DEVELOPMENT OF FOUR NOVEL UWB ANTENNAS
ASSISTED BY FDTD METHOD

DISSERTATION

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By

Kwan-Ho Lee, B.S.E.E., M.S.E.E.

* * * * *

The Ohio State University

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Dissertation Committee:

Robert Lee, Adviser
Chi-Chih Chen
Fernando L. Teixeira

Approved by

Adviser
Graduate program in
Electrical Engineering
ABSTRACT

Due to high demand for wide bandwidth applications, UWB antennas have received significant attention in many commercial and military application areas. They can provide very wide bandwidth information with a single antenna configuration. However, designing UWB antennas have very strict requirement such as broadband matching, broad beamwidth, and good efficiency throughout the operational frequency band which is generally difficult to obtain.

In this work, the finite different time domain (FDTD) method was selected for the design and optimization of UWB antennas in many different application areas. They include ground penetrating radar (GPR), anechoic chamber feed antenna, near field probe antenna and tapered chamber feed. All these antennas require UWB operation, dual linear polarization, and broad beamwidth. For each application area, they have their own detail operation requirements. With the help of the FDTD code and through understanding, the antennas are deeply studied and analyzed for the final design. This process saves time and cost compared to the repeated prototyping. For the verification of the numerical result, prototype antennas are built, measured and compared to its numerical model result. The measurement and the simulations agree due to the realistic modeling of the geometry.
For my parents, my wife Choon-Seon, Son Liam, Daughter Kate and the Truth
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VITA

August 11, 1970 ........................... Born - Sang-Joo, Korea

March, 1997 ............................. B.S. Radio Science and Engineering,
Kwangwoon University, Seoul, Korea

December, 1996 - July, 1997 ............ Researcher, SamSung Telecommunica-
         tion Research Center, Seoul, Korea

September, 1997 - December, 1999 ........ Graduate Research Associate,
The ElectroScience Laboratory,
The Ohio State University

December, 1999 ........................... M.S. Electrical and Computer Engineer-
ing, The Ohio State University

December, 1999 - Present ................. Graduate Research Associate,
The ElectroScience Laboratory,
The Ohio State University

PUBLICATIONS

REFEREED JOURNAL ARTICLES

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FIELDS OF STUDY

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Studies in:

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CHAPTER 1

INTRODUCTION

Ultra Wideband (UWB) technology has been widely used and extensively studied for many decades in military and commercial areas [1, 2]. Basically, UWB radar system uses wide bandwidth instead of short bandwidth signals to achieve better resolution and obtain information more.

The finite difference time domain (FDTD) method has been found suitable for the analysis of the transient response of a complex system, since the FDTD is an explicit approach to solve Maxwell’s equations in the time domain [3]. It also has the advantage of relatively easy handling of the inhomogeneous medium under the consideration and acquiring wide frequency band information from the single run. Assisted by the FDTD method, novel UWB antennas could be designed and optimized for desired radiation performance in different application areas.

With the aid of the FDTD method, four novel UWB antenna designs are examined and optimized for their application areas. Characterizing the electrical properties of these novel UWB antenna designs that reflect the geometric and environmental complexity is one of the major tasks. Based on the physics and electromagnetic theory analysis, understanding and further optimization to meet the UWB antenna requirements is the focus of the dissertation. Common features of a UWB antenna include broad bandwidth, impedance matching, and
frequency independent radiation properties. Four important application areas to be focused are ground penetrating radar (GPR), Near-field range, Tapered Chamber range, and Compact range. Each of these areas needs a UWB antenna with different design specifications. The FDTD method can be used to utilize the UWB antenna designs to predict the potential problems before constructing the prototype antennas.

The GPR system is a non-invasive sensing tool for detection and classification of buried objects in the ground. It has many application areas in geophysics, archeology, civil engineering, environmental engineering, and military \[4\] \[5\]. One key component in a GPR system is the antenna. Desirable features for a good GPR antenna should include broad bandwidth, low antenna ringing, mobility, ground independent matching condition and small size. Even though GPR has many application areas, characteristics of the GPR antennas such as radiation patterns, polarizations and characteristic impedances are not often well established due to their complicated operational environment. For instance, the input impedance of the commonly used dipoles or flat bowtie dipoles are directly affected by the electrical property of the particular ground for antennas operated close to the surface. Moreover, the amount of energy coupled into the ground changes as the permittivity increases and hence the radiation patterns also depend on the soil permittivity \[6, 7, 8\]. One major disadvantage is that the antenna characteristics in the field become dependent on the electrical properties of the ground and surroundings. This also makes calibration more difficult. In order to make antenna characteristics less susceptible to ground characteristics, a new dielectric-loaded horn-fed bowtie (HFB) antenna design was introduced in \[9\]. The HFB antenna was designed to minimize antenna ringing by (1) employing a stable and well matched surge impedance and (2) using specially designed tapered resistive loadings. Unlike most conventional antennas, the surge impedance was designed to be less
dependent on the ground property because the feed point is elevated off the ground. Low loss dielectric material was then used to fill the space between the elevated feed front and the ground surface to reduce ground-surface reflections and increase the electrical height of the feed. Both single-polarized and dual-polarized HFB antenna prototypes have been built and employed in actual applications.

In recent years, the finite-difference time-domain (FDTD) method has been used to simulate GPR measurements [10, 11, 12, 13]. Some of the previous studies modeled GPR antennas as point sources or short dipoles with or without the presence of conducting shields [14]. However, no FDTD models have been applied to more realistic and complicated antennas like the HFB antenna. Here, utilized 3D FDTD model is proposed for further design optimization.

A near field measurement range is a cost-effective and space-saving technique for determining the far-field patterns of an antenna via near-to-far-field (NTF) transformation. The indoor environment provides a better environmental control compared to the use of an outdoor far-field range. Most important issues in a near field range are associated with reducing the probe-antenna under test (AUT) interaction for measurement accuracy. An open-end waveguide (OEW) has commonly been used as a probe in near field ranges. However, significant interaction between probe and AUT is unavoidable due to the conducting waveguide structure. Also, an OEW can only be used for bandwidth no more than 2:1, like all waveguides, and only a single polarization is available at the time of probing. To solve these problems, a new dual polarized UWB dielectric rod probe (DRP) antenna was developed [17].

Cylindrical dielectric waveguides have been extensively studied in various aspects [18, 19] especially, the type of wave guide that supports hybrid mode, the so-called dipole mode
which has no cut-off frequency making it suitable for broadband applications. Far-field patterns for several different shapes of polyrod antennas have been investigated by Muller et al. [20], Watson et al. [21] and others [22, 23]. However, almost all of the previous studies have shown narrow bandwidth performance due to the narrow band characteristics of the feed structure. The new design improves the bandwidth by employing a broad bandwidth launcher which excites the hybrid mode only into the circular dielectric waveguide section. For the broad patterns, techniques of tapering the radiation section is studied. The new design demonstrates a reduced probe-AUT interaction, symmetric E- and H- plane patterns, frequency independent patterns and dual polarization capability. In this study, novel UWB DRP was designed and studied for broadband application. A numerical model of the DRP is created and simulated using FDTD method.

Another important use of a UWB antenna is inside an anechoic chamber. Compact range is a facility for antenna radiation and RCS measurements. The parabolic reflector converts the spherical wave fronts which is radiated from the feed antenna into plane waves. A good reflector feed should have constant beamwidth over the operation frequency band, a fixed radiation center that has to be placed at the focal point, low sidelobes to minimize the leakage, and small size. Horn antennas are often used as a compact range feeds. Disadvantages of these horns include frequency-dependent radiation patterns, asymmetric E- and H- plane patterns, and single polarization. In order to develop a better feed, a new design that uses a solid dielectric horn was introduced [24].

Dielectric horn antenna (DHA) supports hybrid mode or $EH_{11}$ mode, which provides broadband characteristics. The dielectric antenna design has a long history, and related literature can be found in [25, 26]. Solid dielectric antenna patterns have been studied from Salema et al [26], but narrow beamwidth and single polarization were obtained due to their
feeding structures. Later, a narrowband shielded dielectric horn antenna was developed using a circular horn to improve the pattern symmetry and bandwidth. From the idea of using dielectric material as a whole antenna as shown in DRP [27], a novel UWB-DHA is now proposed for broad bandwidth application.

The new UWB DHA design provides dual linear polarization, broad bandwidth (2 - 18 GHz), frequency independent symmetric E- and H-plane patterns, reasonably good efficiency, and broad 3 dB beamwidth. The result obtained from the prototype and FDTD model indicates that the newly developed DHA has a wide bandwidth, symmetric E- and H-plane patterns, dual polarization characteristics. In this dissertation, further optimization and analysis based on the various parameters (horn angle, resistive card profile, launcher arm size etc.) analysis is done to understand the characteristics to improve the mentioned features.

The final application area for UWB antenna design is inside a tapered chamber range. A taper chamber is a spherical range facility that provide a pure spherical incident field for the measurement of antenna radiation and radar cross section (RCS). Common practical feed antennas used as the wave launcher in the tapered chamber include log-periodic and horn antennas [15] [16]. These antennas have suffered from frequency dependent phase center location, diffraction from antenna arms, and single polarization. Also, the radiation center of this antenna, when positioned inside the tapered chamber is not at the apex of the chamber. This causes reflections from chamber, walls and the antenna.

It is also desirable to be able to perform measurements in the tapered chamber as low frequency as possible. This is difficult to achieve using the conventional approach because the absorber material become inefficient at low frequencies. A new wave launcher design that utilizes quad-ridged arms antenna with an appropriate placement of absorber at the
chamber walls was studied. FDTD simulations indicated that this new design could provide the fixed phase center, undisturbed spherical wave front at the test zone, low frequency of operation, and dual polarization. The field distribution effect of the electric property of different absorbing materials and relation between the antenna size and operation frequency band are studied.

The organization of the dissertation is following. Chapter 2 introduces basic FDTD algorithm and modeling techniques regarding accurate numerical model development of the antenna components which are composed of actual measurement system. From the Chapter 3 to Chapter 6, four novel UWB antennas are introduced and characterized from the FDTD analysis. In addition, prototype antenna construction, measurement, and optimization process are discussed. Comparison of the result from measurement and simulation can be found throughout these chapters. Many important issues related to the designing, modeling, and prototyping are deeply studied and discussed. In Chapter 7, the conclusion and the future research for these four UWB antennas will be discussed.
CHAPTER 2

FINITE DIFFERENCING TIME DOMAIN METHOD IN UWB ANTENNA DESIGNS

Finite difference time domain (FDTD) method is a powerful tool for pulsed radiation system analysis. And it has been found in wide application in electromagnetic wave propagation problems [28]. Due to the merit of acquiring wide range of frequency information from the single run and flexibility of inhomogeneous structure handling, FDTD method fits well in the UWB antenna characterizations and optimization. In this chapter, brief review of the FDTD algorithms and modeling issues regarding the accurate UWB antenna design are discussed.

2.1 Three Dimensional Maxwell’s Equations in Rectangular Coordinate System

It is noted that all of the UWB antenna applications in this dissertation have been formulated in full three dimensional analysis. Three dimensional scalar Maxwell’s curl equations in rectangular coordinate system are expressed as:
Yee algorithm solves the above curl equations by applying space and time derivation.

For example, the discretized curl equations can be derived as:

\[
\begin{align*}
\frac{\partial H_x}{\partial t} &= \frac{1}{\mu} \left( \frac{\partial E_y}{\partial z} - \frac{\partial E_z}{\partial y} - \rho' H_x \right) \\
\frac{\partial H_y}{\partial t} &= \frac{1}{\mu} \left( \frac{\partial E_z}{\partial x} - \frac{\partial E_x}{\partial z} - \rho' H_y \right) \\
\frac{\partial H_z}{\partial t} &= \frac{1}{\mu} \left( \frac{\partial E_x}{\partial y} - \frac{\partial E_y}{\partial x} - \rho' H_z \right) \\
\frac{\partial E_x}{\partial t} &= \frac{1}{\varepsilon} \left( \frac{\partial H_z}{\partial y} - \frac{\partial H_y}{\partial z} - \sigma' E_x \right) \\
\frac{\partial E_y}{\partial t} &= \frac{1}{\varepsilon} \left( \frac{\partial H_z}{\partial x} - \frac{\partial H_x}{\partial z} - \sigma' E_y \right) \\
\frac{\partial E_z}{\partial t} &= \frac{1}{\varepsilon} \left( \frac{\partial H_y}{\partial x} - \frac{\partial H_x}{\partial y} - \sigma' E_z \right)
\end{align*}
\]

\[ (2.1) \]

\[
\begin{align*}
H_x|_{i,j,k}^{n+1/2} &= \left( 1 - \frac{\rho'_{i,j,k} \Delta t}{2\mu_{i,j,k}} \right) H_x|_{i,j,k}^{n-1/2} + \left( \frac{\Delta t}{\mu_{i,j,k}} \right) \left( \frac{E_y|_{i,j,k+1/2}^{n+1} - E_y|_{i,j,k-1/2}^{n}}{\Delta y} \right) \\
H_y|_{i,j,k}^{n+1/2} &= \left( 1 - \frac{\rho'_{i,j,k} \Delta t}{2\mu_{i,j,k}} \right) H_y|_{i,j,k}^{n-1/2} + \left( \frac{\Delta t}{\mu_{i,j,k}} \right) \left( \frac{E_x|_{i,j,k+1/2}^{n+1} - E_x|_{i,j,k-1/2}^{n}}{\Delta x} \right) \\
H_z|_{i,j,k}^{n+1/2} &= \left( 1 - \frac{\rho'_{i,j,k} \Delta t}{2\mu_{i,j,k}} \right) H_z|_{i,j,k}^{n-1/2} + \left( \frac{\Delta t}{\mu_{i,j,k}} \right) \left( \frac{E_x|_{i,j,k+1/2}^{n+1} - E_x|_{i,j,k-1/2}^{n}}{\Delta x} \right)
\end{align*}
\]

\[ (2.3) \]
Rest of the scalar curl equations are need be converted into discretized version as shown in
the above for the time and the space update for the electric and magnetic interactions of the
general mediums and the geometries. Figure 2.1 shows the Yee cubic cell with the six field
vector components.

In most cases, satisfying the Courant criteria is a reference for the result to be stable.
And the time and space interval in three dimensional space has been chosen by the relation
in Eq. 2.4.

$$\Delta t = \frac{\Delta}{c \sqrt{3}}$$

(2.4)

For the cube lattice case, $\Delta = \Delta x = \Delta y = \Delta z$. In some cases, this lattice size is mostly depend
on the geometry. If the problem geometry has significantly affect the performance (curved
surfaces), determination of the size of the lattice is following the criteria of non-distorted
geometrical modeling which is much smaller than the length determined from the stability
condition.

### 2.2 Modeling of Resistive card loadings

Modeling and analyzing a fine geometrical object in finite difference time domain
(FDTD) has been deeply studied and demonstrated accurate result in many papers [28].
One of the easiest ways of modeling of fine features is the use of uniform grid configura-
tion. However, the computational cost will be extremely expensive when it requires elec-
trically large geometry construction, since the limitation on FDTD method largely depend
on the computer resources. In order to decrease the computational cost and the resources,
many attempts such as sub-griding and hybrid (FETD) method are suggested [29]. For the
planar material case, the conventional uniform FDTD grid requires the fine grid configu-
ration corresponding to the thickness of the material for the accurate property analysis in
order to take account the effect of tangential and normal electromagnetic field components to the surface of the thin materials. For example, the resistive film is the one of the largely used thin dielectric and conductive materials in many applications. Thus, the new approach using the idea of contour-path method can resolve the computation resources and accuracy problems with least amount of changes in conventional uniform grid FDTD algorithm.

2.2.1 Resistive card loading in three dimensional FDTD

In order to reduce the antenna rignings, resistive loading have been applied to reduce the reflections from the finite size antenna end such as horn type antennas [30] and dipole antennas [31, 32]. In the FDTD model, the R-card is modeled as a single-cell layer with a tapered conductivity ($\sigma$) corresponding to the desired sheet resistance ($R_s$). Conductivity along the $x$ direction is calculated by $\sigma = 1/(R_s \tau)$ where $\tau$ is the thickness of single layer. This assumption is valid when $\tau$ is much greater than the penetration depth but much smaller than the free space wavelength.

Three different profiles for the sheet resistance of the R-card were studied. The three profiles being, linear, quadratic and exponential profiles given as

\[
R_{\text{lin}}(x) = R_{\text{min}} + \left(\frac{x}{l}\right) [R_{\text{max}} - R_{\text{min}}]
\]

\[
R_{\text{quad}}(x) = (R_{\text{max}} - R_{\text{min}}) \left(\frac{x}{l}\right)^2 + R_{\text{min}}
\]

\[
R_{\text{exp}}(x) = R_{\text{max}} e^{\frac{2(l-x)}{\gamma l}}, \quad \text{where} \quad \gamma = -\frac{l}{\ln(R_{\text{min}}/R_{\text{max}})}
\]

where $R_{\text{min}} = 1\,\Omega/\square$ is the sheet resistance of the R-card closest to the PEC arm and $R_{\text{max}} = 300\,\Omega/\square$ is the sheet resistance at the other end of the R-card. $R_{\text{min}}$ was chosen to be small so that there would be no diffraction at the PEC/R-card interface. A more detailed analysis on this aspect can be found in [33]. Figure 2.2-(a) and (b) plot the exponential, quadratic and linear resistive taper. The lateral edges of the R-card were kept aligned to
the edges of the PEC arms to avoid undesired diffractions. To study how well the different profiles attenuate surface currents on the arms of the antenna, the time-domain reflected magnetic field component, $H^r_x(t)$, which is in the neighbor of the R-card and is proportional to the surface current density, is compared for the different resistive tapers in Figure 2.2-(c). From the result, a linear taper provides a better performance, i.e. lower reflection at low frequency end, due to relatively short taper length with respect to wavelength. As an example, Figure 2.3 shows the three dimensional equivalent tapered resistive card model for GPR antenna.
Figure 2.2: Resistivity and conductivity profile and time domain $H_x$ component observed at a point on the PEC arm of the planar bowtie for various cases
Figure 2.3: Example linear profile and its equivalent resistive card model for FDTD simulation
2.3 Coaxial Cable Modeling and Time Domain Pulse Excitation

Wide bandwidth and minimal antenna ringing is the most important in designing the UWB antennas. These two properties are closely related to the condition of impedance matching at the feeding terminal. A good impedance matching prevents antenna late time ringing by eliminating the reflection as the currents reflected back from the ends of antenna areas. Figure 2.4 shows an example of the real antenna feeding structure which is modeled in this section.

In the FDTD grid formation, a grid size of \( dh = 0.0063 \) m, for example, was chosen to model the coaxial cable. The inner and outer conductors of the coaxial cable were extend into the PML to simulate infinitely long cables in the -\( z \) direction. The TEM mode is excited within the coaxial line through the specification of a time dependent current given by

\[
J(t) = \sqrt{2}e^{t/\alpha}e^{-\alpha^2} \quad (2.6)
\]

where \( \alpha = (t - t_0)/T_b \), \( t_0 = 3.2T_b \) and \( T_b = 8.33 \times 10^{-10} \) sec. The parameters for this differential Gaussian pulse were chosen because it contains significant energy in the operating
frequency band of 10 to 800 MHz. Because of the rectangular nature of FDTD, the four coaxial lines are modeled with square inner and outer conductors. The antenna is formed with four conical plates, each of which is connected to the inner conductors of one of the coaxial lines. As shown in Figure 2.5, A1 and A2 are for the one pair of plates, and B1 and B2 are for the other pair of plates. The polarities of the excitation currents are also indicated in Figure 2.5-(b). It should be noted that the coupling between the pairs of coaxial cables can be ignored due to the orthogonality of the fields so that the characteristic impedance one pair of cables does not affect the impedance of the other pair of cables.

2.3.1 Modeling of Coaxial Cables

Similar to the real antenna feeding structure, a pair of coaxial cables was placed in the FDTD grid. From the conventional equations for a coaxial line, the characteristic impedance can be obtained and applied to the rectangular grid formation.

\[
Z_0 = \frac{R + j\omega L}{\gamma} = \sqrt{\frac{R + j\omega L}{G + j\omega C}}
\]  

(2.7)

It is noted that the inner and outer conductors are PECs which imply that the \( R \) and \( G \) go to zero, and the equation 2.7 reduces to

\[
Z_0 = \sqrt{\frac{L}{C}}
\]  

(2.8)

with the capacitance and inductance of the coaxial cable being given by

\[
L = \frac{\mu}{2\pi} \ln \frac{a}{b}, \quad C = \frac{2\pi\varepsilon'}{\ln(a/b)}
\]  

(2.9)

where \( a \) and \( b \) are radii of the inner and outer conductors, respectively. For the square coaxial line, the impedance was found numerically through the FDTD simulation. Figure 2.6 shows the comparison between the simulation and calculation. Impedance difference is come from the approximation and well agree to [34].
Figure 2.5: Coaxial cable modeling in rectangular FDTD grid
Figure 2.6: Comparison between the calculated and simulated characteristic impedance of the coaxial cable: dielectric constant ($\varepsilon_r$) = 1.5

2.4 Perfect electric conductor (PEC) Plate Modeling

The thin plate perfect electric conductors (PEC) models are employed for creating the conductive antenna arms. Mostly, thickness of the conductive antenna arms are electrically very small and can be ignored within the operation frequency. Thus, instead of using cubic cell PEC, thin plate models are generated by applying the zeros values for tangential component of the electric field along the plate geometry.

Since the coordinate system is a rectangular and uniform grid, curved surface was modeled as stair stepping scheme. Figure 2.7 represents the PEC thin plate configuration and its stair step modeling.

2.5 Perfect Matched Layer (PML) Modeling for Inhomogeneous Media

In computation domain, radiation/scattering problem can be solved since computation domain is truncated by the appropriate absorbing boundary condition to obtain the far-field
result as in infinite space. Berenger introduced alternative perfect matched layer (PML) with split field components [35] for absorbing the plane waves in arbitrary frequency and incident angle. Later, unsplit field PML was introduced [36] and this method is employed in here. For mesh truncation example, Figure 2.8 illustrate the computation domain and PML with wave source located at the center.

In ground penetrating radar (GPR) study, antenna is operating on the lossy ground. Thus, anisotropic PML (APML) for GPR applications is needed. An anisotropic medium characterized by

\[
\begin{align*}
\nabla \times \vec{E} &= -j\omega \mu [\Lambda] \vec{H} \\
\nabla \times \vec{H} &= j\omega \varepsilon_{\text{eff}} [\Lambda] \vec{E}
\end{align*}
\]

where \[ [\Lambda] = \begin{pmatrix}
\frac{s_x s_y}{s_z} & 0 & 0 \\
0 & \frac{s_x s_z}{s_y} & 0 \\
0 & 0 & \frac{s_y s_z}{s_x}
\end{pmatrix}, \quad s_p = 1 + \frac{\sigma_p}{j\omega}
\]

(2.10)

can provide a reflectionless interface for all frequencies, polarizations and angles of incidence by proper choice of material properties. However, for lossy media the permittivity
Perfect matched layer (PML)

Computation domain

Radiation Source

Radiation Radiation

Perfect matched layer (PML)

Computation domain

Figure 2.8: 2D FDTD domain configuration example - computation domain is surrounded by perfect matched layers (PML)

takes that form \( \varepsilon_{eff} = \varepsilon + \frac{\sigma}{j\omega} \) and hence the original formulation needs to be modified. The second curl equation given by

\[
\left( 1 + \frac{\sigma_p}{j\omega} \right) \left( \nabla \times \vec{H} \right)_p = j\omega \left( \varepsilon + \frac{\sigma}{j\omega} \right) \left( 1 + \frac{\sigma_q}{j\omega} \right) \left( 1 + \frac{\sigma_r}{j\omega} \right) E_p \tag{2.11}
\]

has \( \frac{1}{(j\omega)^2} \) dependence and can be solved by introducing a scaled electric field \( \tilde{D} \) as shown below. In the above equation, \( p, q \) and \( r \) would represent the coordinate directions.

\[
\left( 1 + \frac{\sigma_p}{j\omega} \right) \left( \nabla \times \vec{H} \right)_p = j\omega \left( 1 + \frac{\sigma_q}{j\omega} \right) \left( 1 + \frac{\sigma_r}{j\omega} \right) D_p \tag{2.12}
\]

where

\[
D_p = \left( \varepsilon + \frac{\sigma}{j\omega} \right) E_p \tag{2.13}
\]
Discretizing these equations in time domain, we obtain

$$D^{n+1}_p = \left[ \frac{2 - (\sigma_q + \sigma_r)\Delta t}{2 + (\sigma_q + \sigma_r)\Delta t} \right] D^n_p - \left[ \frac{2\sigma_q\sigma_r(\Delta t)^2}{2 + (\sigma_q + \sigma_r)\Delta t} \right] \sum_{m=0}^n D^m_p$$

$$+ \left[ \frac{(2 + \sigma_p\Delta t)\Delta t}{2 + (\sigma_q + \sigma_r)\Delta t} \right] \left( \nabla \times \mathbf{H} \right)^{n+\frac{1}{2}}_p + \left[ \frac{2\sigma_p(\Delta t)^2}{2 + (\sigma_q + \sigma_r)\Delta t} \right] \sum_{m=0}^{n-1} \left( \nabla \times \mathbf{H} \right)^{m+\frac{1}{2}}_p$$

$$E^{n+1}_p = \frac{2}{2\varepsilon + \sigma\Delta t} \left( D^{n+1}_p - \sigma\Delta t \sum_{m=0}^n E^m_p \right)$$

(2.14)

The above equations can be used to update the electric field. The update equation for the magnetic field is given by,

$$H^{n+1}_p = \left[ \frac{2 - (\sigma_q + \sigma_r)\Delta t}{2 + (\sigma_q + \sigma_r)\Delta t} \right] H^n_p - \left[ \frac{2\sigma_q\sigma_r(\Delta t)^2}{2 + (\sigma_q + \sigma_r)\Delta t} \right] \sum_{m=0}^n H^m_p$$

$$- \left[ \frac{(2 + \sigma_p\Delta t)\Delta t}{2 + (\sigma_q + \sigma_r)\Delta t} \right] \left( \nabla \times \mathbf{E} \right)^{n+\frac{1}{2}}_p - \left[ \frac{2\sigma_p(\Delta t)^2}{2 + (\sigma_q + \sigma_r)\Delta t} \right] \sum_{m=0}^{n-1} \left( \nabla \times \mathbf{E} \right)^{m+\frac{1}{2}}_p$$

Although the PML is matched to the computational boundaries, it is valid in limited bandwidth when medium is frequency dependent. FDTD is a time domain method which require constitutive parameters such as permittivity and permeability, need to be specified as constants. However, it can be solved by using frequency-dependent FDTD method which uses recursive convolution. It also demands more computer resources such as computation time and storages, compared to the regular FDTD. In most situation, it depends on the material properies within the frequency band of interest. Sometimes frequency-dependent FDTD is useful in spite of larger resources usage. In some cases, sub-frequency technique which require multiple simulation within the limited bandwidth can be useful. Decision can be made by their material property.

2.6 Partition Scheme

To minimize the memory usage, we have adopted a partitioning scheme [37]. When problem geometries are complicated, partitioning the total computation domain could be
an efficient way to model the geometry with saving some memory resources. Partition scheme is a geometrical plan to divide inhomogeneous block into multiple homogeneous blocks. Thus, three dimensional material matrix can be replaced by several constants (permittivity, permeability, electric and magnetic conductivities). For example, partition of the inhomogeneous medium can be done as multiple small inhomogeneous medium and homogeneous medium as shown in Figure 2.9. By reducing the size of inhomogeneous matrices, computation time and space can be reduced.

2.7 Summary of the Chapter

The five modeling issues are briefly discussed to incorporate with the UWB antenna designs. Computed result is well agreed to that of measurement. Nowadays, along with abundance of computation resources such as memory and speed of the CPU, more geometrically detailed features can be discretized without losing their own characteristics. In
most of UWB antenna designs, the feeding structures are one of the most important components to obtain a good impedance match. For example, circular coaxial cable was modeled with uniform rectangular grid with 50 Ω characteristic impedance. Semi-ridged coaxial cables with outer conductor of 3 mm diameter and 1 mm diameter of inner conductor can be modeled more accurately, if sub gridding is adapted. Many previous works for sub-grid schemes could be found [38, 39]. It is a combination of two different domains - coarse grid and finer grid. By the careful treatment of the boundary of two different grid spacing, numerical stability and accuracy was achieved. Thus, if electromagnetic characteristic is sensitive due to their geometry, sub gridding could be a best way to solve the Maxwell’s equation.

Another important modeling issue is in the resistive card design. The resistive card used in horn-fed bowtie (HFB) antenna prototype was constructed in-house using multiple layers of commercial window films [40]. In numerical modeling, one cell thickness equivalent resistive card model was employed. Precisely, thickness of resistive card used in prototype is electrically very small and neglectable for normal field components. Although the size of cubic cell is small enough to ignore the diffraction effect due to the thickness, it is still affecting the reflection coefficient and this can be improved by applying the different modeling technique such as sub-cell implementation [41] and thin sheet approximation [42].
CHAPTER 3

DESIGN OF HORN-FED BOWTIE (HFB) ANTENNA FOR GPR APPLICATIONS

Antennas for ground penetration radar (GPR) applications have stringent specifications in terms of wide bandwidth, no late time ringing effects, and input impedance matching. In order to penetrate deep enough in lossy ground soils and in applications associated with detection and classification of Unexploded Ordnances (UXO) or land mines, it is common to have operating frequencies in the range from 10 MHz to 800 MHz. One of the most popular ultra wide band (UWB) antennas for GPR applications is the bow-tie antenna. In order to achieve a better system mobility over the ground surface and a better focusing the main beam into the ground, a fully polarimetric dielectric loaded horn-fed bow-tie (HFB) antenna has been recently developed [9], [43].

Due to its flexibility, the finite-difference time-domain (FDTD) method has been widely used in recent years for the numerical simulation of GPR systems [10, 11, 12, 13]. Some of the previous studies have modeled GPR antennas as a series of point sources or short dipoles with or without the presence of conducting shields [14, 44]. In order to better characterize HFB antennas and to provide a more convenient tool for their design and optimization, a full-scale detailed three-dimensional (3-D) FDTD model of a dual-polarized
HFB prototype was developed in this work and simulated for GPR applications. To reduce the computational cost, a special partition scheme [37] is adopted for the 3-D FDTD domain. This scheme divides the whole inhomogeneous region into several small homogeneous regions. In each homogeneous region, volumetric material property matrices are replaced by constants to save the memory. This partition scheme for modeling the electrically large HFB antenna in the presence of ground also allows for faster simulations on a personal computer. An anisotropic perfectly matched layer (APML) especially formulated for the dielectric or lossy half spaces [45, 12, 46] is implemented.

This chapter is organized as follows. The HFB antenna design is discussed in Section I. Section II describes the construction of the FDTD model for the dual-polarized HFB design and the performance of the resistive-film loading which is optimized for a given length. Section III presents various HFB antenna characteristics obtained from the FDTD simulations.

3.1 Basic Dual-polarized HFB Antenna Design

Figure 3.1 illustrates the basic structure of the dual-polarized UWB HFB antenna design. This is somewhat similar to a planar bowtie dipole with the feed point being raised off the ground. The feed section resembles that of a small TEM horn except that it is filled with low loss dielectric material. Each antenna arm is smoothly curved in the transition from the horn section to the planar bowtie dipole section. The ends of the dipoles are terminated with tapered resistive cards (R-card) to reduce antenna ringing.

3.1.1 Resistive Taper Section

Tapered R-cards have many useful applications for radiation and scattering control [47, 48, 49], but commercial tapered R-cards are often expensive and have very limited choices
of tapering profile and taper length. The R-card used in the HFB prototype was constructed in-house using multiple layers of commercial window films [47]. These have various sheet resistance for different percentage of light transmission. When multiple films are overlaid properly together, one can obtain a desired resistivity profile with desired taper length. Figure 3.1 illustrates how the tapered R-card was constructed for the HFB prototype.

The objective of the resistive card is to reduce reflections by gradually dissipating the currents propagating toward the end of each antenna arm. This requires the resistivity on the R-card to be tapered from a small value to a large value along the antenna arm. An
exponential taper of the resistivity was adopted in the HFB prototype with a tapering shown below.

\[ R_{\text{exp}}(x) = R_{\text{max}} e^{\frac{2l}{x}} \gamma, \quad \text{where} \quad \gamma = -\left(\frac{l - 1}{\ln(R_{\text{min}}/R_{\text{max}})}\right) \]  

(3.1)

\( R_{\text{min}} = 3.3\Omega/\square \) is the initial sheet resistance of the R-card at the PEC/R-card interface, and \( R_{\text{max}} = 1150\Omega/\square \) is the sheet resistance at the far end of the R-card, \( l = 0.63 \text{ m} \) is the length of the R-card, and \( x \) is the distance along the R-card from the PEC arm.

### 3.1.2 Feed Section

The feed section of HFB resembles a dual polarized TEM horn except that the end of each antenna arm is curved outward gradually to be connected to the flat bowtie section, and the internal space of the horn was filled with low loss dielectric material. The geometry of the horn and the antenna arms was chosen based upon the trade-off among the dielectric constant, size, weight, and cost. The objective was to obtain a surge impedance of 100 \( \Omega \) to match to the characteristic impedance of the feeding twin-coaxial cables shown at the bottom of Figure 3.1 (each cable has a characteristic impedance of 50 ohms). Although tabulated characteristic (or surge) impedances for an infinite TEM horn with arbitrary geometry are available [50] [51], the exact impedance of a dual-polarization TEM horn with dielectric filling is complicated to obtain analytically. The experimental data obtained from [52] was used during the construction of HFB prototype. Note that the center of each coaxial cable was connected to one antenna arm and each pair of the 50 \( \Omega \) coaxial cable feed one polarization. A 0 - 180° broadband hybrid was used as a balun for each pair of cables. Accurate FDTD models recently constructed to calculate the surge impedance for such an
antenna geometry are employed here [33]. The prototype to be analyzed here has a dielectric constant of 5. The plate angle of each antenna arm is 11.5 degrees. The horn angle itself is approximately 150 degrees.

3.2 FDTD Model Description

A full scale model of the UWB HFB antenna prototype requires a minimum of 2.5 m × 2.5 m × 0.63 m space. A spatial cell size of 6.3 mm was chosen to accurately model the geometrical details of the antenna and cable structure [53]. This yields approximately 96 million unknowns. The FDTD grid is shown in Figure 3.2. All dimensions in the model were chosen to be as close to the actual prototype as possible. The four antenna arms were modeled as perfect electric conducting (PEC) plates, and the curved edges and surfaces were approximated by staircases. Each tapered R-card attached to the end of the PEC arm is 63 cm in length and is implemented via a conductive sheet. The ground was assumed to be a lossless half space with relative permittivity of 5.

3.2.1 Heterogeneous FDTD Domain Partition

The antenna geometry under study is very complicated and resides in a complex environment. A traditional FDTD approach to represent the geometry would require either the storage of the material properties for each cell or else a data organization similar to what is used in the finite element method, which would also require a significant amount of memory overhead. To minimize the memory usage, we have adopted a partitioning scheme [37]. The FDTD domain is divided into blocks. The size and number of the blocks are judiciously chosen, so that the material properties within most of the blocks are homogeneous. Within the code, the FDTD algorithm is computed in different ways, and based on the properties of the block, the appropriate FDTD algorithm will be chosen. If the block is
a perfect conductor, the FDTD code will recognize this, and not perform any computations for that block. Thus, there is no need to store either the fields or the material properties for that particular block. If the block is an homogeneous dielectric, the material properties are not treated as a function of the grid points within the block but instead represented just as a constant parameter. Thus only the field values need to be stored for each cell within that block. If the block is an inhomogeneous dielectric, then the FDTD algorithm used will assume a constant permeability and no conductivity. Thus, only the fields and permittivity
must be stored for each cell. In our case, we divide the geometry into 196 blocks with only 5 of the blocks being heterogeneous as demonstrated in Figure 3.2.

3.2.2 Feed Cable Modeling

In the discretized FDTD model, each coaxial cable has a square cross sectional area with a single-cell PEC wire surrounded by four PEC walls. As shown in Figure 3.3 (a), a relative dielectric constant of 1.5 is specified between the center wire and the PEC walls. Each cable is terminated with perfectly matched layer at one end and connected to the tip of an antenna arm at the other end. A balanced excitation is introduced to the opposite pair of cables to excite one antenna polarization as shown in Figure 3.3 (a) and (b). The time history of the response is also recorded at the excitation position to obtain reflection and transmission data. The reflection data \( S_{11} \) is obtained with the excitation and observation points co-located in the same cable and cross-coupling data \( S_{21} \) is obtained with the observation point located at the second cable. A differential Gaussian pulse is chosen as the time-domain excitation current,
Figure 3.3: Coaxial cable modeling in rectangular FDTD grid and the TEM current excitation scheme
where \( \alpha = (t - t_0)/T_b \), \( t_0 = 3.2T_b \) and \( T_b = 8.33 \times 10^{-10} \) sec. These parameters for the Gaussian pulse are determined so as to provide significant spectral energy in the frequency range of 10 to 800 MHz. Figure 3.3-(c) illustrates the pulse.

### 3.2.3 Resistive Card Modeling

In the FDTD model, the R-card is modeled as a single-cell layer with a tapered conductivity \( (\sigma) \) corresponding to the desired sheet resistance \( (R_s) \). Conductivity along the \( x \) direction is calculated by \( \sigma = 1/(R_s \tau) \) where \( \tau \) is the thickness of single layer. This assumption is valid when \( \tau \) is much greater than the penetration depth but much smaller than the free space wavelength [54].

In addition to the exponential taper described in Section 3.1.1, a linear taper with the following taper function was also investigated using the FDTD model as a comparison.

\[
R_{\text{lin}}(x) = R_{\text{min}} + \left( \frac{x}{l} \right) [R_{\text{max}} - R_{\text{min}}]
\]

\( R_{\text{min}} = 3\Omega/\square \) is the initial sheet resistance of the R-card at the PEC/R-card junction, \( R_{\text{max}} = 300\Omega/\square \) is the end sheet resistance. The taper length \( l \) is equal to that of the previous exponential taper, i.e. 0.63 m. As it will be shown shortly, a linear taper provides a better performance, i.e. lower reflection at low frequency end, due to relatively short taper length with respect to wavelength. A more detailed analysis on this aspect can be found in [33]. Figure 3.4 plots the linear resistive taper as well as its position relationship with respect to the antenna arm. The lateral edges of the R-card were kept aligned to the edges of the PEC arms to avoid undesired diffractions (see Figure 3.1).
Figure 3.4: Resistive card overlay configurations for the PEC launcher section ($R_{\max} = 300 \, \Omega/\square$, $R_{\min} = 3 \, \Omega/\square$).

3.3 Characteristics of Dual-polarized HFB Antenna Design

3.3.1 $S_{11}$ & $S_{21}$ and Input Impedance

The simulated and measured reflection and transmission coefficients, $S_{11}$ and $S_{21}$, of the HFB design are compared in Figure 3.5 (a). Note that the antenna is located on the surface of a half space with a dielectric of 5, corresponding to the dry sand in reality. The $S_{22}$ is similar to $S_{11}$ since the both antenna arms have the same design. $S_{11}$ and $S_{22}$ provide the co-polarized backscattering data. $S_{21}$ provides the cross-polarized backscattering data.
Figure 3.5: $S_{11}$, $S_{21}$, and surge impedance of the HFB antenna
A calibration procedure was carried out in a similar manner as done in real measurement using “short” and “matched” (PML) reference loads at the end of the feed cables.

\[
S_{11}(f) = \frac{E_{\text{ant}}(f) - E_{\text{match}}(f)}{E_{\text{short}}(f) - E_{\text{match}}(f)} \quad ; \quad S_{21}(f) = \frac{E_{21}(f)}{E_{\text{inc}}(f)} \quad (3.4)
\]

In the above, \(E_{\text{ant}}\) is the response obtained with the coaxial cables connected to the antenna, \(E_{\text{match}}\) is the response obtained with the coaxial cables connected to matched load, \(E_{\text{short}}\) is the response obtained when the coaxial cables are shorted at the end with a conducting wire, \(E_{21}\) is the response obtained at coaxial cable 2 with antenna connected when the excitation is applied to cable 1, and \(E_{\text{inc}}\) is the incident wave.

It is observed that the both linear and exponential taper have similar performance at frequency above 0.3 GHz where the taper length becomes comparable or longer than one wavelength (considering dielectric constant of 5). It is also observed that the linear taper produces lower reflection level than the exponential taper at frequencies below 0.1 GHz. Overall, the reflection level is less than -10 dB above 0.05 GHz. This verifies broadband characteristic of the HFB design. The measured \(S_{21}\) data is found to be on average 10 dB higher than that predicted from the simulation. This difference is most likely caused by the asymmetry of the construction of prototype antenna arms and the feed structure.

Good agreement between the measurement and simulation is the result of the geometrical fidelity between of the FDTD numerical model and the prototype, including the R-card geometry, conductive plates, dielectric filling and coaxial cable feed modeling. However, the prototype measurement introduces additional environmental variables more difficult to control such as ground loss, slight asymmetry of the antenna arm design due to hand-made fabrication, and discrepancies between the equivalent conductive single layer R-card used in the FDTD model and the thin film R-card conductivity value.
The surge impedance can also be calculated from the $S_{11}$ as shown in Eq. 3.5, often applying a time gate to keep only the first peak associated with the feed point near 0 ns position as shown in Figure 3.6

$$Z_{\text{in}}(f) = \left( \frac{1 + S_{11}(f)}{1 - S_{11}(f)} \right) Z_{0}^{\text{coax}}(f), \quad (3.5)$$

where $Z_{0}^{\text{coax}}$ is the characteristic impedance of the twin-coaxial cable. The resultant surge impedance is shown in Figure 3.5 (b). For most of the band (0.15 GHz $< f <$ 0.6 GHz), the surge impedance is found to be within $100 \pm 10\Omega$ range, as desired.

### 3.3.2 Ground effect

In order to see how the ground properties affect the surge impedance of the HFB design, four different ground dielectric constants: 5, 7, 9 and 11 are simulated. Figure 3.6 shows the reflected field from 0 to 3 nsec. The height of the antenna feed above the ground is equal to 0.1 m. This causes the reflection from the ground surface to be delayed by approximately 1.5 nsec since the antenna dielectric filler has a relative permittivity of 5. This agrees with the significant variations shown in the data near 1.5 nsec position. Most importantly, the first reflection peak arising from the feed point remain unaffected by the ground property, as desired.

### 3.3.3 R-card performance investigation

We investigate two parameters that play an important role in minimizing reflections from the truncated antenna arms. The first parameter is the overlay distance between the PEC and R-card. In the actual HFB prototype, a 5 cm overlay was used to allow the electromagnetic energy to be coupled into the R-card section because the R-card was coated with a protective insulator and could not have a direct electrical contact with the antenna
Figure 3.6: Comparison of reflected electric field difference with various ground profiles

arm. The second parameter is the far-end resistance value that affects the tapering rate of the R-card. If the taper is done too rapidly, undesired diffractions would be produced by the R-card. On the other hand, if the taper is too slow, the far-end reflection may still be too strong.

To investigate the effect of PEC and R-card overlay distances, the following three cases were simulated as shown in Figure 3.4. In case 1 through 3, the overlay distances are 11.3 cm, 5 cm and 0 cm, respectively. The simulated reflection responses are plotted in Figure 3.7. As expected, the overlay distance of the tapered R-card affects the reflection at the PEC end. Note that the R-card in the overlay section is shorted out by the PEC, this section would have an effective resistance of zero regardless of the R-card value. The larger geometric discontinuity in Case 3 provides the stronger junction reflection observed in the
Figure 3.7: Reflected \( E_x \) field in time domain for HFB antennas with different resistive card overlay configurations and using same conductivity profile.

figure. Case 1 and 2 provide a smoother transition and result in a -35 dB reflected field at the end of the R-card. Based on the simulations, we concluded that a linear-tapered R-card with either a 11.3 cm or 5 cm overlay at the PEC/R-card does the best job of suppressing the reflections.

To optimize the choice of \( R_{\text{max}} \), values of 100, 200, 300 and 400 \( \Omega/\square \) were implemented and simulated separately. From the reflected field observed at the feed point, the amount of end-reflection suppression was compared as shown in Figure 3.8, where late time (after 20 ns) antenna reflections can be observed. These results indicate that \( R_{\text{max}} = 300 \Omega/\square \) provides the maximal suppression of the arm end reflections.
Figure 3.8: Comparison of co-polarized \((E_x)\) reflected field in time domain from HFB antenna with the different resistive cards \((R_{\text{max}})\)

### 3.3.4 Antenna Ringing

Figure 3.9 compares snapshots of the instantaneous \(E_x\) field distribution in the vertical (or \(x - z\)) plane with and without the R-card attached to the HFB antenna arms. Without the R-card, significant diffraction and reflection at the end of the PEC arms are observed. The reflected fields later propagate back to the observation point inside the cables as shown in Figure 3.9- (b). On the other hand, the R-card extension significantly reduces the diffraction
Figure 3.9: Snap shots from FDTD simulation for $E_x$ field strength in dB scale where $R_{\text{max}} = 300\Omega/\square$
and reflection at the ends as depicted in Figure 3.9- (a) and lowers the antenna ringing by approximately 20 dB. Note that the signals that propagate back to the feed point are partially reflected due to the imperfect matching. This reflected fields generate the secondary reflection. This process repeats and becomes the well known “antenna ringing” effect, a major clutter source in GPR measurements.

### 3.3.5 Radiated Field Distribution & Polarization

The near-field radiation characteristics are investigated next. Figure 3.10 depicts the simulated horizontal co-polarized and cross-polarized field distributions at a plane 40 cm below the antenna aperture, (corresponding to the ground surface plane), at the center frequency of 400 MHz. The cases with and without the R-card are also plotted for comparison. The fields are nearly linearly polarized in the principal planes. The results with the R-card clearly show a more uniform distribution, because the diffracted fields from the antenna arm ends modify the radiated fields that otherwise would have been close to simple spherical wavefronts. The more uniform field distributions simplify the subsequent signal processing and inverse problem and improve the overall detection/classification capabilities of a GPR system.

As the observation point moves away from the principal planes, the level of depolarization increases and reaches a maximum of approximately -12 dB between the two antenna polarizations. This is, of course, due to the spherical nature of the wavefront. We note that the cross-polarized field levels with the R-card present are a little bit higher. This again may be caused by distributed diffractions along the resistive cards.
Figure 3.10: Comparison of co- and cross-polarized aperture field distributions at $f = 400$ MHz, depth $z = 40$ cm or $0.53 \lambda_{\text{min}}$, $R_{\text{max}} = 300 \Omega/\square$ in R-card
3.4 Conclusion of the Chapter

In this work, a detailed FDTD model was used to incorporate realistic features of UWB HFB antennas such as feeding cables, dielectric loading and tapered resistive terminations. The FDTD model is flexible enough to model different geometries, structures, and materials for both the antenna and the ground medium. Fully-polarimetric simulations were performed to obtain the radiation characteristics of HFB antennas over a broad frequency range. A parametric study on the effect of the resistive taper of the R-card termination was also performed. It was found that a linear taper performs better than the commonly used exponential taper for short taper length. It was also found that a proper overlapping between the PEC and R-card improves the transition and reduces the diffraction at the end of PEC. The R-card termination also significantly reduces the undesired antenna ringing. The surge impedance of the HFB antenna was calculated from the reflection coefficients and was found to be approximately 100 ohms over the entire frequency band of interest. This result confirms the broadband characteristic of the HFB design. The FDTD model also provided useful visualization of dynamic field distributions that can help identify undesired radiations and reflections sources. The near-field distributions of the co-polarized and cross-polarized fields were examined. This information is particularly useful in GPR applications where the depth of the target is unknown. Overall, the simulated results confirm that the optimized HFB antenna design is a very attractive choice for broadband, fully polarimetric GPR applications.

3.4.1 Parameter summary of HFB geometry
### Table 3.1: Important parameters applied in HFB antenna design

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>plate angle</td>
<td>11.5°</td>
</tr>
<tr>
<td>horn angle</td>
<td>150°</td>
</tr>
<tr>
<td>length in x &amp; y</td>
<td>2.52 m</td>
</tr>
<tr>
<td>height in z</td>
<td>0.1 m</td>
</tr>
<tr>
<td>permittivity of dielectric inside the horn</td>
<td>5</td>
</tr>
<tr>
<td>length of resistive card</td>
<td>0.63 m</td>
</tr>
<tr>
<td>resistive card profile</td>
<td>linear profile (3Ω/□ - 100 Ω/□)</td>
</tr>
</tbody>
</table>
CHAPTER 4

DESIGN OF DIELECTRIC ROD PROBE (DRP) ANTENNA FOR NEAR FIELD MEASUREMENT RANGE

Near-field measurements are used to determine the far-field patterns of an antenna under test (AUT) with a better environmental control as well as reduced cost and space compared to a far-field range. Probe correction that reduces the probe effect on the measurement accuracy is a most important issue in near-field measurements and has been discussed in many publications [55, 56]. Such a correction is to remove the probe pattern and polarization effects in a way similar to the removal of the transfer function in a linear system. An ideal probe for a near-field range should have

1. no probe-AUT interaction
2. isotropic pattern
3. dual polarization capability
4. frequency-independent response

Requirement (1) is very important but is often neglected in almost all probe correction techniques. Requirement (2) is neither possible nor practical since there are always range structures behind the probe. A broad pattern (in all planes) with low or no side lobes
makes the probe correction easier. Requirement (3) needs a good isolation between the two polarizations and eliminates the need to perform multiple probing. Requirement (4) implies a broad bandwidth.

As a near field probe antenna, open-end waveguide (OEWG) has been commonly installed and used for measurement. And OEWG characteristics are well known. The limitations of OEWG antenna can be summarized as

1. narrow bandwidth operation
2. asymmetric E- and H-plane patterns
3. frequency dependent radiation patterns
4. strong interaction between the OEWG probe antenna and AUT

The limitation (1) is from the nature of waveguide characteristic. The (4) is mainly caused by the conducting material of the OEWG. In order to overcome these undesired characteristics of OEWG probe antenna, a new design of near field probe is highly desired. Recently, a dielectric rod antenna for GPR application has been developed at OSU and utilized for the land mine detections [27, 57]. For the detection and classification of anti-personnel land mines inside the lossy ground, antenna must have following characteristics - very low antenna-ground interaction, isotropic pattern, wide bandwidth, and low antenna clutter. The mentioned characteristics are required since the target is located at the near the surface of the ground. Because of the characteristics, dielectric rod antenna is an ideal candidate for the near field probe antenna.

In many years, dielectric cylinder waveguide has been intensively studied and investigated [18, 19, 58]. Transverse electric/magnetic (TE/TM) mode attenuation & propagation
study in dielectric cylinder rod can be found in [19]. However, TE/TM mode in dielectric rod has cut-off frequencies which prevents dielectric rod from wide bandwidth operation. However, one of the important characteristics of dielectric rod is hybrid mode that have no cut-off frequencies. Often it is referred to a dipole-mode because an electrical dipole has similar patterns. More detail discussion about the dipole-mode characteristics is found in [59]. The ratio of energy inside and outside the dielectric rod is also determined in [20]. Determination of diameter of dielectric cylinder can be obtained from the energy ratio that hold electromagnetic field inside the rod without dispersion in operation frequency. To be a dielectric rod antenna, termination of dielectric rod has to be done properly to radiate the electromagnetic waves into the desired direction.

More previous works on dielectric rod antenna have been published and can be found in [20, 21]. They are named dielectric rod antenna as polyrod antenna because material manufactured for dielectric rod is from polystyrene. Near/far field patterns from the various shapes of polyrod antenna were reported in [22, 23, 60]. However, the focus of previous works is development of dielectric rod antenna for narrow band applications. In other words, feed structure of the dielectric rod has narrow bandwidth characteristic. In this chapter, a new dielectric rod probe (DRP) antenna implements broad bandwidth feed structure and support hybrid mode in dielectric cylinder waveguide.

In this chapter, a 1.5 inch DRP prototype is designed. It was found that the 1.5 inch DRP was suitable for the desired frequency range (2 - 18 GHz). The design specifications and characteristics of the 1.5 inch DRP is the focus of this chapter. The measured result from the 1.5 inch DRP prototype is demonstrating that the novel antenna design is for the near-field probing application in that it provide broad bandwidth, dual-polarization, and low RCS. The design details are provided in this chapter along with measurement result
associated with important antenna characteristics such as VSWR and far-field radiation patterns.

4.1 A UWB DRP Design Principles

![Diagram of DRP antenna](image)

Figure 4.1: Basic configuration of a DRP antenna

The DRP design contains the feed antenna, waveguide, and radiation sections as illustrated in Figure 4.1. The feed antenna section has a conical shape whose cone angle is chosen in conjunction of the geometry of the feed antenna arms to provide the desired 100 Ω surge impedance. A smooth transition from the feed antenna section to the waveguide section limits the maximum cone angle in practice. The waveguide section provides sufficient isolation between the launcher section and radiation section to allow evanescent modes to decay. Thus, a minimum of two wavelengths is usually chosen. The radiation section generates propagation by removing the waveguide material. The geometry of the radiation section affects the radiation efficiency, patterns, and phase center. The new DRP design differs from the conventional polyrod in two aspects. First, a broadband antenna design replaces the metallic OEWG that has relatively narrower bandwidth. Second, the radiation section is much shorter to provide a broader beamwidth and much less dispersion.
4.2 UWB-DRP Geometry

4.2.1 Feed antenna section

Each pair of feed antenna arms is connected to the center conductor of a pair of balanced 50 Ω coaxial cables as shown in Figure 4.2. The combined impedance of each pair of coaxial cables is 100 Ω. The outer conductor of each cable is tapered to improve the impedance matching. A exponentially tapered resistive film is attached to the end of each feed antenna to avoid the antenna ringing on the arm such that broadband operation is ensured at the price of a less efficiency.

4.2.2 Waveguide section

The length of waveguide section should be at least two wavelengths, approximately 5.4 inches at 3 GHz for dielectric rod with refractive index of 1.48. In order to effectively guide the electromagnetic energy, the diameter of the rod needs to be greater than one wavelength, approximately 2.7 inches at 3 GHz or 1.6 inches at 5 GHz. The current DPR prototype has
a diameter of 1.5 inches. This means that the energy leakage is expected at frequency below 5 GHz. The length of the current dielectric rod is approximately 14 inches.

4.2.3 Radiation section

![Graph of cosine profile for DRP radiation section](image)

Figure 4.3: The radius of cosine profile for DRP radiation section

In the radiation section, the diameter of the rod is gradually tapered down to less than one wavelength such that it is no longer sufficient to confine the energy to the rod. The lowest frequency components begin to radiate sooner at the beginning of the size taper due to the longer wavelength. The higher frequency components remain guided until the size of the rod becomes less than one wavelength. Thus, the physical location of the radiation varies with frequency and the tapered length should be kept as short as possible. The trade-off is a shorter taper may increase the reflection between the waveguide and radiation sections. The taper profile also affect the radiation patterns. The FDTD simulation
indicated that a double cosine profile as shown in Figure 4.3 could provide a much broader beam width. Therefore this profile was selected in the current DRP prototype. Numerical result for the improved radiation section design is discussed following section.

4.3 Numerical result for improved radiation section

4.3.1 Numerical modeling

Using Yee cells, dielectric rod and absorber were modeled in computational domain. For a dielectric constant, 2.2 was applied for DRP region. In microwave absorber modeling, 0.1 $S/m$ of conductivity was applied. Differential Gaussian pulse which contains from 2 to 18 GHz frequency spectrum was implemented at the tip of feed section. Unit length (0.68 mm) of cubic cell was determined based on the wavelength of highest frequency and dielectric constant. Figure 4.4 represents a three dimensional DRP geometry that has double cosine profiles at the radiation section. For the curved surfaces, stair step method was used for rectangular grid domain.
Figure 4.4: Cosine curve applied DRP antenna geometry description
4.3.2 Near field distributions

From the simulations, time history of down-range ($x-z$ plane in Figure 4.4) is recorded in order to investigate the possible diffractions and radiation center. Frequency domain is also obtained by fast Fourier transformation. Figure 4.5 and Figure 4.6 show the down-range near field distributions at different frequencies. Result shows the beginning stage of spherical wave propagation at the radiation tip section. Also, it is indicating that the radiation center does not migrate along the tip section in different frequencies, since the length of the radiation section is electrically short. It is noted that the field along the outside DRP waveguide section is due to the point source excitation which create possible higher mode.

![Normalized Electric field distributions for DRP with double cosine profile at f = 5 GHz](image)

Figure 4.5: Normalized Electric field distributions for DRP with double cosine profile at f = 5 GHz
Figure 4.6: Normalized Electric field distributions for DRP with double cosine profile at f = 8, 10, 12, 15 GHz
4.3.3 Performance comparison for different tip profiles

Before the double cosine profile was determined, several different radiation section candidates are simulated and compared the instantaneous near field (1 inch away from the tip). As shown in the Figure 4.7-(a), concave and convex cases with different angles are considered for the beamwidth study. The tip angles are defined by

- $\theta_{tc} =$ tip angle for concave (0, 30, 90, and 120 degrees)
- $\theta_{tv} =$ tip angle for convex (0 and 90 degrees)

And Figure 4.7-(b) shows the observation points along the spherical arc at 1 inch away from the tip. Thus, the time domain instantaneous near field can be compared for different cases.

Figure 4.8 - (a) compares the near field distributions for $\theta_{tc} = \theta_{tv} = 0^\circ$ and $\theta_{tv} = 90^\circ$ cases. In order to radiate the electromagnetic energy, radiation section needs to be removed. Introduction of abrupt discontinuity as shown in the figure ($\theta_{tc/tv} = 0^\circ$), provide the similar beamwidth. However, one of the requirements for the near field probe is minimum interaction between the AUT and the probe, and flat radiation section might have significant diffraction compared to the convex case. Thus, $\theta_{tv} = 90^\circ$ satisfies the requirements.

For the reference, concave cases ($\theta_{tc}$) are also studied for the beamwidth improvement. Comparison result is shown in the Figure 4.8 - (b). Method of removal at the radiation section is reversed. As expected, rim of the cylindrical rod supports diffractions that result slightly wider beamwidth in $90^\circ$ and $120^\circ$ cases. Due to the strong diffractions at the radiation section, convex rod radiation section DRP is better choice.
Figure 4.7: Side view of convex and concave case profile definitions and near-field observation arc
Figure 4.8: Instantaneous near field distributions comparisons for different cases

(a) Convex case comparison

(b) Concave case comparison

(c) Quadratic vs. others
The final trial before the double-cosine profile was quadratic profile. From the Figure 4.8-(c), we found that the quadratic profile DRP provide the most wide beamwidth among all other profiles. The result from the simulations leads us a important characteristic of the radiation section design. In order to increase the beamwidth, the transition from the waveguide and radiation sections must have electrically short length and fast sloping curve while satisfying the low radar cross section (RCS) level. Therefore, a double-cosine profile is suitable for the near field probe that satisfying the important requirements. For the comparison between the quadratic profile and double-cosine profile, far-field patterns of DRP was calculated. From the Figure 4.9 to 4.11, far-field patterns for the quadratic and double-cosine profile radiation tip section DRP are compared. In operation frequency band, they provide very similar patterns. However, double-cosine profile is believed to be low RCS level due to the smooth transition between waveguide and radiation section.

Figure 4.9: Comparison of calculated far-filed pattern of dielectric rod (concave quadradic and double cosine) (1)
Figure 4.10: Comparison of measurement and calculated far-field pattern of dielectric rod (concave quadratic and double cosine profiles) (2)
Figure 4.11: Comparison of calculated far-filed pattern of dielectric rod (concave quadratic and double cosine profiles) (3)
4.4 Measured UWB-DRP Characteristics

The prototype DRP with double-cosine radiation section was built for the measurement. Picture of a novel DRP is shown in Figure 4.12. And far-field patterns are measured at the anechoic chamber in ElectroScience Lab. The measured E-plane pattern result is also compared to simulated patterns, and discussed in this section. The measured E- and H-plane patterns in entire frequency band are also presented. Another important measured characteristics such as VSWR, polarization and gain can be found in [61, 62].

![Prototype DRP with double cosine radius profile of the radiation section](image)

Figure 4.12: Prototype DRP with double cosine radius profile of the radiation section

4.4.1 Far field radiation patterns

The four frequency patterns are selected for the comparison. They are 8,10,12, and 15 GHz for E-plane far-field patterns. Figure 4.13 and 4.14 show the far-field patterns.
Overall, they are well agreed due to the realistic modeling. However, result shows the difference near the boresight direction between prediction and measurement at low frequency (8 GHz). This is because the microwave absorber has frequency dependent characteristics, absorption at low frequency is relative ineffective compared to that of the higher frequencies. It is noted that the simulated absorber model is frequency independent material that has constant values in operation frequency band.

From Figure 4.15 to 4.23, measured E- and H-plane far-field radiation patterns are plotted within the frequency band of 2 - 18 GHz. As mentioned, symmetrical E- and H-plane patterns are observed in whole frequencies. By introducing the UWB DRP, disadvantages of the narrow beamwidth and asymmetrical E- and H-plane patterns from OEWG can be overcome. Also, relative frequency-independent far-field patterns was achieved in higher frequency band (especially above 8 GHz). In lower frequency band, frequency-dependent and slightly narrower patterns were observed. This is due to the diameter of the DRP waveguide section. As it is mentioned in the beginning of the chapter, operation limit from the energy ratio for 1.5 inch diameter with dielectric constant 2.2 is about 5.3 GHz. Below 5 GHz, electromagnetic waves are started radiating from the rod section which was expected. In general, UWB DRP successfully provide the dual-linear polarizations, extremely low RCS level, broad beamwidth and symmetric E- and H-plane pattern which are all desired features for the near-field probing antenna.
Figure 4.13: Comparison of measurement and calculated far-filed pattern of DRP
Figure 4.14: Comparison of measurement and calculated far-field pattern of DRP
Figure 4.15: Measured far-field E- and H-plane patterns (1)
Figure 4.16: Measured far-field E- and H-plane patterns (2)
Figure 4.17: Measured far-field E- and H-plane patterns (3)
Figure 4.18: Measured far-field E- and H-plane patterns (4)
Figure 4.19: Measured far-field E- and H-plane patterns (5)
Figure 4.20: Measured far-field E- and H-plane patterns (6)
Figure 4.21: Measured far-field E- and H-plane patterns (7)
Figure 4.22: Measured far-field E- and H-plane patterns (8)
4.5 Conclusion of the Chapter

Using the three dimensional FDTD model, UWB DRP antenna was developed for the near-field measurement system. Various radiation sections are applied in order to investigate the characteristics of the DRP radiation patterns. The 1.5 inch diameter was determined by the energy radio with respect to the waveguide section. The beamwidth study from the quadratic and double cosine profile showed the superior performance. It is found that the double-cosine profile in radiation section provide the broader beamwidth while suppressing the undesired diffractions which create the AUT-antenna interactions. From the far-field result, approximately 100 degree of 5 dB beamwidth was achieved from 9 - 18 GHz. In low frequency band (2 - 8 GHz), around 50 to 70 degree beamwidth was obtained. This narrower beamwidth in lower frequencies are come from the size of the diameter that limits...
the operation frequency. Overall, performance and characteristics of novel UWB DRP is superior to OEWG antenna for near-field probing.
CHAPTER 5

DESIGN OF DIELECTRIC HORN ANTENNA (DHA) FOR ANECOXIC CHAMBER FEED

It is very challenging to design a single antenna that provides a broad bandwidth, dual linear polarization, a stationary phase center and a desired beamwidth. An antenna design using only electric conducting structures usually faces undesired trade-offs among the above parameters. For instance, a TEM horn with diffraction treatments such as rolled-edge [63] or resistive loading [64] and flared ridges [65] can achieve a significant bandwidth but suffers from asymmetric patterns in the E- and H-planes. Dual-polarization is difficult for these designs due to the interactions between the two polarizations. The phase center of the ridge-horn design also varies with frequency. The different boundary conditions in E-plane and H-plane imposed by the conducting structures force the field distribution at the aperture to be quite different in the two planes. Various techniques have been developed to improve the symmetry of these field distribution such that a more symmetric radiation pattern can be achieved [66, 67]. The use of corrugated walls and dielectric loading are the two well-known examples. However, they usually have relatively narrow bandwidth. Planar spiral antennas [68] and sinuous antennas [69] can also achieve 9:1 bandwidth with a fairly symmetric radiation pattern. Dual-linear polarization spiral and sinuous antennas
are available commercially. The major disadvantages of these antennas are the dispersion and fixed radiation patterns that may not be suitable for all applications.

Using a solid dielectric body as the main radiator can be found in literatures. For instance, dielectric rod antennas developed in the 40’s [21, 70] used a metallic waveguide to excite guided modes inside a solid polystyrene rod with either circular or square cross-sectional area. This type of antenna was sometimes referred to as a polyrod antenna. By gradually tapering down the cross-sectional dimension of the rod, the guided waves begin to radiate away from the rod. The radiation pattern of a polyrod antenna is governed by the radiation from launcher-rod mismatch, tapering of the rod dimension and the radiation from the end of the rod. Very high directivity could be achieved with a proper combination of these three parameters. Polyrod antennas have narrow bandwidth due to the waveguide launcher. This is the same problem for solid dielectric horn antennas fed by waveguides [67]. Recently, a new broadband dielectric rod antenna design was developed for the detection of buried landmines [17, 27]. The improved bandwidth was achieved in this design using a special well-matched, broadband wave launcher. For wide-beam applications, the rod dimension is kept constant until the end where the radiation occurs. This broadband design motivated the new dielectric horn antenna (DHA) design to be discussed here.

Very comprehensive discussions about the characteristics of a solid dielectric horn antenna can be found in [67]. Many important characteristics such as the field distributions inside the dielectric cone and the far-field radiation pattern were discussed for the HE_{11} mode. For instance, it was pointed out that the radiation pattern from a circular dielectric cone has circular symmetry with respect to boresight. The impact of feed antenna and higher order modes were also addressed. However, the modal expansion approach does not provide insight into the original wave mechanism that established the modes. Some
of the mechanism may not be desirable at all and should be avoided via a careful design. The lateral waves that are excited by the air-dielectric interface affect the radiation pattern significantly. Almost all of the previous dielectric horns were excited by metallic waveguides and thus having relatively small bandwidths. Also, most of the previous dielectric horn designs have single polarization. The proposed new DHA design provides superior bandwidth, dual-polarization capability, and improved pattern stability.

The basic dual-polarization DHA design includes the feeding cables, a broadband wave launcher, the dielectric body and the external absorber. Each component will be discussed in the following sections. In Section 5.1, the basic configuration of the DHA design is introduced. The function and importance of each component will also be briefly discussed. Numerical modeling and performance optimization process are discussed in Section 5.2. Control parameters and their effects are presented. Section 5.3 presents some radiation characteristics such as radiation patterns, beam widths, and efficiency based on both numerical simulated data and measured data from DHA prototypes. Finally, the concluding remarks are given in Section 5.4.

5.1 DHA Design Concept

The basic DHA design is composed of three parts: a feed antenna, a dielectric body and external absorber. A more detail geometry of each component can also be seen in Figure 5.1-(a). The launcher contains two launch arms positioned on opposite sides of the horn. Each launcher arm was composed of a triangular conducting plate terminated with a tapered resistive film. The plate angle was selected to provide an input impedance of approximately 100 $\Omega$. This impedance is also a function of horn angle and dielectric constant [33].
Figure 5.1: 130°-DHA and its launcher geometry
The launcher was electrically connected to a pair of 50 Ω cables that provides a balanced broadband excitation with an effective impedance of 100 Ω. A section of resistively tapered film is attached to the conducting plate for smoother current termination. The width and the length of the resistive film were kept to a minimum to avoid modifying the dielectric-air boundary condition. Such a resistive termination suppresses the antenna ringing and increases the bandwidth at the price of a slightly lower efficiency.

The dielectric material chosen for the DHA should have low loss in the frequency band of interest. A higher dielectric constant reduces the antenna size and increases the antenna efficiency with a tradeoff being a larger reflection at the radiation aperture. Such an undesired reflection could be minimized via a proper dielectric tapering along the radial direction to provide a smooth impedance transition [71]. The Polymethyl-Methacrylate which is an acrylic plastic, was used for our prototype based on its low loss, low cost, ease of fabrication, good UV stability, good thermal stability, low water absorption and high dielectric strength. The refractive index of a typical clear acrylic plastic is approximately 1.48 over a wide frequency range. The horn angle should be chosen based on the desired beamwidth, frequency range, and input impedance. A larger horn angle gives a wider beam width and requires wider launch arms to maintain an input impedance of 100 Ω. The latter may become a problem or increases the cross coupling in a dual polarization design. If the horn angle is too small, the cross-sectional dimension would be much less than one wavelength in the dielectric, and the majority of the energy will be in the external medium [21].

As mentioned, the external absorber reduces the undesired lateral waves within the dielectric body. It also reduces possible leakage radiation arises from the feed point. As expected, this would lower the overall antenna efficiency due to the energy absorption. A
Figure 5.2: Feed antenna arm geometry and its resistance profile of the 130° DHA prototype

major requirement of the absorber is its uniformity and its contact to the dielectric surface. Ideally, the dielectric constant of the absorber should be kept lower than that of the dielectric horn.

Figure 5.1-(b) shows an example of a 130°-DHA prototype consisting of a solid circular dielectric cone with a radius of 3.5 inches. Part of the absorber is removed to reveal the details. The dielectric constant and conductivity of the absorber at 10 GHz are approximately 1.8 and 0.22 S/m, respectively. The geometry of the launcher arm and the resistive profile are also shown in Figure 5.2. The antenna characteristics of this prototype from the numerical modeling will be explored in the following sections.
5.2 Numerical Analysis and Further Optimization

Based on the parameters discussed in the previous section, a three dimensional model is reconstructed for the full wave analysis. The finite difference time domain (FDTD) method was employed for the easy handling of inhomogeneous medium such as absorber, solid dielectric and resistive film etc. Detailed features of the antenna include coaxial cables, conductive feed antenna arms, resistive film and the solid dielectric cone, and they are modeled in a realistic manner. The characteristic impedance of the coaxial cable is also set to be 100 Ω. Since the numerical study is applied to a uniform rectangular grid, the curved surfaces of the geometry are approximated by a stair stepping scheme. For the resistive
strip modeling, the resistive values along the strip is converted to effective conductivity by
the relationship \( \sigma = 1/(R\tau) \) where \( \tau \) is the thickness of the film. The external microwave
absorber is located on the top of the DHA. It is modeled as a homogeneous lossy medium
with relative dielectric constant and conductivity of 1 and 0.25 S/m, respectively. Figure 5.3 depicts the three-dimensional FDTD grids for DHA simulation.

In order to verify the numerical model, measured and computed far-field patterns are
compared. In the FDTD computation, far-field patterns are obtained by the near-to-far-
field transformation [72]. Far-field patterns in E-plane are plotted in Figure 5.4-(a) and
(b). Measured and computed patterns have excellent agreement in the entire operation
frequency band. It is noted that 3 dB dips near the boresight direction at the frequency
above 6 GHz are observed. The 3-dB dips are undesired features and could be fixed through
virtual test with the aid of simulations. H-plane far-field patterns from the measurement
and the computation are also shown in Figure 5.4 (c) and (d). Measured and computed
patterns are in good agreement in the overall frequency band. Fairly symmetric E- and H-
plane patterns are observed from the measurement and computation result. Below 3 GHz,
the dimensions of the dielectric body are small compared to the wavelength, and a broad
radiation patterns are observed.

One of the advantages of the numerical simulations is providing a relatively easy op-
timization process with simple changes in the computation variables instead of building
prototype antennas. In this case, performance control variables are
Figure 5.4: Comparison between measured and computed E and H–plane far-field patterns of prototype 130° horn angle DHA
1. feed antenna

2. dielectric horn angle and cross section

3. resistive profile on the strip

4. microwave absorber.

First, characteristics of the DHA with and without feed antenna are investigated. The design of the feed antenna is the most challenging part and plays the most important roll in the final DHA performance. The feed antenna should be able to couple the UWB electromagnetic energy from feeding cables to the dielectric horn and excite the desired fundamental mode inside the dielectric cone. Since the feed antenna resembles a resistively loaded V dipole, it also has its own radiation properties such as efficiency and patterns. For instance, a longer launcher arms may have a higher efficiency but also pull the radiation energy along the arm direction just like a V-dipole. This could cause a dip in the main beam of the overall pattern in wide beamwidth applications. On the other hand, a shorter launcher arm minimizes the launchers impact on the overall patterns, and it behaves more like a point source excitation. However, a short launcher arm also leads to lower efficiency just like a short dipole. The resistive taper also affects both efficiency and pattern and requires optimization for a given operational band and launcher length. In order to better understand the effect of the launcher, near/inner-field distributions are compared with and without feed antenna in DHA.

In one case, a point source was used to excite the field into the dielectric horn from the vertex. In the other case, the radiated fields were recalculated for the horn excited by the feed antenna. Only x-component electric fields (co-polarization) are plotted from Figures 5.5 to 5.14. The left and the right side figures are the comparison pair with/without
feed at the operation frequency. Similar patterns are observed below 10 GHz where the lengths of the feed antenna arms are relatively short compared to the wavelength. Above 10 GHz, the launcher significantly changes the field distributions and becomes non-uniform compared to those excited by a point source. It is also observed that the presence of the feed antenna reduces the side- and back-directed radiation and thus reduces the front-and-back ratio and side lobe levels, which are all desirable features.

This launcher effect study indicates that the longer launcher arm requires a great deal of care to make sure that there is a gradual termination of the current along the surface of the conductive arms to avoid the significant diffractions from the end, which deteriorates the far-field patterns. Thus, the feed antenna arms with 1/8 inch and 18 degree plate angle and a new exponential resistive profile strip (which will be discussed shortly) are found to be optimal with a newly developed DHA.
Figure 5.5: Inner field magnitude distribution comparison for launcher effect from 120°-DHA (1)
Figure 5.6: Inner field phase distribution comparison for launcher effect from 120°-DHA (1)
Figure 5.7: Inner field magnitude distribution comparison for launcher effect from 120°-DHA (2)
Figure 5.8: Inner field phase distribution comparison for launcher effect from 120°-DHA (2)
Figure 5.9: Inner field magnitude distribution comparison for launcher effect from 120°-DHA (3)
Figure 5.10: Inner field phase distribution comparison for launcher effect from 120°-DHA (3)
Figure 5.11: Inner field magnitude distribution comparison for launcher effect from 120°-DHA (4)
Figure 5.12: Inner field phase distribution comparison for launcher effect from 120°-DHA

(a) f=13GHz (w/o launcher)  (b) f=13GHz (w. launcher)

(c) f=14GHz (w/o launcher)  (d) f=14GHz (w. launcher)

(e) f=15GHz (w/o launcher)  (f) f=15GHz (w. launcher)

92
Figure 5.13: Inner field magnitude distribution comparison for launcher effect from 120°-DHA (5)
Figure 5.14: Inner field phase distribution comparison for launcher effect from 120°-DHA

(5)
Second, the investigation of horn angle with the feed antenna is considered for the improvement of the 3-dB dips from the previous prototype design. Although the analytical solution of the dielectric cone geometry is well known, the closed form solutions for our DHA model which include feed, resistive strip, and absorber is very difficult to obtain due to the geometrical complexity. Thus, numerical computation is essential to predict the performance. In this case, the modified cone angles are $\theta_h = 130$, 120 and 110 degrees.

Far-field patterns are calculated and plotted in Figure 5.15. As mentioned, the feed antenna has its own radiation pattern as a V-dipole. It is found that the radiation patterns are also depend on the horn angle. The dip at the center starts to disappear when the horn angle is 100 degrees with the 1/8 inch launcher. It is noted that the 3-dB beamwidth is more than 60 degrees.

Figure 5.15: Far-field patterns for three different horn angles at $f = 14$ GHz
Instead of increasing the conductivity in the external microwave absorber, reshaping the conical boundary condition is another approach to remove the undesired lateral waves. In addition to a curved surface, a pyramidal cross section geometry is employed instead of circular cone because rectangular waveguide has more bandwidth than the circular waveguide in practice [73]. It is noted that the initial horn angle is now 110 degree instead of 130 degree, because our numerical study indicated that the far-field patterns is also affected by the horn angles such that the beamwidth is reduced for smaller horn angle as shown in the Figure 5.15.

Figure 5.16 depicts the side view and three-dimensional uniform grid of the newly designed 110°-dielectric horn antenna with curved sections (DHAc). As shown in the figure, the excitation region is defined by the pyramidal cross section and the curved surface was approximated by five linear sections for easy construction. The four 1/8 inch arms of the feed antenna and the resistive cards are installed at the apex of the horn.

As shown in the previous prototype far-field patterns, the 3-dB dips near the boresight should be eliminated within the operation frequency band. Thus, a new resistive profile is introduced. The conductivity of the new resistive card is higher than that of previous film in order to increase the effective length of the feed antenna arms. Resistance profiles are measured, and its profile is shown in Figure 5.17. The new resistance of the implemented section is from 5.9 to 116.82 Ω in 2.75 inch length.

Finally, microwave absorber model is considered for better characterization of the DHA. The conductivity and permittivity of the absorber model are based on the averaged values from the measurement. The absorber is a dispersive material that has frequency dependent conductivity and permittivity. Since the non-dispersive FDTD method is only allows constant values for the material properties, more accurate modeling is required to achieve
Figure 5.16: Proposed solid dielectric cone shape
better results. Thus, sub-frequency band simulations were chosen in the absorber analysis. It is noted that for the dispersive material analysis, frequency-dependent FDTD using recursive convolutions can be another approach [74]. However, the frequency-dependent FDTD was not used in this paper due to the much longer computation time and larger storage. For fixed conductivity of 0.25 S/m, relative permittivity varies from 1.0 to 2.0. Figure 5.18-(a) plots the half power beamwidth vs. permittivity at $f = 14$ GHz. As indicated in the result, the beamwidth is proportional to the permittivity value except at the highest permittivity value. This is explained by the boundary transition between the dielectric horn and the absorber material. When the permittivity of the absorber is larger than that of the dielectric cone, the beamwidth is decreased. Thus, the permittivity in the absorber should be lower
than that of the dielectric cone body, and it is important that the sub-frequency simula-
tion should include the accurate permittivity value for the DHA. The conductivity is varied
from 0.1 to 0.3 S/m with fixed a permittivity of 1.0 in Figure 5.18-(b). The interference
is present near the boresight angles (±30 degree) at f = 14 GHz. Since the conductivity
was introduced for the suppression of the lateral waves at the boundary of dielectric cone
body, the far-field patterns are affected by the amount of undesired interferences along the
boresight direction. Thus, the result indicates that higher conductivity absorber reduces the
interferences from the dielectric absorber boundary.

Far-field patterns are calculated for the optimized geometry and shown in Figure 5.10.
From the result, great improvement was made in terms of the 3-dB dips at the center com-
pared to the previous design. The figures show symmetric E-and H-plane patterns and fre-
quency independent characteristics. The 130°-DHA is now optimized into the 110°-DHA$^c$
where the character $c$ indicates the curved edge geometry. Thus, the prototype antenna
geometry, which is derived from the numerical results is constructed. In the next section, it
will be measured to determine the VSWR, antenna gain, and far-field patterns.

Inner field magnitude and phase distributions are also calculated and shown in Fig-
ures 5.20 to 5.25. Overall, no significant amount of diffraction from the end of dielectric
cone can be found over the frequency of interest. The phase distribution indicates that the
fixed radiation center is at the apex of the cone.
Figure 5.18: Study of permittivity and conductivity of the external absorber at $f = 14$ GHz
Figure 5.19: E- and H-plane far field patterns of 110°-DHA C from FDTD calculation
Figure 5.20: Inner field magnitude distribution comparison for launcher effect from 110°-DHAc (1)
Figure 5.21: Inner field magnitude distribution comparison for launcher effect from 110°-DHAc (2)
Figure 5.22: Inner field magnitude distribution comparison for launcher effect from 110°-DHAc (3)
Figure 5.23: Inner field magnitude distribution comparison for launcher effect from 110°-DHAc (4)
Figure 5.24: Inner field magnitude distribution comparison for launcher effect from $110^\circ$-DHA$^c$ (5)
Figure 5.25: Inner field magnitude distribution comparison for launcher effect from 110°-DHAc (6)
5.2.1 Specification of improved $110^\circ$-DHA$^c$

Feed antenna

<table>
<thead>
<tr>
<th>Length</th>
<th>1/8 inch</th>
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</thead>
<tbody>
<tr>
<td>Plate angle</td>
<td>34°</td>
</tr>
</tbody>
</table>

Table 5.1: Feed antenna profile of $110^\circ$-DHA$^c$

Microwave absorber model

<table>
<thead>
<tr>
<th>Relative Permittivity ($\varepsilon_r$)</th>
<th>1.0</th>
</tr>
</thead>
<tbody>
<tr>
<td>Conductivity ($\sigma$)</td>
<td>0.25 S/m</td>
</tr>
</tbody>
</table>

Table 5.2: Microwave absorber property

Resistive card

<table>
<thead>
<tr>
<th>Overlay length</th>
<th>2 mm</th>
<th>Length of R-card</th>
<th>2.75 inch</th>
</tr>
</thead>
<tbody>
<tr>
<td>Length in inch</td>
<td>R ($\Omega/\square$)</td>
<td>$\sigma$ (S/m)</td>
<td>Length in inch</td>
</tr>
<tr>
<td>0</td>
<td>5.9091</td>
<td>203.8925</td>
<td>1.5748</td>
</tr>
<tr>
<td>0.1969</td>
<td>9.2424</td>
<td>130.3575</td>
<td>1.7717</td>
</tr>
<tr>
<td>0.3937</td>
<td>12.7273</td>
<td>94.6644</td>
<td>1.9685</td>
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<tr>
<td>0.5906</td>
<td>16.3636</td>
<td>73.6278</td>
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</tr>
<tr>
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<tr>
<td>1.3780</td>
<td>24.8485</td>
<td>48.4866</td>
<td></td>
</tr>
</tbody>
</table>

Table 5.3: Tapered resistive card profile of $110^\circ$ DHA$^c$
Dielectric body

Feed section angle | 110° | Permittivity ($\varepsilon_r$) | 2.2 |
<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Diameter</td>
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<td>Hight</td>
<td>3.5 inch</td>
</tr>
<tr>
<td>$x$ (m)</td>
<td>$y$ (m)</td>
<td>$x$ (inch)</td>
<td>$y$ (inch)</td>
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<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>0.0009</td>
<td>0.0009</td>
<td>0.0327</td>
<td>0.0327</td>
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<td>0.0017</td>
<td>0.0017</td>
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<td>0.0653</td>
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<td>0.0034</td>
<td>0.0025</td>
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<td>0.2288</td>
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<td>0.0067</td>
<td>0.0050</td>
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<td>0.1960</td>
</tr>
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<td>0.0075</td>
<td>0.0058</td>
<td>0.2941</td>
<td>0.2287</td>
</tr>
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<td>0.0067</td>
<td>0.3595</td>
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<td>0.0075</td>
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<td>0.2941</td>
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<td>0.4575</td>
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<td>0.0133</td>
<td>0.0092</td>
<td>0.5229</td>
<td>0.3594</td>
</tr>
<tr>
<td>x (m)</td>
<td>y (m)</td>
<td>x (inch)</td>
<td>y (inch)</td>
</tr>
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<td>-------</td>
<td>----------</td>
<td>----------</td>
</tr>
<tr>
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<td>0.0100</td>
<td>0.5555</td>
<td>0.3921</td>
</tr>
<tr>
<td>0.0158</td>
<td>0.0108</td>
<td>0.6209</td>
<td>0.4248</td>
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<td>0.4575</td>
</tr>
<tr>
<td>0.0191</td>
<td>0.0125</td>
<td>0.7516</td>
<td>0.4901</td>
</tr>
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<td>0.0208</td>
<td>0.0133</td>
<td>0.8170</td>
<td>0.5228</td>
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<tr>
<td>0.0224</td>
<td>0.0141</td>
<td>0.8823</td>
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<td>0.0241</td>
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<td>0.0258</td>
<td>0.0158</td>
<td>1.0130</td>
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Table 5.4: Characteristics and SBH curvature profile applied in 110°-DHAc
5.3 Measurement Result of Improved 110°-DHAc

5.3.1 New prototype antenna construction

Based on the dimension from the numerical model, a new prototype antenna is constructed. Figure 5.26 shows the features of the 110°-DHAc. For the fixture, a 7 inch PVC pipe is attached so that the antenna can be pointed toward the boresight direction without difficulty. As a balun, 180° hybrid transformer is connected to the 100 Ω coaxial cable. Far-field patterns and antenna gain are measured in the anechoic chamber at the Ohio State University to verify the result from numerical simulations.

![Figure 5.26: New prototype design of 110° DHAc](image)

5.3.2 VSWR

Figure 5.27 plots the measured VSWR for the prototype 110°-DHAc from 2 to 18 GHz. A good impedance matching (VSWR < 1.5) is observed. The slow varying curve also indicates little antenna ringing. Recall that the impedance (impedance at the feed point) of a DHA is determined by the angle of feed antenna s conducting plate, the horn angle, and the dielectric constant.
Figure 5.27: VSWR of the $110^\circ$-DHAc prototype
5.3.3 Radiation patterns

Figure 5.28 plot the measured E-plane and H-plane far-field radiation patterns. Both of the plane patterns remain very stable between 4 GHz and 18 GHz. At 2 GHz, the dimensions of the dielectric body become small compared to the wavelength, and a broad radiation pattern are observed. Figure 5.29 compares the measured 10-dB beamwidth as a function of frequency for the previous and the current design. It is found that the improvement of the 3-dB dips yields the slight asymmetry of E-and H-plane far-field patterns. It was expected from the simulated result. However, the beam widths in both planes are fairly symmetric and stable centered around 110 degrees. The E-plane is slightly wider than the H-plane. The 3-dB beam widths were guaranteed to be at least 60 degrees over the operating frequency.

5.3.4 Antenna gains

Figure 5.30 plots the measured antenna gain. The variation of the gain over the 2-18 GHz band is approximately 5 dB. The negative gain is certainly due to the use of the absorber and the tapered resistive films in the feed antenna. Recall that the absorber was used mainly to control the undesired lateral waves to achieve broadband radiation patterns. Other approaches to controlling the lateral waves are currently under investigation. The level of depolarization of the radiated field was found to approximately 30 dB over the 2-18 GHz frequency range.
Figure 5.28: Measured E- and H-plane far-field patterns of new prototype 110°-DHAc
Figure 5.29: Measured 10-dB beamwidth comparison for a previous and improved DHA designs

Figure 5.30: Antenna gain of 110° DHAc
5.4 Conclusion of the Chapter

A broadband dielectric horn antenna (DHA) design concept was presented. The measured data from a DHA prototype with a horn angle of 130 degrees clearly demonstrated many desirable characteristics of the new DHA design. However, 3-dB dips at the bore-sight direction are the unwanted feature. Thus, using the FDTD method, realistic models are constructed for characterization and optimization. Previous problem was solved by applying curved surface of the dielectric horn body and the thin resistive films with lower resistance profile. A new prototype DHA with 110 degree horn angle was constructed and measured far-field patterns. They have following advantages over the conventional compact range feed. First, a VSWR of less than 1.5 was achieved from 2 to 18 GHz with a relative flat response. Dual-linear polarization capability was achieved with a depolarization level less than 30 dB across the band. The measured radiation patterns are similar in E- and H-plane with a fairly stable 10-dB beam width of approximately 120 degrees. Our study indicated that the beam width can be controlled by the horn angle conveniently. Currently, external absorbers were used to control the undesired lateral waves. This results in a lower antenna gain. Other approaches of controlling the lateral waves or avoiding the excitation of the lateral waves are being investigated to overcome this issue. It was also found the uniformity of commercial flexible absorbers could cause pattern distortion.
Antenna pattern and target radar cross section (RCS) measurements are often performed in large open/closed ranges such as anechoic chambers, ground-bounce ranges or compact ranges [75, 76, 77]. A tapered chamber is an economical alternative for such measurements when the size of the antennas or targets to be tested is not too large [78, 79, 80, 81]. King [82] characterized the tapered anechoic chamber by comparing it with a rectangular anechoic chamber and showed that the tapered chamber has better phase error in terms of multi-path reflections effects. There are several tapered chamber configurations available as commercial packages [83]. The most common configuration contains a tapered section where the fields are launched and a main rectangular chamber where the test antenna or target is located (see 6.1). The tapered section includes a conical tapered feed section. Different absorbers are used on the inner walls everywhere to reduce wall reflections to prevent contamination of the incident fields at the test zone. Note that, ideally, the wave launcher should be located at the apex of the conical section to avoid reflections from walls inside the feed section. However, in practice, most people simply place a commercial antenna such as a horn or log periodic antenna within the feed section for wave-launching purpose. The physical size of these commercial antennas are often too big and causes the actual radiation center to be significantly away from the desirable apex point, resulting in
much poorer bandwidth and field performance. Therefore, there is a need to develop an improved wave-launching mechanism that conforms to the original tapered chamber concept with its radiation center located near the apex of the conical feed section. In addition, dual-linear polarization capability and lower operational frequencies are also desired in many applications.

In this paper, we introduce a novel wave-launching design for feeding a tapered chamber with the following features:

1. wider bandwidth (from below 200 MHz to above 3 GHz);
2. cleaner spherical waves;
3. fixed radiation center located near the desirable chamber apex;
4. dual linear polarization.

This new design was evaluated using the finite difference time domain (FDTD) method due to its flexibility in handling inhomogeneous materials. For accurate simulations, full-scale detailed models of all structures and materials involved were created. Such simulations can be used to analyze the field distributions and identify possible sources of contamination such as reflections or diffractions inside the feed section. A prototype was also constructed and tested as verification of the design [84, 85]. Both simulation and measurement results have demonstrated the superior performance of this new feed design.

This chapter is organized as follows. The basic mechanism of the tapered chamber and geometrical descriptions are provided in Section 6.1. In Section 6.2, numerical configurations of individual structures are discussed. Investigations and development of the UWB wave launcher are given in Section 6.3. In this section, preliminary numerical results
demonstrating the direction of future improvements for the tapered chamber are also dis-
cussed. Conclusion and discussions of the tapered chamber study are found in Section 6.4.

6.1 Basic Tapered Chamber Operation Principle

Figure 6.1: Two dimensional common tapered chamber layout

Figure 6.1 illustrates a common tapered chamber geometry whose optimal dimensions
are recommended in the IEEE Standard [86] as being

\[ H_t < \frac{\lambda R}{4h_r} \]  

(6.1)
Figure 6.2: Common tapered chamber feed section and feed antenna
where $\lambda$ is the free space wavelength, $R$ is the separation between source and test antennas, $h_r$ is the perpendicular distance from the fixed test antenna to the chamber wall, and $h_t$ is the perpendicular distance from the source antenna to the chamber wall. Figure 6.2 shows more details of a typical conical feed section with a wave-launching antenna placed inside as is commonly done in practice. The conical angle is 30 degrees and 1.8 meter high with a 0.96 meter aperture diameter. The thickness of the absorber on the conducting wall is 15.24 cm. The geometry and electrical properties of such an absorber is very important and needs to be carefully determined. For our numerical models, the permittivity and conductivity of the actual absorber was measured in a waveguide and are shown in Figure 6.3.

6.2 Numerical Modeling of Tapered Chamber Feed Antenna

Since the FDTD method is employed for analysis, three dimensional uniform cubic grids were created for the feed section, feed antenna, absorber and cables. For curved portions of the geometry, stair stepping was applied. Outer metallic walls and conductive antenna arms are modeled as a perfect electric conductor (PEC). Averaged values between 100 MHz to 3000 MHz band of permittivity (1.87) and conductivity (0.22 S/m) are chosen for computation.

6.2.1 Modeling of feed section

Figure 6.2-(a) shows the three-dimensional FDTD model of the feed section. The feed section is conical shaped with a conducting wall backing covered with a layer of microwave absorbing material.6-inch in thickness. The dimensions of the model are also shown in the same figure.
Figure 6.3: Wall absorber samples and test setup - measured permittivity and conductivity curves
6.2.2 Modeling of microwave absorber

In order to model the microwave absorber accurately, samples are obtained from the Microwave cumming Inc. and measured at the ElectroScience Laboratory. Three different lengths of samples were measured and averaged for FDTD analysis. The permittivity and conductivity of the absorber used in the feed section were determined using the waveguide transmission method. They are shown in Figure 6.3 - (a) and (b). Measured values are also plotted in Figure 6.3-(c) and (d), respectively. During the FDTD simulation the average values within the frequency band of interest were adopted.

6.2.3 Modeling of wave-launching arm

Figure 6.4: UWB dual linear polarized feed design for the given tapered chamber feed section
Figure 6.4 shows the cutout view of the two wave-launching arms imbedded inside the absorber. Each pair of arms is responsible for launching fields with electrical fields polarized on the planes occupied by arms. The absorber boundary is indicated by the shadowed region. Each pair of arms is fed with a pair of balanced $50 \, \Omega$ coaxial cables. During the simulation, we used a differentiated Gaussian pulse waveform whose spectrum was mainly between 200 MHz and 3 GHz. Note that each pair of arms is similar to a resistively terminated V-shaped bowtie dipole for its broadband property \cite{87, 88}. Each arm has a linear section, blending section, and an elliptically curved section. In the linear section, the plate angle of each arm is $20^\circ$, and the opening angle is $30^\circ$, same as the angle of the conical feed section. The total length of antenna is 1.23 m which is equivalent to approximately 2 wavelengths at 500 MHz. In order to minimize diffraction, the transition between the linear curved section and the elliptically curved section must be done smoothly; therefore, the elliptical curve is applied using an ellipse ratio of 1:8 \cite{89}.

### 6.3 Numerical Simulation and Measurement Result

#### 6.3.1 Launched field distributions

From the FDTD simulation, the time history in the x-z plane at $y = 0$ and in the x-y plane at $z = 1.545$ m were saved and transformed to the frequency domain with the fast Fourier transform. The distribution of the magnitude of the E-plane electric field from the lowest to highest frequencies is compared to geometrical optic (GO) field results in Figure 6.5-(a). They indicate no interference from the absorber walls and embedded antenna structure. Magnitude tapers are found at the boundary region (near $\pm 14^\circ$ in x-axis) due to the presence of the wall absorber. To observe the phase distortion, the GO field was used for reference.
Figure 6.5: Comparison of E-plane magnitude and phase variations for blade feed and point source excitation cases at $z = 1.545$ m with respect to GO field.
Figure 6.5-(b) shows the difference in the phase along a radial arc where the phase $\phi$ is defined as follows:

$$\phi = \text{arg}[\text{GO}] - \text{arg}[E] \quad (6.2)$$

The results agree well to the GO field distributions from the lowest frequency of 200 MHz to 1 GHz with less than 10-degree taper near the wall absorber boundary. In the higher frequency band, more tapered phase variations are observed due to the high conductivity. It is noted that the stair stepped response in phase comes from the rectangular grid point extraction in the FDTD computation. In order to see the improvement compared to a conventional tapered chamber feed as shown in Figure 6.2, a simulation is run with a Hertzian dipole placed 0.75 m from the apex of the conical feed section. Since the radiation center is not at the apex, the interference would be more dramatic compared to the blade configuration. Figure 6.5 - (c) and (d) shows the effect of the radiation center offset, especially in the lower frequency band where the typical tapered chamber performs poorly. From the numerical result, the blade configuration with conical tapered feed section provides a more uniform magnitude and more linear phase at the aperture of the feed section.

In order to investigate the overall performance of the computed tapered chamber feed section, inner field distributions are obtained using the fast Fourier transform. From the Figure 6.6 to 6.9, co-polarized electric field magnitude and phase inside the feed section and aperture are represented. Magnitude distributions indicate the pure spherical wave propagation, and phase shows the linear variation along the $z$ direction in operation frequency band.
Figure 6.6: Inner field distributions of the tapered chamber feed section $x - z$ plane - computed result
Figure 6.7: Inner field distributions of the tapered chamber feed section $x-z$ plane - computed result
Figure 6.8: Inner field distributions of the tapered chamber feed section $x - y$ plane at $z = 5.15$ ft. - computed result
Figure 6.9: Inner field distributions of the tapered chamber feed section $x - y$ plane at $z = 5.15$ ft. - computed result.
6.3.2 Bandwidth of the Launched Field

Next, the launched power spectrum is investigated at an observation point located on axis at $z = 1.545$ m position. Both co-polarization ($E_x$) and cross-polarization ($E_y$) fields were examined. Figure 6.10 plotted the normalized co-pol. and cross-pol. Fields predicted from FDTD simulation. Note that the cross talk between the two polarizations is less than -30dB. It is observed that there is a significant power decreasing below 1 GHz. We suspected that such power decreasing might be caused by the absorber. To investigate this, five different absorber configurations as shown in Figure 6.11 were studied. It is noted that the case where the absorber was completely removed from the feed section was also considered. The resultant spectrum shown in Figure 6.12 indeed indicates that...
as the absorber surrounding the feeding area is reduced, the transmitted power increases at low frequencies significantly. Furthermore, we observed that the onset of the power drop occurred at a frequency related to the exposed length of launcher arm, indicating the antenna ringing effect due to the reflection and the entry point of the absorber. From these findings, the lowest operative frequency limit can be calculated based on the length of the feed antenna that is exposed to free space inside the feed section. Thus, from the

Figure 6.11: Four different wall absorber arrangements for transmitted power study

investigation of aperture field distributions and co-polarized transmitted field strength, a
Figure 6.12: Transmitted electric field strength for four different absorber arrangements at $z = 1.545$ m

1.23 m blade bowtie configuration should not be used in the tapered chamber construction due to the significant attenuation at the lower frequency band. However, rearrangement of absorber walls without modifying the antenna shows potential improvement.

### 6.3.3 Measurement result

Because the feed must be rigid, the prototype feed for the tapered chamber was built using an aluminum (rather than copper) plate. In order to avoid undesired reflections from the outer shield at lower frequencies due to the electrically small absorber thickness, the throat region is introduced to trap the electromagnetic waves and enable the TEM mode. The throat region is gradually opened till the end of the feed section in a similar manner.
as the TEM horn antennas. Figure 6.13 shows the prototype feed antenna geometry. For the reconstruction, next section provide the two dimensional retangular coordinate points of quad-ridged antenna arm. Also, it is designed to avoid the inductive region \( kz \ll 1 \), where \( k \) is a wave number in free space) of the conical feed section, which is specified in [90]. As mentioned in the previous section, the low frequency limit is related to the length of the

![Figure 6.13: Prototype UWB feed antenna geometry for conical feed section of tapered chamber](image)

blade arm which is exposed inside the feed section. Thus, the arms are extended to 1.8 m which is the same as the total length of the feed section of the tapered chamber. Finally, the
wall absorber arrangement of case 4 in Figure 6.11 was applied during the construction. The launcher is connected to the 100 Ω coaxial cable. As a balun, a hybrid transformer is installed. In order to solder the inner conductor of the coaxial cable with the aluminum plate, adhesive copper plates are applied at the contact. Figure 6.15-(a) and (b) show the shapes of the coaxial cable-feed antenna section. The dimensions of a part of the tapered chamber feed section in shown in Figure 6.15-(c). The other polarization arms are inserted and fixed at the orthogonal slot inside the absorber. After the completion of the entire assembly, a near field probe of the aperture was performed and shown in the Figure 6.15-(d). Employing the 1.5 inch monopole, results from the vertical and the horizontal scan have been collected. It is noted that the aperture is facing toward the anechoic chamber to prevent the undesired reflections from the building structure.

A numerical model was also created, and the E- and H-plane near field distributions with respect to the GO field distributions at $z = 1.545$ m are calculated and plotted in Figure 6.14-(a) and (b). The strong field near the edges (near $\theta = \pm 15^\circ$) is due to the exposed PEC blades. Relatively constant E-plane magnitude and tapered H-plane magnitude are observed as similar to circular waveguide $TM_{11}$ mode distribution. Phase plots are compared to GO field distributions and shown in Figure 6.14-(c) and (d). As expected, the phase in the low frequency band is greater than 0°. This is due to the migration of the radiation center as a function of frequency. Although the throat opening guaranteed TEM mode excitation, the phase center was shifted and needed to be calibrated for RCS/radiation measurements. Overall, near field distributions indicate that the prototype feed design produces a spherical wave front at the aperture of the given conical feed section.

Near field normalized magnitude and phase from 200 MHz to 3 GHz are shown in the Figure 6.20. It is noted that the time gating was applied to eliminate the potential reflections...
from the indoor structure except for the chamber response. In H-plane magnitude distributions, patterns in the operating frequency are fairly symmetrical, and no interferences from the conductive arms are observed. It is evident that the phase distributions have linear variations in the frequency domain. However, relatively asymmetrical patterns are obtained in E-plane. This is caused by the near field probe which has an asymmetrical feed. Thus, the patterns in E-plane are believed to be dependent upon the probe effect. In order to correct the potential probing problems, a dipole-type probe is recommended. Overall, the aperture field indicates that the feed section with the prototype antenna provides the spherical wavefront over the whole frequency band.
Figure 6.14: Computed near field probe distributions of prototype tapered chamber feed section
Figure 6.15: Process of the construction for prototype tapered chamber feed antenna
Figure 6.16: Inner field distributions of the tapered chamber feed section $x - z$ plane - computed result
Figure 6.17: Inner field distributions of the tapered chamber feed section $x - z$ plane - computed result
Figure 6.18: Inner field distributions of the tapered chamber feed section $x - y$ plane at $z = 5.15$ ft. - computed result
Figure 6.19: Inner field distribuitons of the tapered chamber feed section $x - y$ plane at $z = 5.15$ ft. - computed result
Figure 6.20: Measured near field patterns at the aperture of the feed section
6.3.4 Specification of tapered chamber feed section and feed antenna

Feed section geometry

![Diagram of feed section geometry]

Figure 6.21: Top and side view of the feed section
Microwave absorber

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Table 6.1: Measured permittivity and conductivity of three absorber samples

Improved quad-ridged antenna

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Table 6.2: Extracted points of prototype quad-ridged antenna arms (see Figure 6.13)

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Table 6.2: Extracted points of prototype quad-ridged antenna arms (see Figure 6.13)
6.3.5 Proposed improved UWB dual-polarization wave - launching Design

The modified feed antenna and absorber arrangement provide the desired magnitude and phase variation except for the fixed radiation center. Investigation of the results obtained from numerical results and measurements has lead to better designs of the tapered chamber feed section. Instead of using the conical feed section, a cylindrical feed section is believed to be a good candidate for the tapered chamber feed section. Figure 6.22 - (a) illustrates the new feed section geometry including the configuration of the wall absorber. Major differences compared to the conical feed section are the apex region and the rectangular arrangement of the absorber. In order to avoid undesired reflections from the outer wall in the conical feed section near the apex, a cylindrical waveguide configuration is applied. It is noted that the cut-off frequency due to the diameter of the cylinder is lower than the operating frequency. For feed designs, a straight 30 degree horn angle blade is employed without the throat section. The curved section of the feed is merged into the absorber. The geometry of the feed antenna is shown in Figure 6.22 - (b).
Figure 6.22: Proposed feed section of the tapered chamber/UWB feed antenna
Figure 6.23: Near field probe distributions of proposed tapered chamber feed section
Figure 6.24: Spectrum comparison of transmitted co-polarized field at $z = 1.545$ m

Preliminary numerical results are obtained from the computation and shown in Figure 6.23 from (a) to (d). They are compared to the reference GO field distributions. No significant interference from the conductive blade arms are observed in phase and magnitude for both the E- and H-planes. From the phase plots, the radiation center is stays at the same location over the entire frequency band of interest. Finally, transmitted co-polarized field magnitudes are compared Figure 6.24 for the original blade feed, the prototype feed, and the proposed feed. Remarkable improvement is demonstrated for the new absorber arrangement with the prototype feed. However, in proposed case, although this configuration provides the best power delivery at low frequency, a resonance is found near the $1$ GHz region. It is believed to be from the reflection between the feed and the outer wall of
metallic cover. Thus, further improvement could be done with a new absorber arrangement that minimize the reflection at the junction between the antenna arms wall-type absorber.

6.4 Conclusion of the Chapter

In this dissertation, a novel UWB, dual polarized wave launcher for the tapered chamber was developed and analyzed using the FDTD method. The quad-blade configuration moved the radiation center closer to the desired chamber apex and reduced undesired reflections from side walls. The bandwidth was also significantly improved through proper design of the blade antenna and absorber geometry. Numerical result indicated that the operating frequency can be lower than 200 MHz which is lower than that of commercially available tapered chamber. The study indicated that the low frequency limit is related to the reflections arising from the placement of blade/absorber interface (see Figure 7), and these reflections could be minimized by a proper geometry design. A prototype of this new feed section was constructed and measured. The results confirmed improved magnitude and phase purity of the launched fields. Our finding indicated that further improvement can be achieved without using the conical geometry. Instead, a cylindrical feed section geometry would be more desirable and will be investigated in the future.
CHAPTER 7

CONCLUSION

By the assistance of the FDTD method, the UWB antennas for three different application areas are studied, designed, optimized and constructed for their individual application areas. All of the UWB antenna designs include geometrical and environmental complexity in both of the numerical modeling and the prototype construction. Areas of study are GPR system, compact range and tapered chamber. Using the FDTD method, final result commonly provide UWB operation, dual polarization and broadband matching. Each of the detail specification requirements are obtained based on the numerical result and interpretation of physical characteristics involving the electromagnetic theory. This approach saves time and construction of UWB antennas in many different application areas.

First, fully realistic features of the HFB antenna in a three dimensional FDTD numerical model has been created in this dissertation. They involved the modeling of dielectric loading, thin PEC plate antenna, coaxial feeding cable, tapered resistive cards, and the half-space medium. The tapered resistive cards were attached to the end of conductive antenna arms to reduce the late time antenna ringing. A linear tapered resistivity profile ranging from 3 to 300 $\Omega/\square$ was found to be effective, and this resistive card suppressed the late time antenna ringing co-polarization field by up to 50 dB compared to the case without resistive cards. The antenna’s surge impedance and reflection coefficient were calculated
with the feed cables in the same manner as is actually done in the real measurement cali-
bration and found to be close to the theoretical prediction for the no-dielectric filling cases. The characteristic impedance of the HFB antenna was found to be approximately 100 Ω in its operational band and was well matched to the coaxial cable when the dielectric constant of the filler was 5. For the fully polarimetric antenna analysis, the level of depolarization in the off-axis plane was found to be less than -12 dB relative to the co-polarization components along the antenna axis. These results validate the existing fully polarimetric HFB design.

Second application area was near-field measurement. In order to overcome the undesired characteristic of widely used OEWG antenna, a novel broad band dual-polarized DRP was developed and optimized by the help of the FDTD method. Various radiation section profiles are modeled and simulated in order to understand the relationships between the radiation section and the effects of beamwidth. It is found that the double-cosine profile provide the widest beamwidth with extremely low RCS level that yields low AUT-probe interactions. This novel UWB DRP have met all the near-field probe antenna requirements - low AUT-probe interactions, close to isotropic patterns, dual linear polarizations, frequency independent patterns, and symmetric E- and H-plane far-field patterns. Prototype antenna was built for the test measurement. And, simulated and measured far-field patterns are compared. They are well agreed in operation frequencies. Thus, FDTD model and computed result support the performance optimization in terms of beamwidth study of this application area with saving time and cost.
As a third application area, compact range feed design was studied and constructed by the assistant of FDTD method to optimize the radiation performance. Prototype antenna have dual linear polarized characteristics, reasonable efficiency, frequency independent radiation patterns, almost symmetric E- and H-plane patterns which are all attractive properties for compact range feed antenna.

DHA consist of five major components to construct: antenna, resistive card, solid dielectric horn, feeding structure and absorber materials. These configurations and basic criteria are discussed in Chapter 5. In addition, above variables are considered in numerical analysis. In numerical simulations, three dimensional FDTD grid of DHA has been created including the coaxial feeding structure, resistive cards, absorber, solid dielectric and four antenna arms. To improve beam formation into the broadside, internal field study was done in 120 degree and 110 degree cases. We found that 110 degree in nature, provide more focused beam into the broadside direction. Furthermore, by changing the boundary condition from circular cone to pyramidal cone, more 3 dB beam width especially in E-plane was achieved. In order to understand the antenna effect, excitation with antenna and point source cases are performed and compared. Used antenna was 1/8 inch in length and 30 degree plate angle. 2 inch long R-card strips are attached at the truncation of the conductive plates. Inner field (co-polarization field) distribution indicates that antenna reduces side lobes compared to that of point source excitation case. However, it reduces the 3 dB beamwidth. In previous design, conical shape introduced the undesired diffractions from the edges of spherical apertures. Corrections were made from the FDTD simulations by finding the sources of diffractions and curving the side walls. This idea is come from slot-line bowtie horn (SBH) antenna curve and successfully implemented into the DHA.
design. Numerical result provides improved 3 dB beamwidth in both plane patterns. Prototype antennas were built employing the predetermined dimensions from the numerical simulations. Horn angle is 110 degree with 7 inch diameter spherical aperture. 2.5 inch resistive card was newly measured and attached to the four arm antenna. 7 inch diameter PVC pipe was used to fix the DHA and feeding cable. Absorber (CL12) was mounted on the top of the solid dielectric horn and closed out by the cap. Antenna was measured in OSU anechoic chamber and showed that approximately 110 degree of 10 dB beamwidth in 2 - 18 GHz. Cross-polarization is also found to be 25 - 30 dB lower than the co-polarization level.

Final application area is the tapered chamber antenna development. The tapered chamber feed antenna operating from less than 200 MHz to higher than 3 GHz was designed in this dissertation. Feed antenna has a dual linear polarization characteristic, a minimum diffractions, and broadband matching. This type of antenna is unique in terms of operation frequency band due to the geometrical characteristic of taking the advantage of modal characteristics of conical feed section. As discussed about the cut-off issue in design, the improved feed structure of the chamber, quad-ridges inside the conical section was installed by the following the criteria which is used in circular waveguide due to the similarity of the modal characteristics. And avoiding the inductive region, it is possible to realize the low frequency operation with the chamber facility. These operation frequency restrictions which have been suffered in long times, are solved during the study.

7.1 Future study of UWB Antennas Design

Using the FDTD method and through understanding of the result, many different type of antennas can be developed and optimized before the prototype constructions. It is very
effective and provide the wide range of freedom to change, compare, and design in different application areas. For the future UWB antenna development, a fully automated process of numerical support will provide more effective and inexpensive procedure to deliver the final form of designs.

In UWB DRP design, focus of the numerical study is limited to the radiation section development. The other performance control variables such as feed antenna, waveguide section, and microwave absorber-are also very important to design further optimized DRP antenna. Thus, for the future direction of the DRP, automated numerical tools for the determination of feed antenna design which provide the desired operation frequencies and optimized radiation section profiles under the various contrains are highly desirable so that the final design of DRPs for the different environmental situations can be easily done.

There are two major future studies for DHA development. First one is antenna efficiency. Since DHA include 1/8 inch size four conductive plates which is electrically small below 10 GHz and excite the field as similar as point source excitation. One way to improve the efficiency is employing longer and wider plates. However, larger antenna creates significant amount of diffractions and distort the field formation unless special treatment at the any of discontinuous edges is applied. Alternative can be removing the absorbers which is assigned for eliminating possible lateral waves. More tentative study in this area is required for further improvement. Second future work will be a reducing the DHA size. As we studied inner field distributions, spherical wave front is settled down within the short distance from the feed point. Thus, small size DHA design can be possible with a new launcher.

For the tapered chamber antenna, further improvement can be achieved by the investigation of microwave absorber material characteristics. In the prototype design, absorber
samples are obtained from the Microwave Cummings Co. and installed inside the delivered conical feed section. It is found that the conductivity of the absorber is too high resulting the significant attenuation of the electromagnetic wave that yields poor power delivery into the test zone. Thus, study about the different absorber material within the chamber is highly suggested. In the last section of the Chapter 6, a proposed new chamber geometry is a cylindrical feed section. The research for the interaction between quad ridge antenna and absorber can provide the more improvement in the purity of the launched field into the test zone.


[83] www.microwavecuming.com


