry, (b) resistance, and (c) reactance.
resistance of the conical antenna is roughly 40 Ω for this specific width/height ratio. It is observed that the single-ellipse IHA exhibited high impedance, and only the eleven-ellipse concave IHA achieved 50 Ω, with relatively low oscillations, necessary for UWB operation. That is, for a given aperture size and height, the multi-ellipse IHA is capable of achieving a desired input impedance.

To sum up, the multi-ellipse IHA is better than a conventional conical antenna in the following two aspects:

1. **Impedance Control.** Thanks to various practical reasons, there is often a strong restriction on antenna height, whereas the footprint (width) is allowed to be relatively large. IHA’s impedance control feature is thereby quite attractive to extremely low profile designs, of which the radiation resistance is typically very low. For instance, the input impedance of a 24” wide and 2” tall conical antenna is only 10 Ω beyond the first resonance. Comparing to the simple concave tapered profile that increases the average impedance level at the price of introducing significant oscillations (see the single-ellipse IHA in Figure 3.13), the multi-ellipse concave IHA is able to achieve better matching since multiple elliptical segments effectively suppress large impedance variations as shown in Figure 3.13 (b) and (c).

2. **Uniform Pattern.** Figure 3.14 compares the $E$-plane radiation pattern of a 15” wide and 6” tall conical antenna to that of the 11-ellipse convex IHA in Figure 3.12 at 2 GHz. Clearly, the IHA design provides much broader 3-dB beamwidth. This pattern uniformity implies that high frequencies has radiated out prior to reaching the top-edge, validating the concept of frequency separation by creating a series of scaled diffraction structures.
Figure 3.14: $E$-plane radiation pattern comparison between a 15” × 6” conical antenna and the 11-ellipse convex IHA in Figure 3.12 at 2 GHz.

### 3.4.3 Measurement Demonstration

Measurements were carried out to validate the multi-ellipse IHA design. As demonstrated in the previous subsection, the 11-ellipse convex IHA delivers a highly uniform radiation pattern at 2 GHz. To compensate its low radiation resistance ($\sim 25 \, \Omega$), the IHA tip near the feed was replaced by a tapered concave profile as depicted in Figure 3.15(a). This resulted in a modified seven-ellipse 15” (381 mm) × 6” (152 mm) IHA that was enclosed in a 16” (406 mm) radome having a dielectric constant of $\epsilon_r = 2.2$ (see Figure 3.15(b)).

A reasonably good agreement is observed between measurements and calculations. The reflection coefficient and realized gain in the horizontal plane ($\theta = 90^\circ$) are depicted in Figure 3.15(c)–(d). Fed by a 50 $\Omega$ coaxial cable, the modified seven-ellipse IHA prototype provided a 10 dB return loss from 0.2 GHz to 2 GHz, indicating
a 10:1 impedance bandwidth. Satisfactory radiation was also achieved. It is worth mentioning that one of the reference horn antennas had a lowest operational frequency of 1 GHz. As a result, some gain inconsistencies can occur at this frequency. Due to the 17” (432 mm) finite ground plane, both measured and simulated gain over the horizon are lower than that of the infinite ground plane case. Use of ferrite material on the ground plane is likely to recover some of this gain reduction [65].

Figure 3.15: Measurements vs. calculation for a modified seven-ellipse IHA: (a) geometry, (b) fabricated prototype with radome, (c) reflection coefficient, and (d) realized gain in the horizontal plane ($\theta = 90^\circ$).
Figure 3.16: Measured and calculated normalized gain patterns for a modified seven-ellipse IHA (in the elevation plane).

Figure 3.16 shows the calculated and measured normalized gain patterns in the elevation plane for different frequencies. Clearly, the IHA exhibits the desired monopole-like patterns at all frequencies. This performance demonstrates the effectiveness of the diffraction points at the junctions of the adjoining ellipses forming the IHA outer surface.

3.5 Summary

In this chapter, a novel compact low-profile UWB inverted-hat antenna (IHA) was presented. The proposed broadband RF aperture achieves a reasonably good VP gain at the low VHF frequencies while maintaining a desired impedance matching at
higher frequencies (up to 2 GHz). A fabricated double-ellipse IHA design was 15” in width and 6” in height, and achieved a gain of -15 dBi at 75 MHz (λ/10 aperture) when mounted on a 21” finite ground plane. On an infinite ground plane, simulations showed the same performance down to 50 MHz for the same aperture and height, which is quite close to the theoretical limit presented in Figure 1.2. The obtained calculated and measured gain data were found to be in good agreement. Design curves were given to provide design guidelines in terms of gain versus aperture width and height sizes. As a continuation work, a frequency-scaled IHA design, of which the outer surface is composed of multiple ellipses that follow the growth rate of an exponential spiral, was presented. The mathematical details for constructing the outer IHA surface aimed at providing UWB performance was provided. A nearly frequency-independent behavior is observed for a moderate number of elliptical segments used to construct the IHA. Impedance control could also be achieved by employing different outer surface profiles. A modified seven-ellipse design was finally tested to validate the concept. This design approach manages to maintain a similar scattering geometry

Figure 3.17: A 15”×6” double-ellipse IHA was mounted on top of a ground vehicle to test its radiation properties.
in terms of wavelength, providing controllable input impedance and uniform radiation patterns, important in guiding the development of future low-profile UWB antennas. In particular, the proposed compact wideband antenna can be applied across a variety of vehicular platforms, as shown in Figure 3.17.
Chapter 4: Body-of-Revolution Monopole Antenna Analysis

4.1 Introduction

In Chapter 3, a compact body-of-revolution (BoR) inverted-hat antenna (IHA) operating from 50 MHz to 2 GHz with excellent vertical polarization near grazing incidences is presented. The low-frequency operation of this antenna is primarily limited by its maximum width and height. Correspondingly, the high-frequency performance is controlled by introducing a moderate number of appropriately scaled elliptical segments and by adjusting the outer surface growth profile (see Figure 4.1). This compact wideband aperture offers controllable input impedance and uniform radiation patterns over the entire operating bandwidth [66, 67]. To further improve its bandwidth, it is necessary to consider different outer shapes other than the shown

Figure 4.1: Configuration of an UWB inverted-hat antenna: (a)cross-sectional diagram and (b)photograph of a fabricated 24” wide and 6” tall prototype.
collection of elliptical curves in Figure 4.1. Generation of such shapes requires a fast analysis tool coupled with formal optimization techniques (e.g. the genetic algorithm (GA)).

As the antenna being considered is in the form of body-of-revolution, it is appropriate to also use a BoR method of moments (MoM) analysis. Harrington and Mautz [68] were the first to develop a MoM-based solution for BoRs. This method reduces a 3-D analysis to a two and a half dimensional one. Consequently, the solution time is reduced dramatically, making it suitable for insertion in an optimization loop. Successive efforts involve extending the solution to homogeneous [69] and inhomogeneous [70, 71] dielectric BoRs. Although BoRs have been used for scattering problems [72–74] and a few antenna designs [75–77], to best of my knowledge, the BoR-MoM analysis has not been considered for UWB antennas.

The general BoR-MoM analysis can be greatly simplified for an electric dipole or monopole configuration utilizing the fact that its excitation is always vertically polarized (no circumferential dependence). Formulations are given in this chapter, along with presenting a validation via the measurement of a 24” wide and 20” tall dome-dipole antenna over a 100:1 bandwidth. The calculations and measurements are shown to be in reasonably good agreement. This agreement provides confidence in integrating this BoR analysis into a formal optimization technique to design low-profile BoR monopole antennas for UWB operation. A design example resulting from the GA optimization of a 6” tall IHA is shown to deliver a nearly frequency-independent performance across the frequency range from 1 GHz to 6 GHz.
4.2 Body-of-Revolution Moment Method

As depicted in Figure 4.1, the proposed monopole will be placed over a metallic surface (such as the top of a ground vehicle). Therefore, for this analysis, it can be assumed that an infinite ground plane as the installation platform of the monopole fed at its bottom. Under this assumption, image theory can be invoked to instead consider a BoR dipole configuration (see Figure 4.2). The goal is to compute the outer surface currents using a BoR moment method as discussed in the introduction of this chapter. Following [68], the surface current density \( \vec{J}(\vec{r}) \) on the BoR is expanded as follows

\[
\vec{J}(\vec{r}) = \sum_{m=-\infty}^{\infty} \sum_{n=1}^{N} [a_{mn}^t \vec{J}_{mn}^t(\vec{r}) + a_{mn}^\phi \vec{J}_{mn}^\phi(\vec{r})],
\]

(4.1)

Figure 4.2: Discretization and triangular basis functions for a body-of-revolution (BoR) dipole antenna analysis.
where $\vec{J}_{t,mn}(\vec{r})$ and $\vec{J}_{\phi,mn}(\vec{r})$ refer to the longitudinal and azimuthal tangential currents of the BoR surface. Also, as usual, $a_{t,mn}$ and $a_{\phi,mn}$ are the coefficients to be found via the moment method. The total number of basis functions is $N$. As seen from Figure 4.2, $\hat{t}$ is the unit vector tangent to the BoR surface and $\hat{\phi} = \hat{t} \times \hat{n}$, where $\hat{n}$ is the unit normal to the BoR generating curve. The $\phi$ (circumferential) dependence of the currents are sinusoidal due to their $2\pi$ period. Thus,

$$\vec{J}_{t,mn}(\vec{r}) = \hat{t}f_n(t)e^{j m \phi}$$

$$\vec{J}_{\phi,mn}(\vec{r}) = \hat{\phi}f_n(t)e^{j m \phi}$$

where $f_n(t)$ refers to the basis functions along the longitudinal direction ($t$-direction) of the BoR dipole. $f_n(t)$ can be further written as

$$f_n(t) = \frac{T_n(t)}{\rho(t)}.$$  \hspace{1cm} (4.3)

Here, $T_n(t)$ is a local triangular basis function spanning four segments, and $\rho(t)$ is a normalization factor. The trapezoidal rule is used to perform the integrations using four samples, as shown in Figure 4.3. More explicitly,

$$T_1 = \frac{1}{4}, \quad \hat{T}_1 = 1$$

$$T_2 = \frac{3}{4}, \quad \hat{T}_2 = 1$$

$$T_3 = \frac{3}{4}, \quad \hat{T}_3 = -1$$

$$T_4 = \frac{1}{4}, \quad \hat{T}_4 = -1$$  \hspace{1cm} (4.4)

To solve for the coefficients $a_{t,mn}$ and $a_{\phi,mn}$, continuity of the tangential fields through the BoR surface is enforced to construct the electric field integral equation (EFIE), yielding

$$-\frac{j}{\omega \mu} \hat{t} \cdot \vec{E}(\vec{r}) = \int_S \left(1 + \frac{1}{k^2} \nabla \nabla \cdot \right) \sum_{n=1}^N \left[a_{t,mn} \vec{J}_{t,mn}(\vec{r}) + a_{\phi,mn} \vec{J}_{\phi,mn}(\vec{r}) \right] \frac{e^{-jkr}}{4\pi r} d\vec{r}.$$  \hspace{1cm} (4.5)
As the antenna feed is modeled by a delta gap source [78], the impressed electric field $\vec{E}_i(\vec{r})$ only exists inside one triangular basis function (at the origin) and is zero outside. That is

$$\vec{E}_i(\vec{r}) = \begin{cases} \frac{V_0}{\Delta z} \hat{z} & : \vec{r} = 0 \\ 0 & : \text{else} \end{cases},$$

where $\Delta z$ is the width of the gap, and $V_0$ is set to unity.
Upon employing Galerkin’s testing, the matrix system \([Z][I] = [V]\) for each mode \(m\) is obtained, which is expressed implicitly as

\[
\begin{bmatrix}
Z_{tt}^m & Z_{t\phi}^m \\
Z_{\phi t}^m & Z_{\phi\phi}^m
\end{bmatrix}
\begin{bmatrix}
I_m^t \\
I_m^\phi
\end{bmatrix} =
\begin{bmatrix}
V_m^t \\
V_m^\phi
\end{bmatrix},
\]  

(4.7)

where \(I_m^t\) and \(I_m^\phi\) are rows containing the unknown coefficients \(a_{tmn}\) and \(a_{\phi mn}\). The submatrices in \(Z\) and \(V\) refer to the impedance matrices and excitation vectors, respectively.

For the geometry at hand, the excitation is independent of \(\phi\), which excites the \(m = 0\) mode only. Also, there is no coupling between \(J^t(\vec{r})\) and \(J^\phi(\vec{r})\), resulting in \(Z_{0t}^\phi = Z_{0\phi}^{t*} = 0\). Further, since the excitation has only an \(E^t\) component, no \(\phi\)-directed current exists \((I_0^\phi = 0)\). The matrix system (4.7) reduces to

\[
Z_0^{tt} \cdot I_0^t = V_0^t.
\]  

(4.8)

The impedance matrix elements of this system are given by

\[(z_0^{tt})_{\alpha\beta} = \sum_{p=1}^{4} \sum_{q=1}^{4} [j\omega\mu T_p T_q (\sin v_p \sin v_q G_{20} + \cos v_p \cos v_q G_{10}) - \frac{j}{\omega \varepsilon} T_p T_q G_{10}], \]

(4.9)

with

\[
G_{10} = \Delta t_p \Delta t_q \int_{0}^{\pi} g(\phi) d\phi
\]

(4.10)

and

\[
g(\phi) = \frac{e^{-jkR_{pq}}}{R_{pq}}
\]

(4.11)

\[
R_{pq} = \begin{cases} 
\sqrt{(\rho_p - \rho_q)^2 + (z_p - z_q)^2 + 2\rho_p\rho_q(1 - \cos \phi)}, & \text{if } p \neq q \\
\sqrt{(\Delta t_p/4)^2 + 2\rho_p^2(1 - \cos \phi)}, & \text{if } p = q 
\end{cases}
\]

(4.12)

The index \(\alpha\) refers to the \(\alpha\)th triangular basis function spanning the \(\alpha\)th annulus on the BoR surface \(S\) and the index \(\beta\) refers to the \(\beta\)th triangular basis function.
The angles \( v_p \) and \( v_q \) are measured between the \( z \)-axis and the segments \( p \) and \( q \), respectively. \( \rho_p \) is the value of \( \rho(t) \) at the \( p \)th point along the \( \alpha \)th annulus and \( \Delta t_p \) is the segment length at the \( p \)th point along the BoR arc.

The excitation vector \( V^t_0 \) has nonzero entries only for basis functions having support on that element, which is given by

\[
v^t_0 = 2\pi \sum_{p=1}^{4} T_p \Delta t_p \frac{V_0}{\Delta z}.
\]

The surface currents \( \vec{J}(\vec{r}) \) can, thus, be determined by solving for \( I^t_0 \). The far-zone radiated electric field \( \vec{E}(\vec{r}) \) is calculated via

\[
\vec{E}(\vec{r}) = -\frac{j\omega\mu}{4\pi} \int \int \int \vec{J}(\vec{r}') e^{jk\vec{r} \cdot \vec{r}'} d\vec{r}'.
\]

The antenna input impedance \( Z_{in} \) is given by

\[
Z_{in} = \frac{V_0}{I_{in}},
\]

where \( I_{in} \) is calculated by integrating \( \vec{J}(\vec{r}) \) over \( 2\pi \) on the feed element.

### 4.3 Numerical Analysis and Experimental Validation

Figure 4.4 shows the fabricated broadband dome-dipole prototype to validate the BoR analysis. The dipole configuration can be viewed as a 3-D version of the flare dipole [44, 79]. A similar configuration with an ellipsoidal outer surface was considered in [80] and achieved a 1.5:1 bandwidth. The actual geometry in Figure 4.4 has an exponential outer surface and its height of 20” and width of 24” were chosen to achieve 0 dB gain down to 100 MHz. The mathematical description of the tapered profile is given by

\[
z = 1.7(e^{0.161y} - 1).
\]
In (4.16), it is assumed that the origin of the coordinate system is located at the feed point and \( z \) refers to the vertical axis. The growth rate \( \alpha = 0.161 \) was chosen so that the electrical separation between the upper and lower surfaces would be relatively small at high frequencies, which avoids pattern distortions associated with most electrically large antennas. Two aluminum dome-like components were fabricated using standard metal spinning technology, and supported by three 0.75” thick PVC pipes, as shown in Figure 4.4.

The prototype antenna described above was simulated using the BoR analysis. Measurements were then performed to validate the analysis. Figure 4.5 and Figure 4.6 give the voltage standing wave ratio (VSWR) and realized gain (in the horizontal plane), respectively. These plots demonstrate the wideband performance of the exponential monopole. Specifically, the antenna achieves VSWR < 2 from 180 MHz up to 2 GHz with a 50 Ω reference impedance. The gain is also kept greater

![Figure 4.4: Fabricated 24” wide and 20” tall BoR dome-dipole antenna.](image)
than 0 dB within this range. Of importance is that simulations and measurements are in good agreement.

The same dome-dipole antenna has excellent pattern performance up to 10 GHz. The patterns for several frequencies are given in Figure 4.7. It is seen that the patterns are pretty much consistent over a 100:1 bandwidth from 100 MHz to 10 GHz, a critically sought after feature. As expected, the main lobe beamwidth decreases with frequency (as the aperture becomes electrically larger). Also, the desired maximum radiation direction is in the horizontal plane ($\theta = 90^\circ$) without beam tilting.

More importantly, the use of BoR principle leads to a dramatic reduction of the memory requirement as well as the calculation time. With the aid of Intel®Core™2 Duo Processor of 3 GHz and 4 GB RAM, for a moderate size problem (e.g. 41 equally spaced sampling points taken from 30 MHz up to 2 GHz), it is shown that

![Figure 4.5: Simulated and measured VSWR for the 24” wide and 20” tall BoR dome-dipole antenna in Figure 4.4.](image.png)
Figure 4.6: Simulated and measured realized gain in the elevation plane ($\theta = 90^\circ$) for the 24" wide and 20" tall BoR dome-dipole antenna in Figure 4.4.

The CPU time has been reduced by a factor of 7 using this in-house BoR code (155s) in comparison to the commercial MoM-based EM simulation package FEKO (1,116s). Furthermore, as the antenna becomes significantly electrically large (e.g. 12\(^\lambda\) wide and 10\(^\lambda\) tall at 6 GHz), the effectiveness of employing the BoR analysis is more notable, as shown in Table 4.1. The BoR analysis has been proven to be fast and accurate, and it will be used to optimize the IHA shape for an enhanced bandwidth in the next section.

<table>
<thead>
<tr>
<th>Solver</th>
<th># of unknowns</th>
<th>CPU time (s)</th>
</tr>
</thead>
<tbody>
<tr>
<td>FEKO</td>
<td>101,310</td>
<td>5,306</td>
</tr>
<tr>
<td>BoR</td>
<td>249</td>
<td>56</td>
</tr>
</tbody>
</table>

Table 4.1: Computational statistics of the 24" wide and 20" tall BoR dome-dipole antenna in Figure 4.4 at 6 GHz using FEKO and BoR analysis.
Figure 4.7: $E$-plane radiation patterns for the 24” wide and 20” tall BoR dome-dipole using BoR analysis: (a) $f = 100$ MHz, (b) $f = 2$ GHz, (c) $f = 4$ GHz, (d) $f = 6$ GHz, (e) $f = 8$ GHz, and (f) $f = 10$ GHz.
4.4 Parameter Optimization of Inverted-Hat Antenna

In this section, the BoR analysis is integrated with the GA for IHA shape optimization subject to gain and size constraints. The goal is to achieve a frequency-independent design, which delivers a constant gain, constant impedance, and uniform radiation patterns over a large bandwidth (1 GHz–6 GHz). The latter are manifested via the objective function that is tweaked to minimize the impedance and gain variations across the band of interest while satisfying a pre-defined gain level (toward the horizon). Utilizing the weighted sum method, the objective function is constructed as

\[
\text{COST} = \alpha_{\text{real}} P_{\text{real}} + \alpha_{\text{imag}} P_{\text{imag}} + \alpha_{\text{dir}} P_{\text{dir}} + \alpha_{\text{ripple}} P_{\text{ripple}},
\]

where

\[
P_{\text{real}} = \sqrt{\frac{1}{N_f} \sum_{N_f} |R(f) - \text{avg}(R(f))|^2}, \quad \alpha_{\text{real}} = 0.5,
\]

\[
P_{\text{imag}} = \frac{1}{N_f} \sum_{N_f} |X(f)|, \quad \alpha_{\text{imag}} = 0.5,
\]

\[
P_{\text{dir}} = -\frac{1}{N_f} \sum_{N_f} P_{\text{dir}}(f), \quad P_{\text{dir}} = \begin{cases} D(f) : & \text{if } D(f) < 5 \text{ dB} \\ 5 \text{ dB} : & \text{else} \end{cases}, \quad \alpha_{\text{dir}} = 0.8,
\]

\[
P_{\text{ripple}} = \sqrt{\frac{1}{N_f} \sum_{N_f} |D(f) - \text{avg}(D(f))|^2}, \quad \alpha_{\text{ripple}} = 10.
\]

Here, \(R(f), X(f)\) and \(D(f)\) refer to the antenna input resistance, reactance, and directivity (toward the horizon) at frequency \(f\), respectively. \(N_f\) is the count of frequency steps between 1 GHz and 6 GHz (step size = 500 MHz). \(\alpha\) were chosen based on the relative importance of each objective. The aperture size, profile type, number of elliptical segments and curvature of the profile are encoded in chromosomes for GA operations, as listed in Table 4.2. Tournament selection, uniform crossover,
and random mutation strategy are deployed in the evolution process. The population of each generation is 16, maximum number of generation is 20, and the mutation rate is set to be 0.05.

<table>
<thead>
<tr>
<th>Width</th>
<th>10”, 12”, 14”, ..., 36”, 38”, 40”</th>
<th>4 bits</th>
</tr>
</thead>
<tbody>
<tr>
<td>Global Profile</td>
<td>convex/concave</td>
<td>1 bit</td>
</tr>
<tr>
<td>Curvature</td>
<td>0.1, 0.2, ..., 0.9, 1, 1.5, 2, 2.5, 3, 4, 5</td>
<td>4 bits</td>
</tr>
<tr>
<td># of Ellipse</td>
<td>3, 5, 7, ..., 29, 31, 33</td>
<td>4 bits</td>
</tr>
</tbody>
</table>

Table 4.2: Codification of a 6” tall IHA parameters for GA operation.

The resulting antenna profile (side view) from the GA is shown in Figure 4.8. It is a 28” wide and 6” tall 31-ellipse convex IHA on an infinite ground plane. For comparison, the profile of a 24” wide and 6” tall 11-ellipse concave IHA [67], which was designed to operate from 50 MHz to 2 GHz, is also depicted in the same figure. As seen in Figure 4.9, the GA optimized IHA exhibits a much wider bandwidth than the non-optimized case. By smoothing out the curvature close to the feed, maximum radiation toward the horizon is continuously maintained up to at least 6 GHz, leading to a highly stable directivity (see Figure 4.9(a)) and radiation patterns. As depicted

![Figure 4.8: Profile (side view) of the GA optimized and non-optimized 6” tall IHA.](image-url)
Figure 4.9: (a) Directivity (at $\theta = 90^\circ$) and (b) impedance of the optimized and non-optimized 6” tall IHA from 1 GHz to 6 GHz.

in Figure 4.10, the main beamwidth of the optimized IHA is quite broad, and the side lobe level is much lower in comparison to the non-optimized case. This was achieved by forcing minimization of the standard deviation (STD) of the directivity across the band in the GA objective function. Concurrently, the antenna input impedance

Figure 4.10: Radiation patterns ($E$-plane) of the (a) optimized and (b) non-optimized 6” tall IHA at 1 GHz, 3 GHz, and 6 GHz.
approaches to an almost constant value (around 15 Ω) as shown in Figure 4.9(b), which further demonstrates the effort toward accomplishing a frequency-independent design. Note that the optimized IHA is similar to an exponential monopole with a large bottom flare angle, exhibiting quite low radiation resistances. Therefore, the uniform pattern and impedance matching (to 50 Ω) are in conflict with each other for a BoR monopole design. It is natural that consideration be given to optimization techniques to find out a compromise solution. Alternatively, as will be shown in the next chapter, resistive loading is another effective means to resolve this issue.

4.5 Summary

In this chapter, the validation of a BoR analysis for an UWB dome-dipole antenna is presented. A 24” wide and 20” tall prototype was fabricated and demonstrated to deliver an operational bandwidth from 180 MHz to at least 2 GHz. VSWR and gain measurements for this prototype were compared with the calculated results based on a BoR moment method. Excellent agreements between measurements and simulations were observed over a 100:1 bandwidth. This agreement validates the BoR analysis, offering confidence on using it in optimizing the outer surface of low-profile UWB BoR monopole antennas. As a design example, it was shown that, utilizing the BoR analysis along with the GA, an optimized 6” tall IHA could deliver a nearly frequency-independent performance (e.g. constant gain, impedance, and pattern) across the frequency range from 1 GHz to 6 GHz. In the following chapter, a more general approach will be developed to systematically design the outer surface of a BoR monopole antenna for different goals set out by specific applications.
5.1 Introduction

Formal optimization techniques have been frequently employed in all kinds of antenna designs [44, 75–77, 81–95]. A conventional single-antenna optimization problem can be categorized into either parameter fine-tuning [44, 90–93] or topology exploration [84–88, 94]. The parameter fine-tuning approach refers to the case when the basic antenna configuration is known and only the geometrical and/or material parameters needs to be adjusted so as to achieve the desired objective, for example, the parameter optimization of the inverted-hat antenna (IHA) performed in Chapter 4. By contrast, if the designer’s prior EM knowledge cannot be applied to the initial design, or if it is favorable to set a least amount of constraints in the design space, the topology exploration is advantageous. This method usually yields improved or innovative antenna designs that exhibit non-intuitive shapes, providing a novel insight into antenna operating principles. The success of this type of optimization heavily relies on the accurate geometry description and appropriate cost function selection.

Computational methods for electromagnetics (e.g. FEM, FDTD, and MoM) utilizing the axisymmetric property of BoRs were reported in [75–77] integrated with
optimization engines for different design purposes, including an UWB monopole antenna with extremely low reflection coefficient [75], a size-reduced conical horn [76], and evolved-profile dielectric rod antennas with improved radiation patterns [77]. However, none of them discussed the completeness of its geometry description, making it difficult in judging whether the resulting designs are truly globally optimal. Besides, they were only targeted for single objective such as impedance matching or radiation pattern, not for optimizing a collection of objective functions simultaneously.

In this chapter, a systematic approach named the RGB method is proposed for the design of profiled UWB monopole antennas that satisfy specified radiation objectives and size constraints. As a class of topology exploration, a random walk scheme, which is able to achieve a fairly diverse geometric space, is used to generate the antenna profile. This is combined with a genetic algorithm (GA) along with a BoR-MoM analysis. The weighted global criterion method is then introduced to construct a single cost function consisting of multiple design objectives. Minimization of this cost function amounts to minimizing the difference between achievable and ideal design points within the criterion space, whereas the ideal (utopia) point represents the scenario when all of design objectives reach their own optimum simultaneously. Utilizing the proposed shape optimization method, two examples are presented and demonstrated to deliver improved performance in comparison to previously developed IHAs of the same size. Furthermore, resistive loading is considered as a post-optimization treatment for better matching and pattern stability. Measurements are given at the end of this chapter to verify the design concepts.
5.2 Optimization Methodology

A systematic approach to design and optimize BoR monopole antennas is proposed in this section. The design package consists of random walk (RW) for shape representation, genetic algorithm (GA) as optimizer, and BoR moment method analysis. It will be therefore referred to an RGB method. A system block diagram is shown in Figure 5.1. INTERFACE is the mastermind of the system; it operates as an “information hub” to interconnect the RW, the GA and the BoR. The steps of the optimization procedure are depicted in Figure 5.2.

![Figure 5.1: The proposed RGB method combining a random-walk (RW) scheme, a genetic algorithm (GA), and a body-of-revolution (BoR) solver.](image)

The optimization process amounts to designing the 2-D profile of a monopole having a given width $D$ and height $H$. To discretize the line forming the BoR, a random walk scheme was considered to represent the antenna shape. In particular, the line profile is made up of a sequence of line segments characterized by their length.
Figure 5.2: A flowchart of the optimization procedure.

$r_j$ and angle $\alpha_j$ (with respect to the horizontal plane), as depicted in Figure 5.3. Two bits were allocated for the length and angle of each step. Therefore, for each step during the $N$-step walk, there are $2^2$ possible lengths (1, 2.667, 4.333 and 6) and $2^2$ possible angles ($18^\circ$, $36^\circ$, $54^\circ$ and $72^\circ$). The resulting profile from the random walk was scaled down further to satisfy the pre-specified aperture size $D$ and height $H$ for the specific design. Doing so, the coordinate of node $i$ $(x_i, y_i)$ on the polyline is given
\[
x_i = \frac{\sum_{j=1}^{i} r_j \cos \alpha_j}{\sum_{j=1}^{N} r_j \cos \alpha_j} \cdot \frac{D}{2},
\]

\[
y_i = \frac{\sum_{j=1}^{i} r_j \sin \alpha_j}{\sum_{j=1}^{N} r_j \sin \alpha_j} \cdot H,
\]

with

\[
r_j \in \{1, 2.667, 4.333, 6\}
\]

\[
\alpha_j \in \{18^\circ, 36^\circ, 54^\circ, 72^\circ\}.
\]

To examine how well this RW approach actually represents a random curve from the start \((0, 0)\) to the end \((\frac{D}{2}, H)\), consider a generic \textit{random number generation} (RNG) scheme. Without loss of generality, a “normalized” square region defined by 4 vertices \(((0, 0), (0, 1), (1, 1)\) and \((1, 0))\) in the \(x\)-\(y\) plane is selected. A random number generator is triggered to produce a real number between 0 and 1, with equal
probability. There are \( M \) control vertices that completely define the subject’s 2-D profile; these points are generated in the following manner:

1. Run the random number generator \( M \) times and arrange the results in ascending order, forming a data array \( P = \{p_i|i = 1, 2, \ldots M\} \).

2. Repeat step 1. Name the resulting array \( Q = \{q_i|i = 1, 2, \ldots M\} \).

3. The coordinate of the control vertex \( i \) is \((p_i, q_i)\).

The profile is only allowed to grow monotonally away from the origin to avoid possible fabrication difficulties during the spinning process. The abovementioned RNG approach is generic enough, and thus it would be used as a benchmark to evaluate the proposed RW method. To this end, a Monte Carlo experiment was conducted. Specifically, antenna 2-D profiles \((D = 2, H = 1)\) were generated 50,000 times using the RW \((N = 6)\) and the RNG \((M = 5)\), respectively. The coordinate statistics were recorded and shown in Figure 5.4(a)–(e).
Figure 5.4: Coordinate statistics from a Monte Carlo simulation: antenna 2-D profiles with 5 control vertices are generated 50,000 times using the RW \((x_i, y_i)\) and the RNG \((p_i, q_i)\) scheme, respectively.
In all of the subplots, the leftmost ones represent $x_i$—the $x$-coordinate of the $i^{th}$ point generated by the RW scheme; the middle ones represent $y_i$—the $y$-coordinate of point $i$ yielded from the RW scheme; the rightmost ones stand for both $p_i$ ($x$-coordinate) and $q_i$ ($y$-coordinate) of the $i^{th}$ point resulted from the RNG scheme since the two coordinates have an identical probability distribution. A couple of interesting facts were found from this experiment. First, the statistical distribution of point $i$ is mirror-symmetric to that of point $M - i + 1$. This is because the problem of finding a path from $(0, 0)$ to $(1, 1)$ is essentially identical to that of forming a curve from $(1, 1)$ to $(0, 0)$, and the generation of points is inherently independent from one to another in both RW and RNG schemes. Second, the coordinate distributions of two schemes are, by and large, similar. This, in turn, demonstrates that the proposed RW approach is able to achieve a fairly diverse geometric space. It is also worth mentioning that this well-behaved distribution largely comes from the appropriate selection of the initial length array $r_j$ and the angle array $\alpha_j$ in the RW scheme. One would expect dissimilar patterns if different design parameters were used. Last but not least, the statistical discrepancy between two schemes indicates that a few geometries are unobtainable using the RW. For instance, in subplot (a), no point, which possesses an $x$-coordinate or $y$-coordinate that is greater than 0.6, was produced by the RW method, whereas a few do exist from the RNG scheme. However, this incompleteness is negligible as compared to the convenience brought by the RW method: it easily encodes geometric information into binary strings for GA operation. Hence, the RW scheme will be frequently employed as the profile generator in the remainder of this dissertation.

Successful design examples using constraint nonlinear programming [90], simulated annealing algorithm [75], and particle swarm optimization [94] are available in
the literature. As one of the global optimization techniques, the genetic algorithm that relies on Darwin’s theory of natural selection [81–83] is well-known for its robustness and ease of implementation. Following the geometrical prescription given above, the line vectors were encoded in chromosomes for GA optimization. The length and angle of each step were represented by 2 bits, respectively. The total number of bits in each chromosome is therefore \((2 + 2) \times N = 4N\), where \(N\) represents the total steps of a random walk \((N = 6\) for this dissertation unless specified). Roulette-Wheel selection, single-point crossover, random mutation and elitist strategy were deployed in the evolution process. Other parameters involved in the algorithm were as follows: population size = 100, maximum generation = 30, crossover probability = 0.9, and mutation probability = 0.02.

The in-house BoR antenna solver developed in Chapter 4 is used to perform EM analyses called by the INTERFACE program, as shown in Figure 5.1.

5.3 Cost Function Construction

Multi-objective optimization (MOO) is defined as the process of optimizing a collection of objective functions systematically and simultaneously [96]. Surveys have been conducted to provide comprehensive coverage of popular methods along with their general applicability to engineering problems [96–98]. A common conclusion is that no single approach is superior. The selection of a specific method typically depends on the type and amount of input information, the decision-maker’s preference, the solution requirements, and the ease of implementing associated algorithms.

For the remainder of this dissertation, the weighted global criterion method [99, 100] will be adapted to construct the cost function. As a method with a priori
articulation of preferences, it allows the user to specify the relative importance of different objectives. Moreover, as discussed below, the idea of a *compromise solution*, which entails minimizing the difference between a potential optimal point and a *utopia point* [101], provides a clear-cut physical interpretation to the optimization process.

According to [101], a utopia point $F^o (\in \mathbb{Z}^k)$ is defined as “a point that for each $i = 1, 2 \cdots, k, F^o_i = \text{minimum}\{F_i(x)| x \in X\}.”$ $x$ is a vector of design variables (point in the design space); $X$ is the feasible design space; $F$ is a vector of objective functions (point in the criterion space); $Z$ is the feasible criterion space; $k$ is the number of objective functions. As the utopia point $F^o$ is unattainable, optimization leads to a solution point as close as possible to $F^o$. A common way to realize the

Figure 5.5: Illustration of the weighted global criterion method. The goal is to minimize the distance $N(x)$ between the solution point $F(x)$ and the utopia point $F^o$.
term *close* is to minimize the Euclidean distance \(N(x)\) given by

\[
N(x) = |F(x) - F^o| = \left\{ \sum_{i=1}^{k} [F_i(x) - F^o_i]^2 \right\}^{\frac{1}{2}},
\]  

which is illustrated in Figure 5.5. In fact, the term *distance* can be defined in a broader sense [102]. One of the most general *utility functions*, which combines all objective functions to form a single function and models a decision-maker’s preference, is expressed in the following form [103]:

\[
U = \left\{ \sum_{i=1}^{k} w_i [F_i(x) - F^o_i]^p \right\}^{\frac{1}{p}},
\]  

where \(w\) is a vector of weights satisfying \(\sum_{i=1}^{k} w_i = 1\) and \(w_i > 0\). According to [103] and [104], (5.3) is sufficient for *Pareto optimality* [105] as long as \(w_i > 0\). The Pareto optimality is a situation in which the output of one objective function cannot be reduced without increasing the output of another objective function.

However, in practice, if different objective functions in (5.3) have different units, the physical meaning of *distance* becomes vague, leading to an ill-represented *closeness*. Besides, under such conditions, it might be difficult for decision-makers to come up with a proper weighting strategy since the outputs of individual object functions are inherently incomparable. It is therefore advantageous to transform the original objective functions to dimensionless ones. One of widely used transformation functions is given by [106, 107]

\[
F_i^{\text{trans}}(x) = \frac{F_i(x)}{F^o_i},
\]  

in which \(F^o_i > 0\) is assumed. Accordingly, (5.3) is modified as follows

\[
U = \left\{ \sum_{i=1}^{k} w_i [F_i^{\text{trans}}(x) - 1]^p \right\}^{\frac{1}{p}}.
\]  

65
One can interpret (5.5) as components of a distance function that represents the relative deviation of a solution point to the utopia point in the criterion space. It will be used to construct the single cost function for the following design optimizations.

5.4 Design Examples

In this section, two design examples subject to different radiation objectives are provided to illustrate the optimization procedure developed in the preceding sections. The interpretation of resulting configurations is important to understand the underlying antenna operating principles.

5.4.1 24” × 6” UWB Monopole for 40 MHz to 2 GHz Operation

In practice, a structure that can operate down to low VHF frequencies while maintaining a good performance at higher frequencies is desirable. The antenna of interest is a 24” wide and 6” tall BoR monopole on an infinite ground plane. A figure of merit to characterize the low frequency performance is the onset of its -15 dBi gain. Obviously, it is favorable to have this onset to occur at a lower frequency. For higher frequencies, a moderate realized gain (> 0 dBi) over the horizon (θ = 90°) with minimum oscillations across the band is preferred. The voltage standing wave ratio (VSWR) should also be minimized over a large bandwidth for impedance matching.

The above three objectives—low frequency gain ($F_1$), impedance matching ($F_2$), and uniform directivity ($F_3$)—are given by:

\[ F_1 = f@-15 \text{ dBi}, \]
\[ F_2 = \frac{1}{N_f} \sum_{f}^{N_f} \text{VSWR}(f) \quad f = 200, 300, \ldots, 2000 \text{ MHz}, \]

66
where parameter $N_f$ refers to the frequency steps, and $D(f)$ is the directivity as a function of frequency. Equation (5.6) is self-explanatory. Equation (5.7) represents the average VSWR across the frequency band from 200 MHz to 2 GHz. Below 200 MHz performance is dominated by the size constraints instead of geometrical details. Therefore, the VSWR at these frequencies are always too large to be considered. It is remarked that an alternative of imposing constraints on continuous bandwidth is to minimize the maximum VSWR within the operational frequency range. Equation (5.8) aims to minimize the standard deviation of the directivity over the horizon ($\theta = 90^\circ$). Low frequency radiation is primarily related to edge diffractions at the aperture end, the calculating bandwidth was therefore set from 600 MHz up to 2 GHz.

As a first step, separate optimizations for each objective function were carried out to find out the utopia point. The resulting geometries along with the minimum outputs $F_i^o (i = 1, 2, 3)$ are shown in Figure 5.6 and Table 5.1. For the first criterion

![Figure 5.6](image)

Figure 5.6: Side view of the optimum monopole antenna profiles (24” wide and 6” tall) associated with objective function $F_1$, $F_2$, and $F_3$, respectively.
Table 5.1: Utopia point: minimum outputs for the objective function $F_1$, $F_2$, and $F_3$.

<table>
<thead>
<tr>
<th>$F_1^o$</th>
<th>$F_2^o$</th>
<th>$F_3^o$</th>
</tr>
</thead>
<tbody>
<tr>
<td>39.7 MHz</td>
<td>1.393</td>
<td>0.321 dB</td>
</tr>
</tbody>
</table>

$F_1$, it is observed that optimization managed to select a configuration (optimum antenna #1) similar to the top-loaded monopole, which is miniaturized by introducing a capacitive top plate [108]. The optimum profile associated with the second objective function $F_2$ is quite close to that of a conical antenna, except for a narrower cone angle at the feed region. The input impedance of a monocone is approximated by [8]

$$Z_{in} = 60 \ln(\cot \frac{\alpha}{4}),$$  

where $\alpha$ refers to the bottom cone angle. Thus, to achieve 50 $\Omega$, the opening angle needs to be $94^\circ$, defined by the first three line segments of the optimum antenna #2.

When the uniform directivity ($F_3$) criterion was applied, GA yielded an aperture that gradually tapers its surface away from the feed, a common feature shared by Vivadi-type and tapered-slot antennas (TSAs) [62]. This smooth curvature (see optimum antenna #3) facilitates high-frequency radiating out prior to reaching the aperture edge, leading to relatively small horizontal gain variations across the band.

A global weighted single dimensionless cost function that combines (5.4)–(5.8) along with $F_i^o$ given in Table 5.1 was then constructed as follows

$$\text{COST} = \sqrt{w_1\left(\frac{F_1}{F_1^o} - 1\right)^2 + w_2\left(\frac{F_2}{F_2^o} - 1\right)^2 + w_3\left(\frac{F_3}{F_3^o} - 1\right)^2}. \quad (5.10)$$

Note that $k = 3$ and $p = 2$ were used in (5.4) and (5.5). After some trial and error, it was found that $w_1 = 0.38$, $w_2 = 0.6$, and $w_3 = 0.02$ are good choices, as they emphasize low frequency gain and impedance bandwidth for the final design.
Figure 5.7: (a) Side view and (b) 3-D view of the final optimized 24” wide and 6” tall BoR monopole antenna.

The resulting optimum profile via this multi-objective optimization is given in Figure 5.7. The realized gain in the horizontal plane ($\theta = 90^\circ$) along with the input impedance (on the Smith Chart) of this final design are shown in Figure 5.8 and Figure 5.9, respectively.

Figure 5.8: Horizontal realized gain ($\theta = 90^\circ$) of the 24” × 6” final design via shape optimization (inset figure displays the low frequency performance).
The gain level is, by and large, stable, reaching +5 dBi at frequencies above 800 MHz. More importantly, the onset frequency point at -15 dBi gain arrives at 41 MHz (see inset in Figure 5.8), i.e. quite close to the theoretical limit (32 MHz) predicted by [10]. The electrical size of this antenna is only $0.08\lambda$ (24”) in width and $0.02\lambda$ (6”) in height at 41 MHz. As frequency increases, the input impedance converges to the center of the Smith Chart (50 $\Omega$), a desired feature.

The “distance” between the compromise solution point and the utopia point can be visualized and quantified in the following manner. Figure 5.10–Figure 5.12 display the performance comparison between the individual optimum antennas and the final design. One can easily relate the objective function $F_i$ to the corresponding geometric feature. For instance, as can be seen from Figure 5.10, in terms of the maximum possible gain at low frequencies, the final design is fairly “close” to the optimum...
Figure 5.10: Low frequency gain comparison between the final design and optimum antenna #1 (inset figure displays the associated antenna configurations).

Figure 5.11: VSWR comparison between the final design and optimum antenna #2 (inset figure displays the associated antenna configurations).
Figure 5.12: Directivity ($\theta = 90^\circ$) comparison between the final design and optimum antenna #3 (inset figure displays the associated antenna configurations).

antenna #1. This is largely due to the “top-hat” defined by the outer two segments. Also, the bottom opening angle is critical for impedance matching—an evident fact shown in Figure 5.11. The introduction of a new local diffraction point (i.e. the 3rd control vertex on the final design profile) might account for the directivity spike at 1200 MHz indicated in Figure 5.12.

\[
\begin{array}{|c|c|c|c|}
\hline
i & \mathcal{F}_i & i = 1 & i = 2 & i = 3 \\
\hline
\mathcal{F}_{i_{\text{p}}} & 41.2 \text{ MHz} & 1.646 & 0.698 \text{ dB} \\
\mathcal{F}_{i_{\text{p,trans}}} - 1 & 0.038 & 0.182 & 0.897 \\
\hline
\end{array}
\]

Table 5.2: Objective function value $\mathcal{F}_i$ of the final optimized design and the associated fraction deviation ($\mathcal{F}_{i_{\text{p,trans}}} - 1$) from the utopia point.
Table 5.2 reflects the *closeness* of the optimal design point (P) to the utopia point (O) in the criterion (objective) space. Specifically, \( F_{i}^{\text{p,trans}} - 1 \) \( (F_{i}^{\text{p,trans}} \text{ is given by (5.4))} \) indicates the relative deviation from the utopia point. As mentioned earlier, the low frequency gain \( (F_1) \) and impedance bandwidth \( (F_2) \) are two key design objectives. Consequently, the ultimate design point is only 3.8% and 18.2% away from the utopia point in terms of \( F_1 \) and \( F_2 \), as shown in Table 5.2. Figure 5.13 depicts the lowest cost of each generation (viz. the progress toward convergence) implying that the GA entered a steady state after approximate 10 generations. The final design is the best

![Figure 5.13: Costs associated with the best individual of each generation through the GA optimization.](image)

<table>
<thead>
<tr>
<th>( i )</th>
<th>0</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
</tr>
</thead>
<tbody>
<tr>
<td>( x_i ) (in)</td>
<td>0.00</td>
<td>1.60</td>
<td>1.96</td>
<td>5.00</td>
<td>5.84</td>
<td>9.42</td>
<td>12.00</td>
</tr>
<tr>
<td>( y_i ) (in)</td>
<td>0.00</td>
<td>1.39</td>
<td>1.71</td>
<td>3.10</td>
<td>4.74</td>
<td>5.47</td>
<td>6.00</td>
</tr>
</tbody>
</table>

Table 5.3: Coordinates of the control vertices defining the final optimal profile.
individual from the 15th generation. Control vertices that define this optimal profile are given in Table 5.3.

Lastly, the electrical performance of the shape-optimized design (in Figure 5.7) is compared with that of the 11-ellipse IHA presented in [67] (in Figure 4.1). Both designs are 24” in aperture width and 6” in height. As can be seen in Figure 5.14, design optimizations carried out in this subsection yield a significant performance improvement. Specifically, the 11-ellipse IHA reaches VSWR < 3 from 380 MHz up to 2 GHz, while the optimized aperture achieves the same matching condition at 210 MHz and above. In addition to the enhanced bandwidth, the in-band reflection are also much less for the latter design. Meanwhile, the optimized antenna exhibits a rather uniform horizontal gain at higher frequencies (e.g. 1 GHz–2 GHz), which translates into less directive patterns. This indeed results from the minimization of the standard deviation of antenna directivity across the band of interest. It is

![Figure 5.14: Performance comparison of the 24” wide and 6” tall shape-optimized antenna vs. 11-ellipse IHA: (a) VSWR and (b) realized gain at horizon ($\theta = 90^\circ$).](image-url)
remarked that using a spline representation [44, 77, 89] given control vertices might lead to further improvements.

5.4.2 VHF Antenna Miniaturization

As already noted, the onset frequency at -15 dBi gain is a figure of merit to characterize the low frequency performance. In this subsection, a group of antennas with different dimensions are considered. The goal of this study is to provide reference charts that reflect the trade-offs between BoR antenna size and its maximum achievable gain at low VHF band. The RGB method is used for shape optimization. The optimum performance subject to a given aperture width and height is compared with that predicted by the small antenna theory [10] as well as previously developed IHAs.

The random walk (RW) scheme described in section 4.1 only allows the profile to grow monotonically with respect to the feed. This effort simplifies the fabrication process but limits the antenna shape that can be represented. For instance, a top-loaded monopole (i.e. a conventional wire monopole topped by a disc) cannot be generated using the existing algorithm, because the line segments forming the profile are prohibited to be in parallel with the ground plane. To overcome this restriction, the angle choices ($\alpha_j$) of the RW scheme were expanded so as to include direction pointing horizontally ($\alpha_j = 0$) and downward ($\alpha_j < 0$). The array of angles was modified to be

$$\alpha_j \in \{-30^\circ, -15^\circ, 0^\circ, 15^\circ, 30^\circ, 45^\circ, 60^\circ, 75^\circ\},$$

while the length choices ($r_j$) were maintained the same as before. The modification on the RW gives rise to problems for the GA and BoR solver. The profile might be invalid when the aperture touches the ground plane or even goes below it. Thus, an
extra step was added in the optimization loop to check and generate geometries only with positive coordinates.

The cost function for the shape optimization is given by (5.6). Figure 5.15 and Figure 5.16 display the onset frequency points at -15 dBi gain for different aperture widths and heights. As seen, the increase of the aperture size leads to a lower operating frequency with regard to a given gain requirement. On the one hand, the theoretical study shows that a 6” tall antenna is able to achieve -15 dBi gain around 50 MHz and 26.5 MHz as the aperture width is 12” (0.051\(\lambda\)) and 30” (0.067\(\lambda\)), respectively. Because the modified shape representation scheme still cannot generate a completely arbitrary profile given size constraints, the design points resulting from the optimization are not able to reach the theoretical limit. However, comparing to the similar study conducted on the 6” tall double-ellipse IHAs (see section 3.3), the

![Figure 5.15: The onset frequency point at -15 dBi gain of a 6” tall body-of-revolution monopole antenna over infinite ground plane for different aperture widths.](image-url)
Figure 5.16: The onset frequency point at -15 dBi gain of a 24” wide body-of-revolution monopole antenna over infinite ground plane for different heights.

optimized design curve approaches closer to the limit, a sought-after feature. On the other hand, when the aperture width is kept at 24”, the variation of height (2”–8”) plays a significant role on the antenna low frequency gain. Specifically, as the width to height ratio increases, the design curve produced by the shape optimization starts deviating from the theoretical limit. This is largely because the high-order diffractions occurred at the aperture edge cancel each other out when the spacing between the aperture and ground plane is too narrow, resulting in poor net radiation. Due to the fact that amplitude of the field along the diffracted ray is inversely proportional to the square root of the distance between the observation point and diffraction point [111], with the increase of height (decrease of the width/height ratio), high-order diffraction
interferences are much less pronounced—the first-order diffraction dominates the radiation mechanism. As a result, a linear behavior is observed in Figure 5.16 for the optimized design curve (4”–8”).

It is also worth mentioning that the theoretical limit derived in this chapter is based upon Chu’s work, which is mostly accurate for spherical geometries (width/height = 2). Different performance bounds might hold for non-spherical geometries [112], such as ellipsoidal antennas with width to height ratio in excess of 10. Thereby, it would be interesting to examine two extremities: a 12” × 6” configuration (width/height = 2) and a 24” × 2” configuration (width/height = 12). As shown in Figure 5.17, the low width/height ratio design makes a better use of its volume by introducing a

Figure 5.17: Side view of (a) 12” wide and 6” tall and (b) 24” wide and 2” tall BoR monopole antenna achieving lowest frequency at -15 dBi realized gain via shape optimization.
zigzag arm as inductive loading. This observation agrees with Chu’s statement that highest directivity can be achieved when the radiator occupies maximum volume of the sphere that encloses it. On the contrary, the high width/height ratio design is in the form of a disc-loaded monopole, which maximizes the capacitive loading effect. The meander line feature is not seen in this design. Based on above findings, one can realize that, in addition to width and height, width to height ratio is also an important parameter that determines the radiation property of electrically small body-of-revolution monopole antennas. Given an ultra low profile (e.g. 24” × 2” or λ/16 × λ/200 at 30 MHz), [113] shows that strategically placed ferrite loading increases realized gain up to the theoretical limit. However, the resulting design operates in a loop antenna mode, leading to inconsistent beam pattern change as the frequency varies.

5.5 Post-Optimization Improvement

The optimized antenna found in section 4.4.1 can be further improved via resistive loading. By introducing additional material losses, the antenna VSWR can be reduced at the expense of efficiency [33, 109, 110]. Due to the fundamental limits on small antennas, the current design exhibits a significant amount of mismatch loss particularly at the VHF band (e.g. 30 MHz–200 MHz). For certain high power applications, the reflected energy might destroy the transmitter if no protection circuitry exists. In addition to the poor matching at lower frequencies, the irregular pattern at higher frequencies, which largely stems from the interference among diffracted field
components at the aperture edge, is also a major concern. For example, the horizontal gain “drop” at 500 MHz and “spike” at 1200 MHz in Figure 5.8 is because of pattern variations.

It is thus provoked to lower the VSWR and smooth out patterns concurrently. To this end, the metal surface defined by the last two segments in Figure 5.18(a) is considered to be replaced by an impedance sheet, minimizing reflections back into the receiver and reducing undesired edge diffractions. A controlled experiment was then conducted to verify this hypothesis. Specifically, as seen from Figure 5.18(b), a portion of metallic ream was removed from the initial design. An R-card surface with sheet resistance of 100 Ω/□ (Figure 5.18(c)) and 500 Ω/□ (Figure 5.18(d)) was mounted on the truncated structure to maintain an identical size as that of the

![Graphs showing antenna performance](image)

Figure 5.18: Side view of the antennas considered for a resistive loading treatment: (a) the shape-optimized antenna found in section 4.4.1, (b) a smaller PEC aperture, (c) a smaller PEC aperture incorporated with an R-card having a sheet impedance of 100 Ω/□, and (d) a smaller PEC aperture incorporated with an R-card having a sheet impedance of 500 Ω/□.
original PEC one (Figure 5.18(a)). For ease of fabrication, a 1” PEC strip ring was kept (the last PEC segment shown in Figure 5.18(b)–(d)) to overlap with the R-card. This part also serves to guide wave propagating along the resistive surface.

Low frequency (30 MHz–200 MHz) performances of the subject antennas are presented in Figure 5.19. Impedance matching is improved when resistive loadings are applied. As a result of additional material losses, the onset frequency at -15 dBi gain point shifts from 41 MHz (untreated) to 44 MHz (500 Ω/□ R-card). One can predict that, as the sheet resistance increases, the realized gain curve of the resistively-loaded antenna will approach that of the smaller PEC aperture (Figure 5.18(b)). This is because an extremely thin sheet with infinite resistance is essentially equivalent to a perfect insulator (air).

A far more significant performance improvement is observed at high frequencies (200 MHz–2 GHz). In particular, the antenna incorporated with a 500 Ω/□ impedance sheet delivers a rather smooth gain along the horizon, as displayed in

![Figure 5.19](a) Reflection coefficient and (b) realized gain in the horizontal plane (θ = 90°) of the untreated and resistively-loaded antennas (30 MHz–200 MHz).

81
Figure 5.20: (a) Reflection coefficient and (b) realized gain in the horizontal plane ($\theta = 90^\circ$) of the untreated and resistively-loaded antennas (30 MHz–2 GHz).

Figure 5.20(b). The gain drop at 500 MHz is eliminated thanks to the successful pattern control (see Figure 5.21(a)). The strong local diffracted currents along the large aperture edge are effectively suppressed by the resistive surface, equivalently shrinking the radiation aperture to a smaller one. In addition, the R-card kills the energy reflected back to the feed, resulting in a better matching condition as can be seen in Figure 5.20(a). Note that the smaller PEC aperture itself radiates well at frequencies above 500 MHz, which explains why a moderate gain ($\sim 5$ dBi) is still achievable, even though lossy material has been introduced into the structure. Furthermore, side lobes of the resistively-loaded antenna are far less pronounced in comparison to the untreated design, a critically sought-after feature.

This study shows that the use of resistive loading leads to: (a) lower VSWR; (b) smoother gain from 200 MHz to 2 GHz without much compromise in low frequencies; (c) more uniform monopole-like pattern. In practice, the location and sheet resistance of the R-card should be selected strategically to accommodate different needs at the
low frequency and high frequency ends, respectively. Tapered R-card achieves a gradual change from perfect conductor to impedance surface and thus might further smooth radiation patterns [33, 114].

5.6 Measurements and Validation

In this section, measurements are presented to validate the shape optimized design and resistive loading treatment, respectively. The 24” wide and 6” tall aluminum component was fabricated using standard metal spinning technology. It was then
mounted on a 26” diameter ground plane (see Figure 5.22(a)) and fed by a 50 Ω coaxial cable. The resistively-loaded prototype was subsequently constructed by replacing the top aluminum surface with an R-card section. Due to the limited resistive films available in the lab, the sheet impedance was chosen to be 120 Ω/□. Insulating foam sealant, verified to have little impact on the band of interest, was sprayed between the antenna and ground plane, as shown in Figure 5.22(b). It should be noted that all previous analyses were based on an infinite ground plane. Performance degradation is expected when measured with a finite ground plane.

Measurements were then conducted (see Figure 5.23) and compared with simulation results for the finite ground case. The VSWR and realized gain in the horizontal
plane ($\theta = 90^\circ$) of these two prototypes are depicted in Figure 5.24 (a) and (b). A reasonably good agreement is observed between measurements and calculations. The introduction of R-card reduces VSWR, resulting in an 8:1 bandwidth ($\text{VSWR} < 3$ from 260 MHz up to at least 2 GHz). Meanwhile, the selection of sheet resistance (120 $\Omega/\square$) was demonstrated to yield an insignificant gain reduction in comparison to that of the untreated case.

Figure 5.25 presents the simulated and measured $E$-plane radiation patterns for the two prototypes at different frequencies. The main beam direction could not be maintained at $\theta = 90^\circ$ because of the finite ground plane. However, of importance is that pattern stability was enhanced via resistive loading. Specifically, at 2 GHz, the main beam is broader ($30^\circ$–$120^\circ$), while the side lobes are less pronounced. It is noted that differences between simulation and measurement at some angles are due to fabrication tolerances. Nevertheless, the measurements validate the efficacy of
Figure 5.24: Measured and simulated (a) VSWR and (b) realized gain at horizon ($\theta = 90^\circ$) of the 24” wide and 6” tall aluminum and resistively-loaded prototype shown in Figure 5.22.
Figure 5.25: Measured and simulated radiation patterns (E-plane) of the 24” wide and 6” tall aluminum and resistively-loaded prototype shown in Figure 5.22: (a) 1000 MHz, (b) 1500 MHz, (c) 1800 MHz, and (d) 2000 MHz.

using resistive loading to improve antenna impedance matching and reduce pattern variations.
5.7 Summary

In this chapter, a systematic design approach was presented to optimize the shape of low-profile UWB body-of-revolution monopole antennas. The proposed RGB method combines a random walk scheme with a genetic algorithm and a BoR-MoM solver. The cost function construction was based on the weighted global criterion method, aimed at converging to a solution point as close as possible to the utopia point that achieves the optimal performance for all design objectives simultaneously. Two examples were given to elaborate the developed optimization methodology and cost function selection strategy. An optimized 24” wide and 6” tall aperture reaches -15 dBi gain at 41 MHz on an infinite ground plane, as it is only 0.08λ in width and 0.02λ in height. The same antenna also provides a VSWR < 2 bandwidth from 280 MHz up to at least 2 GHz, while maintaining a realized gain around +5 dBi with moderate oscillations across the band of interest. In comparison to the 11-ellipse IHA of the same size, this optimized design offers an enhanced bandwidth as well as more stable horizontal gain. The second example generated reference charts that relate BoR antenna size with its maximum achievable gain at low VHF band. Design curves from shape optimization lie closer to the theoretical limit in comparison with the non-optimized IHA cases. A resistive loading treatment was further applied at the monopole ream to improve matching at low frequencies and smooth pattern at high frequencies. The optimized designs were validated via measurements. This shape optimization approach can be readily applied to a variety of body-of-revolution monopole antennas.
Chapter 6: Conclusions

6.1 Summary

Low-profile ultra-wideband (UWB) monopole antennas are attractive for aircraft and ground vehicle imaging/voice communication systems. The continued need for UWB and small size antennas implies significant challenges in radio frequency aperture design. Not surprisingly, this topic has been of strong interest in the research community over the last decade.

This dissertation focuses on designing the outer surface of a body-of-revolution (BoR) monopole antenna subject to given size constraints and performance requirements. To this end, two types of antennas were developed following different design philosophies. By manually creating multiple diffraction points along the profile, the inverted-hat antenna (IHA) managed to maintain a similar scattering geometry in terms of wavelength, leading to UWB operation. It was demonstrated that controllable input impedance and uniform radiation patterns can be accomplished by this design. For the latter property, increasing diffraction from the elliptical curvatures was exploited to achieve a broader radiation pattern at higher frequencies. This concept was elaborated and verified in Chapter 3. Alternatively, the outer surface of the antenna can be constructed following a design optimization approach (Chapter
4–Chapter 5), which incorporated a random walk for antenna outer profile generation, a genetic algorithm for optimization, and a BoR moment method analysis—all combined in a single design package. Of importance is the cost function, which represents the difference between a potential optimal point and a utopia point within the solution space. The utopia point combines separate optimal designs associated with all key objectives, offering a deep insight into antenna operating principles. Example designs were included to demonstrate performance improvements due to antenna shape optimization. In comparison to the original IHAs of the same size, the optimized designs were shown to deliver larger bandwidth, higher gain at low VHF band, and more stable horizontal gain at higher frequencies.

It is worth noting that the IHA design has a clear physical reasoning owing to the fact that it is based on diffraction theory principles. However, the resulting configuration is not guaranteed to be optimal since geometrical parameters are selected manually according to the designer’s intention. If the objectives are manifold, it is impractical to come up with a reasonable solution simply by means of trial and errors. On the contrary, the design optimization approach is heavily dependent upon the cost function selection strategy once the program has been set up correctly. Little prior knowledge in the theory of antennas is needed, making it suitable for exploring a variety of unknown topics. The interpretation of geometries yielded from shape optimization is important as it might suggest non-intuitive physics that could be used to guide future designs.

The contents of each chapter are summarized as follows.

A survey was conducted in Chapter 2 on the VHF/UHF UWB antennas according to their radiation pattern shape. They were categorized into either dipoles/monopoles
or spirals. This historical review suggests that a great interest exists in developing small UWB antennas to replace conventional narrow-band protruding radiators. In particular, there is a strong need to develop a systematic approach for the shape design of broadband monopole antennas that offer excellent vertically polarized (VP) radiation over the horizon while maintaining a low-profile.

Chapter 3 introduced the concept of frequency-scaled inverted-hat antenna, of which the outer surface is composed of multiple ellipses that follow the growth rate of an exponential spiral. A nearly frequency-independent behavior was observed for a moderate number of elliptical segments used to construct the IHA. Impedance control and uniform radiation patterns were accomplished by employing different outer surface growth profiles. Several prototypes were fabricated and tested to validate the concept. Specifically, a 15” (λ/10) wide and 6” (λ/25) tall double-ellipse IHA achieved a gain of -15 dBi at 75 MHz when mounted on a 21” finite ground plane. A seven-ellipse design of the same size was shown to deliver a 10:1 impedance bandwidth along with satisfactory radiation performances.

Chapter 4 presented a body-of-revolution moment method analysis for an UWB dome-dipole antenna. This method reduces a 3-D analysis to a simple 1-D one, leading to a dramatic solution time reduction and thus making it suitable for insertion in an optimization loop. To verify the performance predicted by the BoR analysis, a 24” wide and 20” tall dome-dipole prototype was fabricated and demonstrated to deliver an operational bandwidth from 180 MHz to at least 2 GHz. Excellent agreements between measurements and calculations were observed over a 100:1 bandwidth. This agreement validated the BoR analysis, delivering confidence on using it in optimizing the outer surface of low-profile UWB monopole antennas.
In Chapter 5, a systematic approach named the RGB method was proposed for the design and optimization of profiled UWB monopole antennas that satisfy specified radiation objectives and size constraints. A random walk scheme, which is able to achieve a fairly diverse geometric space, was used for the shape representation. The genetic algorithm was chosen as the optimizer, and the BoR analysis developed in Chapter 4 was deployed to rapidly carry out EM evaluations. The cost function construction is based upon the weighted global criterion method, which entails minimizing the difference between a potential optimal point and a utopia point within the objective space. Utilizing the proposed shape optimization method, two examples were presented and demonstrated to deliver improved performance in comparison to previously developed IHAs of the same size. A resistive loading treatment was further applied at the monopole ream to improve matching at low frequencies and smooth pattern at high frequencies. Measurements agreed well with simulations.

6.2 Future Work

6.2.1 Development of Advanced Shape Optimization Method

The RGB method proposed in this dissertation is demonstrated to be a powerful tool in antenna shape optimization. Nevertheless, to extend the scope of its application, extra effort has to be made in the following aspects.

The random walk scheme for constructing the antenna profile can be modified for a broader sense of shape representation. Specifically, it is not necessary to set the point \((w/2, h)\), where \(w\) is the maximum aperture width and \(h\) refers to the antenna height, as the unique destination for the random walk. An arbitrary path that satisfies the size constraint should be allowed as long as there is no intersection between any
two segments. This freedom in geometry generation might result in some interesting profiles. For example, the teardrop shape, which has no sharp diffraction points, is likely to maintain a stable radiation pattern across a large bandwidth. In addition, the geometric detail becomes increasingly important for high frequency operation. In this context, not only the control vertices that define the profile but also the smoothness of the curve is crucial. A rational curve [75, 95] or a spline representation [44, 77, 89] might be a good alternative. For a multi-objective optimization problem, instead of exclusively looking for a single optimum design with explicitly defined constraints, one may also consider generating a pool of candidates that lie on a surface Pareto front [87, 88]. This strategy provides antenna designers with optimal alternatives since an absolutely perfect solution rarely exists in the real world. Also, the BoR-MoM analysis should not be limited to PEC structures only. Inclusion of magnetic surface currents along the BoR surface allows for applying this algorithm to lossy dielectric apertures [69, 73, 74, 77], which will facilitate users to optimize sheet impedance values when the resistive loading technique is employed.

6.2.2 Integration of Spiral and UWB Monopole for Pattern Diversity

As summarized in Chapter 2, an UWB antenna operating at VHF and UHF band that offers either monopole-like pattern or patch-like pattern has been continuously studied over the past few decades. It is natural that consideration be given to combining these two RF apertures into a single antenna system for pattern diversity. That is, a spiral provides circularly polarized (CP) overhead coverage (patch-like pattern) while a wideband monopole offers omni-directional VP radiation (monopole-like pattern) down to lower VHF band. Configurations of a demo hybrid design (backed by a
22" ground plane) along with individual elements are displayed in Figure 6.1. A preliminary test suggests that the 15” × 6” double-ellipse IHA and the volumetric spiral antenna of 12” diameter functioned normally when only one of the two radiators was active (see Figure 6.2). Further challenges for constructing this hybrid system are the design of a dual-connector feeding scheme to excite each antenna simultaneously and mitigating the interference/mutual couplings.
Figure 6.1: (a) 15” wide and 6” tall double-ellipse inverted-hat antenna (IHA), (b) 12” spiral antenna, and (c) the hybrid design placed on a 22” ground plane.

Figure 6.2: Preliminary gain measurement results of the monopole-spiral hybrid design in Figure 6.1: (a) comparison of the measured horizontal gain (with and without spiral on top of the double-ellipse IHA) and (b) boresight gain of the spiral (IHA was terminated).
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Appendix A: Body-of-Revolution Monopole Antenna Shape Optimization Code (BoRMASOC) User Guide

A.1 Program Structure

The “Body-of-Revolution Monopole Antenna Shape Optimization Code” (BoR-MASOC) is an in-house MATLAB program for optimizing the outer surface of a body-of-revolution (BoR) monopole antenna (on an infinite ground plane) given size constraints and radiation objectives. This code package consists of 9 scripts and/or functions, which are shown in Figure A.1. To perform a single optimization run, the user should follow the 3 steps below:

1. Initialization (Input.m)

2. Optimization (Interface.m, bin2rv.m, BoRCal.m, BoR.m, CostFun.m, and BGA.m)

3. Post-processing (PfmPlot.m and PfmAnim.m)

The user can manually execute the scripts labeled by black blocks, whereas the functions in red blocks are only accessible by the upper level scripts or functions.
Figure A.1: Program structure of the Body-of-Revolution Monopole Antenna Shape Optimization Code (BoRMASOC).
A.2 Scripts and Functions

A.2.1 Input.m

This script provides the user with an interface to set up program parameters associated with design specifications and the RGB method addressed in Chapter 5.

- **Antenna Specifications**
  - $D$: aperture width (unit: inch)
  - $H$: height (unit: inch)
  - $freq$: frequency range (unit: Hz)

- **Random Walk**
  - $nBitR$: number of bits representing the length choices
  - $r$: length choices
  - $alpha$: angle choices
  - $nStep$: number of steps during the random walk

- **Genetic Algorithm**
  - $nPop$: population size
  - $nMaxGen$: maximum number of generations

- **Miscellaneous**
  - $flagGenInfo$: indicator for the storage of the up-to-date simulation result.
    
    If it is 1, the program saves all existing calculation results to a subdirectory
“History”. This operation repeats every generation. The user is therefore able to access the latest result even if the program is still running.

– ProjectName: A name root shared by the output files.

All defined variables and their values in current workspace are saved in a binary data file XXXX_input.mat, where XXXX is assigned in ProjectName.

A.2.2 Interface.m

As the kernel of the entire optimization program, this script interconnects the random walk scheme, the genetic algorithm (GA), and the BoR-MoM solver. It also makes a record of all calculation results for the sake of post-processing. In addition, this piece of code checks the geometry resulting from the GA to guarantee positive valued control vertices. Other default setups include:

• poolsize for parallel computation: 2

• convergence criterion: exit the optimization loop if the relative absolute cost variation across the latest 5 generations is less than 1%

All defined variables and their values in current workspace are saved in a binary data file XXXX_output.mat, where XXXX is assigned in ProjectName.

A.2.3 bin2rv.m

This function implements the random walk scheme by translating Gray-coded binary strings carrying length and angle choices into real-valued control vertices that define the antenna 2-D profile.

Inputs:
- `chrom`: chromosome (binary string)
- `D`: aperture width (unit: inch)
- `H`: height (unit: inch)
- `nStep`: number of steps during the random walk
- `r`: length choices
- `alpha`: angle choices

**Outputs:**
- `rho`: `x`-coordinates of the control vertices (unit: inch)
- `z`: `y`-coordinates of the control vertices (unit: inch)

### A.2.4 BoRCal.m

This function performs frequency-sweep analysis on a BoR monopole antenna by calling the subroutine `BoR.m` and calculates the associated costs by invoking the subroutine `CostFun.m`.

**Inputs:**
- `rho_in`: `x`-coordinates of the input geometry (unit: inch)
- `z_in`: `y`-coordinates of the input geometry (unit: inch)
- `freq`: frequency range (unit: Hz)

**Outputs:**
• \textit{Dir.}_90\text{deg}.dB: directivity (unit: dB) in the horizontal plane (\(\theta = 90^\circ\)) at each frequency

• \textit{Zin}: complex input impedance (unit: \(\Omega\)) at each frequency

• \textit{Rgain\_patt}.dB: realized gain (unit: dB) within the angle sector (\(\theta = 0^\circ\text{--}90^\circ\)) at each frequency

• \textit{cost}: total cost

• \(P\): individual cost items (the default number of cost items is 5)

## A.2.5 BoR.m

This function implements the BoR-MoM analysis developed in Chapter 4. The input geometry must be of a dipole configuration and made of PEC only. Thanks to the rotational symmetry and image theory, control vertices in the first quadrant are sufficient to describe the antenna profile completely.

**Inputs:**

• \textit{freq}: the single operating frequency (unit: Hz)

• \textit{rho}: \(x\)-coordinates of the input geometry (unit: m)

• \textit{z}: \(y\)-coordinates of the input geometry (unit: m)

**Outputs:**

• \textit{Zin}: complex input impedance (unit: \(\Omega\))

• \textit{E\_ff}: \(E_\theta\) in the far field (ignoring the term \(\frac{e^{-jkr}}{r}\)) (unit: V/m)

• \textit{I}: tangential current density along the profile (unit: A/m)
• $N_{tri}$: total number of basis functions

### A.2.6 CostFun.m

This is a highly customized function that defines the cost function(s). The user is suggested to strategically modify the passed variables as well as associated operations to accommodate his or her needs. By default, the inputs and outputs are given as follows:

**Inputs:**

- $freq$: frequency range (unit: Hz)
- VSWR: voltage standing wave ratio (VSWR) at each frequency
- $Dir_{90deg\,dB}$: directivity (unit: dB) in the horizontal plane ($\theta = 90^\circ$) at each frequency
- $Rgain_{90deg\,dB}$: realized gain (unit: dB) in the horizontal plane ($\theta = 90^\circ$) at each frequency
- $Rgain_{patt\,dB}$: realized gain (unit: dB) within the angle sector ($\theta = 0^\circ$–$90^\circ$) at each frequency

**Outputs:**

- $cost$: total cost
- $P_{freq}$: onset frequency point at -15 dBi realized gain (unit: MHz)
- $P_{VSWR}$: mean of the VSWR across the band of interest
• $P_{\text{StdDir}}$: standard deviation of the horizontal directivity across the band of interest (unit: dB)

• $P_{Rgain_{\text{avg}}}$: mean of the horizontal realized gain across the band of interest (unit: dB)

• $P_{\text{patt}}$: average maximum pattern difference between the simulated aperture and a resonant quarter-wave thin monopole across the band of interest (unit: dB)

In case of the utopia point optimization, the total cost is the output from a single objective function. If the user is carrying out multi-objective optimization for a compromised solution, numerical values of the utopia point must be allocated in this script. In addition, the user is responsible for selecting appropriate weights to specify the relative importance of different objectives.

**A.2.7 BGA.m**

This is a typical genetic algorithm that works with binary variables. The evolution process utilizes the Roulette-Wheel selection, single-point crossover, random mutation, and elitist strategy.

Inputs:

• *Gene*: a collection of all chromosomes in a single generation before reproduction

• *cost*: costs associated with these chromosomes

Outputs:

• *Gene_new*: a new group of chromosomes after reproduction
Other default program setups are:

- `crossrate`: crossover rate (0.9)
- `mutrate`: mutation rate (0.02)
- `rplrate`: replacement rate (0.75)
- `el`: number of chromosomes (the best individuals) passed to the next generation without crossover or mutation (2)

### A.2.8 PfmPlot.m

This script consists of a group of Boolean variables for generating desired plots. 1 refers to execution, whereas 0 stands for ignorance. All output files are saved in the subdirectory “Plot”. Below is a list of the indicators:

For the ultimately optimal design:

- `flagPlotProfile`: antenna 2-D profile
- `flagPlotDir`: horizontal directivity vs. frequency
- `flagPlotImped`: complex input impedance vs. frequency
- `flagPlotS11`: reflection coefficient vs. frequency
- `flagPlotVSWR`: VSWR vs. frequency
- `flagPlotRgain`: horizontal realized gain vs. frequency
- `flagPlotPatt`: E-plane radiation pattern ($-90^\circ$ to $90^\circ$)
- `f_patt`: frequency (unit: Hz) at which the radiation pattern is plotted
For the GA performance check:

- *flagPlotCost*: cost vs. generation
- *flagTextCostFront*: best individuals of each generation

Note that the last indicator *flagTextCostFront* is for exporting a text file that contains the best individuals of each generation throughout the GA optimization. The formatted outputs include the binary string (chromosome), total cost, control vertices, complex input impedance, horizontal directivity, and individual cost items.

**A.2.9 PfmAnim.m**

This script consists of a group of Boolean variables for generating animations that reflect the profile and performance variation of the best individual in each generation. 1 refers to execution, whereas 0 stands for ignorance. All output files are saved in the subdirectory “Animation”. Below is a list of the indicators:

- *flagAnimProfile*: antenna 2-D profile
- *flagAnimDir*: horizontal directivity vs. frequency
- *flagAnimVSWR*: VSWR vs. frequency
- *flagAnimRgain*: horizontal realized gain vs. frequency
- *flagAnimCost*: cost vs. generation