Compact Planar Ultra-wideband Antennas for
Ground Penetrating Radar

A Dissertation Presented
by
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to
The Department of Electrical and Computer Engineering
in partial fulfillment of the requirements
for the degree of
Doctor of Philosophy
in the field of
Electrical Engineering
Northeastern University
Boston, Massachusetts
November, 2013
Acknowledgements

The study and research life at Northeastern University has been a much valuable and meaningful experience to me. When recalling the period of the PhD program, I genuinely feel blessed and grateful. Without the help from so many teachers and friends, it would be impossible for me to complete my study and research work here at Northeastern University.

First I would like to thank my advisor, Prof. Nian-Xiang Sun, and my co-advisor Prof. Ming-Liang Wang for their constant help and support during the time that I spent on my research. Their patient guidance and intriguing advice helped me in exploring and learning, which taught me to be a better researcher.

Also, I want to express my great gratitude to Prof. Ralf Birken of the Civil and Environmental Engineering Department. He spent so much time working together with us, doing GPR testing and helping me to improve my research paper writing. I also appreciate Dr. Dan Busuioc’s helpful discussion and consulting work on antenna design and manufacturing.

I would also like to thank Prof. Marvin for his role on my dissertation committee. His suggestions and comments were really helpful.

The research work could only be done with the friendly help from the other group members. I would like to thank Jing Wu, Yifeng Lu, Reid Vilbig, Hao Liu, Ziyao Zhou, Xi Yang, Yuan Gao, Tianxiang Nan, Ming Liu, Jing Lou, and Shawn Beguhn for the
invaluable help and collaboration in the laboratory. Many thanks should also be given to Ms. Veena Teli and Dr. Yinhong Cao for their constant help on my research work.
Abstract

Compact and low-cost Ground Penetrating Radar (GPR) systems are attractive in detecting pavement layers and subsurface defects such as rebar corrosion at driving speed. Ultra-wideband (UWB) antennas are key elements for the high-speed operation of the air-coupled GPR systems. This thesis studies and develops a number of planar antennas that have been manufactured with low-cost printed circuit board (PCB) technology. It also presents a brief methodology for the design process in order to frame the context and boundary conditions of the antenna problem, and to satisfy the regulatory specifications such as FCC compliance.

A variety of compact low-profile UWB antennas are designed, fabricated, characterized and tested with GPR systems, including rounded Bowtie antenna, Bowtie slot antenna and Vivaldi antennas. All antennas are intended to operate within the 1.1 – 4 GHz frequency band and benefit from compact size while providing high gain to allow for the detection of pavement layers and rebar in bridge decks to a depth of up to 2 feet. In-field measurements of the antennas, together with the GPR system, are presented for static testing scenarios such as buried rebar in a sand box and concrete slab. The antenna testing over the sand box and concrete slab demonstrates the great potential of utilizing the proposed antennas in air-coupled GPR systems, especially the compact rounded Bowtie and Bowtie slot antennas.

Antenna array is also developed to achieve enhanced penetrating capability of the GPR system while maintaining high resolution. When two Pacman antennas in the array are positioned in parallel, the gain of the array can be augmented up to 4 dB when compared
to the original single Pacman application. Anti-parallel Pacman arrangement in the antenna array has also been proposed and tested, which could eliminate the ground reflection and direct coupling effectively due to the 180° phase difference between the two Pacmans in one antenna array.

The antenna dispersion problem is also summarized and clarified in detail with theoretical analysis, simulation models and experimental characterization to aid the UWB antenna development for impulsive GPR system.
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Chapter 1 Motivation and Overview

1.1 Motivation
Ground Penetrating Radar (GPR) is a non-destructive technology that uses electromagnetic radiation of the RF/microwave band to detect or image the subsurface structures. It has extensive applications in a variety of fields, such as mine detection in geophysical research, and buried objects exploration in archaeology [1]. Due to its strong ability to precisely locate underground utilities (such as rebar and pipes) and to assess the quality and deterioration of asphalt or concrete pavement and bridge decks, GPR is an attractive non-destructive testing and evaluation method commonly used for health monitoring of roadways and bridge decks. [2]

Generally, GPR systems rely on ultra-wideband (UWB) pulse radar technology, in which broadband radio frequency electromagnetic waves are transmitted and received by antennas. The properties of the subsurface structures can be characterized according to the variation of the reflected signals. Historically, GPR systems are ground-coupled with the antenna operated in contact with the ground as it reduces the loss at the air-ground-interface and allows for a greater depth of investigation. For the mandate of an in-traffic operated GPR, such systems are impractical and an air-coupled system is a requirement [3].

One of the difficulties for such a GPR system is to develop a suitable Ultra-wideband (UWB) antenna of small electrical size and low-profile, with high gain and unidirectional stable radiation pattern. The compact and rugged physical characteristics are especially important for mounting the system in confined space, like the underside of a vehicle.
The VOTERS (Versatile Onboard Traffic-Embedded Roaming Sensors) project (www.neu.edu/voters) at Northeastern University in Boston provides a framework and prototype system to complement periodical localized inspections of roadways and bridge decks with continuous network-wide health monitoring [4]. Utilizing traffic-embedded Vehicles of Opportunity (VOOs) roaming through daily traffic eliminates hazardous, congestion-prone work zones that are typically set up to gather these critical inspection data sets. To realize this vision, VOTERS has developed four new prototype sensing systems. These systems collect data containing surface and subsurface (maximum of 1 m deep) condition information of roadways and bridge decks to locate and map defects at traffic speed and domain appropriate spacing. One of the sensing technologies developed is an improved air-coupled Ground penetrating radar (GPR) array system that is geared towards mapping subsurface defects such as corroded rebar, trapped moisture, voids, and the pavement layers (thicknesses and electromagnetic properties) [5,6]. This development was undertaken as current commercial systems didn’t fit the requirements of such a system. These requirements are: low-cost, multi-channel, low-profile (ideally for under vehicle deployment), low power, and capable of collecting GPR traces at cm intervals within traffic.

The main concern of this research is to design and develop compact planar Ultra-wideband antennas for the air-coupled Ground Penetrating Radar system.

1.2 Dissertation Overview
Traditionally the UWB antennas for ground penetrating radar system are bulky and contact the ground surface in operation. The work presented in this thesis focuses on the development of compact planar UWB antennas for GPR system, especially in air-coupled GPR application.
Chapter 1 presents the motivation and overview of the VOTERS research project at Northeastern University, which has been developing versatile sensors including RF/microwave sensor – the air-coupled ground penetrating radar system.

In Chapter 2, the fundamentals of Ultra-wideband Antennas are introduced. Basic concepts to characterize UWB antennas will be presented, including impedance bandwidth, radiation pattern, gain, efficiency, polarization and antenna dispersion. These parameters will be discussed separately in this chapter. Then the UWB antennas will be classified, and requirements of UWB antenna for GPR application will also be discussed.

Chapter 3 discusses a low-profile, high-gain reflector antenna, which is composed of a rounded bowtie antenna and a metal cavity as a planar reflector. By performing careful design based on parametric analysis and simulation, this antenna can operate at a broad bandwidth of 0.8 – 3.5 GHz in theory. Such good performance makes it a promising design for the under-vehicle GPR, but it will be found in this chapter that the ringing issue may hinder its application in subsurface radar system.

Chapter 4 can be divided into two major parts: the first part provides a detailed overview and summary on the concept of antenna dispersion, which plays an essential role in the performance of impulsive ground penetrating radar system. The antenna dispersion topic will be explained from theoretical analysis, time domain and frequency domain characterization experiments, and simulation model. The second part of this chapter discusses some improvement on the Cavity-backed Rounded Bowtie Antenna Design presented in Chapter 3. The modified antenna will be characterized as well in comparison with the previous measurement in Chapter 3, and the difference between the two versions of rounded bowtie antennas will be discussed.
Chapter 5 presents a new type of antenna design – a highly compact bowtie slot antenna for ground penetrating radar application. The proposed bowtie slot antenna is eyes-shaped and fed by 50 Ohm coplanar waveguide (CPW) structure, which is quite suitable for system integration. The operation frequency range of the bowtie slot antenna is 1.1 – 3.5 GHz. Careful parameter analysis during the numerical simulation leads to a unidirectional, high-gain antenna with stable radiation patterns. The antenna dispersion of the bowtie slot antenna is also simulated and compared with experimental results, so that the time domain response of such antenna can be monitored during the design and verification process.

Chapter 6 discusses about the antenna testing with ground penetrating radar systems. This chapter serves as a summary of different types of antennas developed for the VOTERS GPR project, including rounded bowtie antenna, bowtie slot antenna, and Vivaldi antennas (planar horn antenna). New models of Vivaldi antenna designed at Northeastern University will also be presented with GPR testing results. Furthermore, how the GPR system works with the UWB antennas will be covered in detail, which is supposed to provide useful reference and guidance for antenna design for practical impulsive GPR system. Then GPR experimental results of several types of UWB antennas, including miniaturized rounded bowtie antenna, bowtie slot antenna and one Vivaldi antenna will be presented on both sand box and concrete slab testing. The performance of each antenna collaborating with the compact GPR system at VOTERS project will be demonstrated clearly in this chapter.

Chapter 7 presents the possibility and practical development of antenna arrays for impulsive ground penetrating radar application. Conventionally only single UWB
antenna is used as transmitter or receiver antenna for impulsive GPR. In this chapter, we will explore the feasibility of antenna arrays for impulsive radar application. The analysis and experimental work successfully demonstrate the plausibility of antenna array used as transmitter and receiver antenna. An array made of two Pacman antennas is constructed for GPR measurement on both sand box and concrete slab testing. The antenna array can provide stronger penetrating capability, compared to the traditional single-antenna GPR. With some special arrangement of individual antennas in the array, the GPR system could demonstrate some interesting and useful features, such as automatic ground reflection cancellation and direction coupling reduction. The antenna dispersion problem is also discussed in order to maintain good time domain performance for antenna array application. This work provides great potential of antenna array application in impulsive subsurface radar system.

Chapter 8 presents the summary of the research work on UWB antennas design for GPR system.

1.3 References


Chapter 2 Fundamentals of Ultra-wideband Antennas and Ground Penetrating Radar

2.1 Definition of Ultra-wideband Antennas
An antenna is used as a transducer between electromagnetic waves in free space and signals on a transmission line, or vice-versa. (John antennas) A transmitting antenna converts signals on the transmission line into electromagnetic waves, and a receiving antenna collects the electromagnetic waves in the space, and converts them back into signals on the transmission line. The radiation, propagation and reception of electromagnetic waves obey the Maxwell’s equations. [1]

\[
\begin{align*}
\nabla \times E &= -\frac{\partial B}{\partial t} - M \\
\nabla \times H &= -\frac{\partial D}{\partial t} + J \\
\n\nabla \cdot D &= \rho \\
\n\n\nabla \cdot B &= 0
\end{align*}
\] (2-1)

where \( E \) is the electric field, \( D \) is the electric displacement field, \( H \) is the magnetic field, \( B \) is the magnetic flux density, \( M \) is the magnetic current density, \( J \) is the electric current density, and \( \rho \) is the electric charge density.

For linear materials, the constitutive relations between \( E, D, H \) and \( B \) are

\[
\begin{align*}
D &= \varepsilon_r \varepsilon_0 E \\
B &= \mu_r \mu_0 H
\end{align*}
\] (2-2)

where \( \varepsilon_0 \) is the permittivity of the free-space, \( \mu_0 \) is the permeability of free-space. \( \varepsilon_r \) is the relative permittivity and \( \mu_r \) the relative permeability of the material. For free-space, \( \varepsilon_r = 1, \mu_r = 1 \).

Antennas are ubiquitous. They exist in any equipment that uses wireless technology, such as cell phones, Bluetooth, satellite communications, as well as other devices such
as Radio-frequency identification (RFID) tags and garage door openers. Most antennas used in the everyday communication systems are narrowband. For instance, the 2G Global System for Mobile Communications (GSM) networks in US are allocated 850MHz and 1900MHz frequency bands. All satellites for the Global Positioning System (GPS) broadcast at the same two frequencies, 1.57542 GHz (L1 signal) and 1.2276 GHz (L2 signal) when we drive with a GPS navigator. The antenna bandwidth for such systems is usually about 1% of the operating frequency.

It is not surprising that ultra-wideband (UWB) antennas distinguish themselves from ordinary narrowband antennas by their large bandwidth. According to the U.S. Federal Communications Commission (FCC), Ultra-wideband refers to radio technology with a bandwidth exceeding 500 MHz or 20% of the arithmetic center frequency.[2] That is, Ultra-wideband refers to bandwidth with

\[ bw = 2 \frac{f_u - f_l}{f_u + f_l} \geq 0.2 \]

\[ or \ bw = f_u - f_l \geq 500 \text{ MHz} \quad , \quad (2-3) \]

where \( f_u \) is the upper end of the antenna’s operational band, and \( f_l \) refers to the lower end of the antenna’s operational band. The \( f_u \) and \( f_l \) are defined by the points where the radiated power is down 10 dB from the peak level in the figure of radiated power spectral density. Since the spectral power density is not that intuitive in the antenna experiment, the upper and lower ends of the antenna’s operational band will be defined by points where the S11 (reflection coefficient) is less than -10 dB in the impedance bandwidth graph.
2.2 Essential Concepts for UWB antennas
Before the discussion of UWB antenna application in Ground penetrating radar system, some fundamental concepts to characterize UWB antennas will be presented first. It is essential to understand these concepts, since the UWB antenna performance is parameterized by such quantities as impedance bandwidth, radiation pattern, gain, efficiency, polarization and antenna dispersion. These parameters will be discussed in this section separately.

To clarify these concepts, the small thin-wire dipole antenna can be taken as a good example. This kind of antenna is also referred to as infinitesimal dipole, which is assumed to be very small ($l \ll \lambda$) and very thin ($a \ll \lambda$). Here $l$ is the length of the Infinitesimal dipole, and $a$ is the diameter of the dipole wire, as shown in Figure 2-1. [3]

![Figure 2-1. Infinitesimal electrical dipole](image)

The spatial variation of the current is assumed to be constant, and the time-dependent current distribution $I$ can be written as:
When \( T(t) = e^{i\omega t} \), it is assumed that the current distribution is harmonic time dependent.

The region of charge in the dipole is considered to be confined to the end points, and the relation between the charge \( q \) and the current distribution \( I \) is \( I = \frac{dq}{dt} \). The electrical dipole moment [4]

\[
p(t) = q(t) \cdot l \hat{2} \quad (2-5)
\]

The Infinitesimal dipole is chosen here for three reasons: 1) Due to its simple structure, the Infinitesimal dipole can keep the mathematical details to a minimum, and meanwhile can still illustrate the fundamental concepts of antenna radiation. 2) Some other types of broadband antennas can be derived from this basic model, such as biconical antenna, bowtie antenna and ellipse antenna. 3) It will also serve as a useful model in the future chapters when we discuss the reflectors and antenna dispersion.

In general, the vector potential excited by current distribution \( J(r', t) \) is

\[
A(r, t) = \frac{\mu_0}{4\pi} \int \frac{J(r', t - \frac{|r - r'|}{c})}{|r - r'|} \, dr' , \quad (2-6)
\]

where \( r \) represents the observation point coordinates, \( r' \) represents the coordinates of the source, and \( c \) is the light speed.

Let \( R = r - r' \), \( \hat{R} = R/R \), \( R = |R| \), and define a function \( T(t) \) at retarded time

\[
t_r = t - R/c
\]

\[
[T(t)] = T(t - R/c) \quad (2-7)
\]

Using (2-5), the magnetic flux density

\[
I(z, t) = \dot{l} = 2l_0 T(t) \quad (2-4)
\]
\[ B = \nabla \times A = \frac{\mu_0}{4\pi} \left( \nabla \frac{1}{R} \times \int [J(r', t)]dr' + \frac{1}{R} \int \nabla \times [J(r', t)]dr' \right) \]
\[ = \frac{\mu_0}{4\pi} \int \left( \frac{[J] \times \ddot{R}}{R^2} + \frac{1}{Rc} \frac{\partial J}{\partial t} \times \ddot{R} \right)dr' \]  (2-8)

The equation (2-7) is the Jefimenko form of the Biot-Savart Law. It is the time-dependent generalization of the Biot-Savart law to electrodynamics, which were originally true only for steady currents.

For the infinitesimal electric dipole, the vector potential can be approximated as

\[ A(r, t) = \frac{\mu_0}{4\pi r} [\dot{p}] \]  (2-9)

In the following, \([\dot{p}] = p(t - r/c)\)  (2-10)

Applying the Taylor series about the origin, the magnetic flux density becomes

\[ B(r, t) = \nabla \times A = -\frac{\mu_0}{4\pi r} \left( \frac{1}{r^2} \hat{p} \times [\ddot{p}] + \frac{1}{cr} \hat{p} \times [\ddot{p}] \right) \]  (2-11)

The scalar potential for the two point charges is

\[ \Phi(r, t) = \frac{1}{4\pi \varepsilon_0} \left( \frac{q(t-s_1/c)}{s_1} - \frac{q(t-s_2/c)}{s_2} \right) \]  (2-12)

With some approximation, the \(\Phi(r, t)\) can be written as

\[ \Phi(r, t) = \frac{1}{4\pi \varepsilon_0} \frac{\hat{r} \cdot p}{r} \left( t - \frac{r}{c} \right) + \frac{\hat{r}}{c} \cdot \dot{p} \left( t - \frac{r}{c} \right) \]  (2-13)

Then the radiated electrical field is [5]

\[ E = -\nabla \Phi - \frac{\partial A}{\partial t} = -\frac{1}{4\pi \varepsilon_0} \left( (1 - 3\hat{r} \cdot \ddot{r}) \left( \frac{1}{r^3} [p] + \frac{1}{cr^2} [\dot{p}] \right) - \frac{1}{c^2 r \hat{r}} \times (\hat{r} \times [\dddot{p}]) \right) \]  (2-14)

Equations (2-11) and (2-14) represent the general time dependence of radiated field for an infinitesimal electric dipole, which will help the discussion of the time-domain response of UWB antennas in Chapter 4.

For harmonic time dependence, since \(r \gg L\) and \(\lambda \gg L\) for the Infinitesimal dipole, with some approximation during the calculation, the electric field \(E\) can be got as [3]:
The magnetic field $\mathbf{H}$ is:

$$
\begin{align*}
E_r &= \eta \frac{\ell d \cos \theta}{{2\pi}r^2} \left[ 1 + \frac{1}{kj} \right] e^{-jkr} \\
E_\theta &= j\eta \frac{k l \ell \sin \theta}{{4\pi}r} \left[ 1 + \frac{1}{kj} \right] e^{-jkr} \\
E_\phi &= 0
\end{align*}
$$

(2-15)

In the far-field region where $kr \gg 1$, we have

$$
\begin{align*}
H_\phi &= j\eta \frac{k l \ell \sin \theta}{{4\pi}r} e^{-jkr} \\
H_r &= H_\theta = 0
\end{align*}
$$

(2-16)

Then we will discuss about each individual concepts for UWB antenna characterization.

### 2.2.1 Radiation Pattern, Directivity and Gain

The radiation pattern is one of the factors indicating which antenna should be used in the very application. For instance, the cell phone should employ an antenna with nearly omni-directional radiation, so that the mobile can always receive optimal signals in any position or direction. When applied in Ground Penetrating Radar system, a unidirectional radiation pattern is desired, since the radiation leakage would give rise to noise from environment and some other unwanted couplings between antennas.

According to the IEEE Standard Definitions of Terms for Antennas, an antenna radiation pattern is defined as ‘a mathematical function or a graphical representation of the radiation properties of the antenna as a function of space coordinates. In most cases, the
radiation pattern is determined in the far-field region and is represented as a function of
the directional coordinates. Radiation properties include power flux density, radiation
intensity, field strength phase or polarization. In future discussion, the radiation
intensity \( U \) and the parameters Gain or Directivity derived from radiation intensity are
usually employed to describe the antenna radiation pattern.

Using equation (2-17), the average radiation power density of the Infinitesimal dipole
antenna can be written as [3]

\[
W_{av} = \frac{1}{2} Re(E \times H^*) = a_r \frac{1}{2\eta} |E_\theta|^2 \quad (2-18)
\]

The radiation intensity is defined by

\[
U = r^2 W_{av} = \frac{r^2}{2\eta} |E_\theta|^2 = \frac{\eta}{2} \left( \frac{k|l_0|^2}{4\pi} \right)^2 \sin^2 \theta, \quad (2-19)
\]

which represents the power radiated from an antenna per unit solid angle.

The total power radiated by the Infinitesimal dipole is

\[
P_{rad} = \int_0^{2\pi} \int_0^\pi U(\theta, \phi) \sin \theta d\theta d\phi = \frac{\eta \pi}{3} \left( \frac{l_0}{\lambda} \right)^2 = \frac{1}{2} l_0^2 R_r, \quad (2-20)
\]

where \( R_r = \frac{2\eta}{3} \left( \frac{l_0}{\lambda} \right)^2 \) is defined as the radiation resistance of the antenna.

The average radiation intensity of an antenna is \( P_{rad}/4\pi \), which is the total radiated
power divided by the entire solid angle of \( 4\pi \).

Then the directivity of the Infinitesimal dipole antenna can be calculated as

\[
D = \frac{U(\theta, \phi)}{P_{rad}/4\pi} = \frac{3}{2} \sin^2 \theta, \quad (2-21)
\]
which represents ‘the ratio of the radiation intensity in a given direction from the antenna to the radiation intensity averaged over all directions’.

From the above equation (2-21), we can plot the three-dimensional radiation pattern. If the direction is not specified, it implies the direction of maximum directivity where \( \theta = \pi/2 \), and \( D_0 = D_{\text{max}} = \frac{3}{2} = 1.76 \text{ dBi} \). \hspace{1cm} (2-22)

Here dBi means that the directivity is referred to an isotropic radiator.

Figure 2-2 illustrates a 3D radiation pattern of electric dipole simulated in HFSS. It is measured on a spherical coordinate system indicating relative strength of radiation power in the far field. The length of the electric dipole is \( l = 0.2\lambda \), and the diameter of the cylinder is \( a = 0.01\lambda \). From the simulation we find that the directivity is 2.135 dBi, which is slightly larger than the theoretical value shown above, since the antenna in the simulation is not an ideal infinitesimal dipole. Instead, it is a small antenna.
Figure 2-2. Electric dipole model and its radiation pattern (a) an electric dipole model, the length $l = 0.2\lambda$, the diameter of the cylinder $a = 0.01\lambda$. (b) Simulated 3D radiation pattern of Infinitesimal Dipole in HFSS

Conventionally, for a linearly polarized antenna, ‘the plane containing the electric field vector and the direction of maximum radiation’ is defined as the E-plane, and ‘the plane containing the magnetic field vector and the direction of maximum radiation’ is defined as the H-plane. For the above example, the x-z plane (elevation plane) is the E-plane, and the x-y plane (azimuthal plane) is the H-plane. In reality, example of Figure 2-2
demonstrates infinite number of E-planes (elevation planes; \( \phi = constant \)) and single H-plane (azimuthal plane; \( \theta = \pi/2 \)).

Also, the radiation patterns can be illustrated in two-dimensional way, the E-plane and H-plane. Figure 2-3 shows the two dimensional radiation pattern plots for Infinitesimal dipole. In Figure 2-3 (a) the radiation pattern of E-plane is demonstrated, where \( \theta \) varies with \( \phi = 0 \). Figure 2-3 (b) shows the radiation pattern of H-plane, where \( \phi \) varies with \( \theta = \pi/2 \). It can be found that the maximum directivity is achieved at \( \theta = \pi/2 \), which is consistent with the theoretical analysis above.
Figure 2-3. Two dimensional radiation pattern plot for Infinitesimal dipole (a) E-plane, varying $\theta$ with $\phi = 0$, (b) H-plane, varying $\phi$ with $\theta = \pi/2$. 

<table>
<thead>
<tr>
<th>Name</th>
<th>Theta</th>
<th>Angle</th>
<th>Mag</th>
</tr>
</thead>
<tbody>
<tr>
<td>m1</td>
<td>90.0000</td>
<td>90.0000</td>
<td>2.1355</td>
</tr>
<tr>
<td>m2</td>
<td>51.0000</td>
<td>51.0000</td>
<td>-0.9443</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Name</th>
<th>Phi</th>
<th>Angle</th>
<th>Mag</th>
</tr>
</thead>
<tbody>
<tr>
<td>m1</td>
<td>360.0000</td>
<td>-6.0000</td>
<td>2.1355</td>
</tr>
</tbody>
</table>
The half power beam width (HPBW) is also shown in Figure 2-3. The HPBW is referred to the angular distance from the center of the main beam to the point at which the radiation power is reduced by 3 dB. This parameter is useful when describing quantitatively how directive an antenna is. For example, the radiation pattern of the Infinitesimal dipole in the E-plane exhibits an HPWB of around 78 degree, while in H-plane there is no radiated power level variance.

The gain of an antenna is defined as [3]

\[ G = 4\pi \frac{U(\theta, \phi)}{P_{in}}, \quad (2-23) \]

where \( P_{in} \) is the total (accepted) input power by the antenna.

The total input power \( P_{in} \) and the total radiated power \( P_{rad} \) is related by

\[ e_{cd} = \frac{P_{rad}}{P_{in}}, \quad (2-24) \]

where \( e_{cd} \) is the antenna radiation efficiency we will mention again in the following section.

Consequently there is a relation between gain and directivity

\[ e_{cd} = \frac{G(\theta, \phi)}{D(\theta, \phi)} \quad (2-25) \]

Therefore gain takes into account the efficiency and the directional capability of an antenna, while directivity considers merely the directional properties and is controlled by the radiation pattern only.
In the GPR application, the radiation gain has a profound influence in the radar’s penetrating capability. This property can also be verified through the concept of Antenna Factor.

The Antenna Factor (AF) can directly relate a received voltage to incident field strength. It is defined as

$$AF = \frac{|E_{inc}|}{V_{rec}}, \quad (2-26)$$

where $|E_{inc}|$ is the incident electric field strength and $V_{rec}$ is the received voltage by antenna. Therefore, the Antenna Factor can be described as the required electric field strength that produces 1 Volt at the terminals of an antenna.

When the load impedance is 50 Ohms, the Antenna Factor is related to gain $G$ and the wavelength $\lambda$ as:

$$AF = \frac{9.73}{\lambda \sqrt{G}} \quad (2-27)$$

This formula is valid only when the receive antenna is aligned in the direction of maximum gain in the far-field of the transmit antenna. From the above formula (2-26) and (2-27), it can be found that the higher the antenna gain, the larger voltage signal the antenna can receive for the same incident electric field.

### 2.2.2 Impedance Bandwidth

Reflections will occur when impedance discontinuity happens on the transmission line. A good impedance match can maximize the power transfer and efficiency of an antenna.
Two parameters are usually used to measure the quality of antenna impedance matching: Voltage standing wave ratio (VSWR) and a scattering parameter (S11). They are not independent, and can be related by the parameter $\Gamma$, the voltage reflection coefficient at the input terminals (a, b) as shown in Figure 2-. The definition of VSWR and S11 are:

\[
\text{VSWR} = \frac{1+|\Gamma|}{1-|\Gamma|} \quad (2-28)
\]

\[
S11(\text{dB}) = 20\log|\Gamma| \quad (2-29)
\]

**Figure 2-4. Reflection and impedance match of an antenna.** The antenna shown above is the Pacman antenna designed for the VOTERS GPR.

For the configuration in Figure 2-4, the reflection coefficient can be written as

\[
\Gamma = \frac{Z_A - Z_0}{Z_A + Z_0} \quad (2-30)
\]

Impedance bandwidth denotes the bandwidth for which the antenna is sufficiently matched to its input transmission line such that no more than 10% of the incident signal is lost due to reflections. That is, $S11 < -10$ dB. When the antenna input impedance $Z_A$ equals to the transmission line impedance $Z_0$ within the special operating frequency range, then the UWB antenna is well matched.
In Figure 2-5, the S11 of the dipole is plotted with length $l = 0.2\lambda$, the diameter of the cylinder $a = 0.005\lambda$. Here $\lambda = 300$ mm. The central (resonance) frequency of the dipole is 2.20 GHz.

**Figure 2-5. The S11 of a dipole with length $l = 0.2\lambda$, the diameter of the cylinder $a = 0.005\lambda$.**

The bandwidth of this dipole antenna is 320 MHz = $f_u - f_l$, where $f_l = 2.032$ GHz, and $f_u = 2.350$ GHz. Or the bandwidth can be represented as $bw = (2.350 - 2.032)/(2.350 + 2.032)*2 = 14\%$. According to the definition of ultra-wideband, this dipole is a narrow band antenna.

### 2.2.3 Efficiency

The total antenna efficiency $e_0$ describes the losses at the input terminals and within the antenna structure. In general, it consists of two main parts: 1) efficiency $e_r$ related to reflection (mismatch) between the transmission line and the antenna; 2) efficiency $e_{cd}$
related to conduction and dielectric loss in the antenna. The total efficiency can be written as

\[ e_0 = e_r e_{cd} = e_{cd} (1 - |\Gamma|^2) \]  \hspace{1cm} (2-31)

where \( e_{cd} \) is also referred to as antenna radiation efficiency, which is used to relate the gain and directivity of an antenna. [3]

![Figure 2-6. The equivalent circuit of a transmitting antenna](image)

Figure 2-6 shows the equivalent circuit of a transmitting antenna, where the antenna is attached to a generator with internal impedance \( Z_g \). For an antenna of (input) impedance

\[ Z_A = R_A + jX_A = R_r + R_L + jX_A \]  \hspace{1cm} (2-32)

where \( R_A \) is antenna resistance, \( X_A \) is antenna reactance, \( R_r \) is radiation resistance of the antenna, and \( R_L \) is the conduction and dielectric loss resistance of the antenna, the radiation efficiency can be written as

\[ e_{cd} = \frac{R_r}{R_r + R_L} \]  \hspace{1cm} (2-33)
For an infinitesimal dipole of length $l = \lambda/50$, the radiation resistance $R_r = \frac{2\eta}{3} \left( \frac{l}{\lambda} \right)^2 = 80\pi^2 \left( \frac{l}{\lambda} \right)^2 = 0.316$ Ohms. Therefore, it will exhibit a very large mismatch when connected to a practical transmission line of 50 or 75 Ohms. Considering the comparable loss resistance of this kind of antenna, the radiation efficiency and the overall efficiency hence will be very low due to the impedance mismatching between the transmission line and the antenna.

### 2.2.4 Polarization

The far-field waves transmitted (or radiated) by the antenna are transverse to the direction of propagation, that is, the transverse electromagnetic (TEM) waves. Then the polarization of the antenna can be defined as the orientation of the electric field of the TEM wave.

The polarization of an antenna can be classified as linear, circular or elliptical. If the electric field vector at a point in space as a function of time always lies along a line, the polarization of the electric field is defined as linear. But usually the path the electric field traces out is an ellipse, so the polarization is said to be elliptical. When the magnitudes of the two components of the electric field $\mathbf{E}$ are the same and the phase difference between them is odd multiples of $\pi/2$, the elliptical polarization will become circular.

For most UWB antennas in the GPR application, their polarization is linear. The electric field of a horizontally polarized antenna is horizontally oriented, while the electric field of a vertically polarized antenna is vertically oriented. When the polarization axes of a
transmit antenna and receive antenna are placed out of alignment by an angle $\phi$, the signal power received by the receive antenna will be $\cos^2 \phi$. As a result, there might be a detection null point when the electric fields of two linearly polarized antennas lie orthogonally to each other, that is, when $\phi = \pi/2$. Similarly, linearly targets such as small pies and rebar buried in the concrete slab could also produce a sinusoidal variation in received signal when the linearly polarized antenna is aligned with a certain angle to the target direction.

An attractive option is to make the antenna radiate circularly, so that the target can always be detected even when its orientation is unknown. One of such antenna design is equi-angular spiral antenna, whose polarization is circular. Unfortunately, such antenna display big deficiency which hinders its application for impulsive GPR. The spiral antenna is seriously dispersive, which causes a stretch-out in the transmitted waveforms, and the radiated pulse takes the form of a 'chirp'. This is also named as ringing effect, since the waveform oscillation lasts some time before it disappears.

In reality, however, an antenna is never 100% polarized in a single mode (linear, circular, etc). Hence, the polarization is sometimes resolved into a pair of orthogonal polarizations, the co-polarization (or desired polarization component) and the cross-polarization (orthogonal to a specified polarization, usually the co-polarization).

### 2.2.5 Antenna Dispersion

Antenna dispersion is essential for the UWB antenna performance in impulsive Ground Penetrating Radar system, especially the resolution of GPR. The section 2.2.4 has
mentioned that equi-angular spiral antenna, one kind of frequency-independent UWB antenna, would demonstrate serious antenna dispersion, which could prevent its application in impulsive GPR application. This phenomenon sometimes is also referred to as ringing effect. This topic will be covered in detail in Chapter 4.

2.3 UWB Antenna Classification
Four different classes of UWB antennas are summarized according to their forms and functions as the major categories in UWB system application, especially for GPR.

2.3.1 Element antennas
Element antennas tend to possess small size and low profile, which makes them attractive for application in space-confined condition. Also, they are generally characterized by linear polarization and omni-directional radiation. Examples of element antennas include monopoles, dipoles, conical antennas and bow-tie antennas. Disadvantages of element antennas include low directivity, low gain, and relatively limited bandwidth.

To overcome the disadvantages and also to meet different requirements of application, various techniques have been employed in the antenna design. For example, by increasing the diameter of the dipole cylinder, the bandwidth can be increased. More generally, when the dipole goes fatter and fatter in the three dimensional space and toward a sphere, its bandwidth could be extended to a maximum limit. For planar antenna design, when the bowtie antenna becomes bulbous or when its corner is rounded, the bandwidth is also broadened, and the antenna dispersion could be decreased. A myriad of antenna configurations have evolved from the element antennas, including broadband dipole, ellipse antenna, rounded Bowtie antenna, bottom-fed planar rounded Bowtie antenna. One example of ellipse antenna is shown in the Figure 2-7.
Another technique to increase the limited bandwidth is resistive loading, by sacrificing the radiation efficiency of the antenna [6]. Typically, a resistively loaded antenna exhibits a gain at least 3 dB lower than the unloaded antenna, which makes such method unfavorable. This technique, however, could effectively reduce the notorious UWB antenna ringing. The mechanism behind it is that a travelling wave distribution of current can be produced by suitable resistive loading. As a result, less or no reflections are formed at the antenna ends, which produce less or no standing waves. This method actually is very popular when discussing about traveling-wave antennas. The long wire traveling wave antenna, the V antenna and the rhombic antennas are all terminated by a resistor load.

In order to enhance the antenna gain, antenna array can be assembled for the dipole or bowtie arrays. There are not many literatures discussing this topic, due to its inherent restrictions such as large antenna dispersion and difficulty in array assembly. In this
dissertation, several types of antenna array are discussed, including the theoretical analysis, numerical simulation and experimental verification. Promising results are demonstrated to design powerful UWB antennas for GPR penetrating ability.

To improve the radiation gain, an alternative way is to use reflectors, including planar reflector, metal-cavity reflector as well as parabolic reflector. Such method could also bring in unwanted ringing effect, which will mask the target signal and reduce the GPR resolution.

**2.3.2 Travelling wave antennas**

Unlike common dipole antennas, which exhibit current or voltage standing wave patterns formed by reflections from the open end of the wire, travelling wave antennas demonstrate the current and voltage distributions represented by one or more traveling waves, usually in the same direction.

A large variety of antennas with different configurations fall into this category, such as long wire antenna, the V antenna, the rhombic antennas, dielectric rod antenna, and helix antennas. Besides, aperture antennas like reflectors and horns can be treated as traveling wave antennas. Generally these antennas are composed of a pair of conductors in V structure, and are capable of supporting a forward travelling TEM wave. Vivaldi antenna, one kind of planar horn antenna, is a good example, as shown in Figure 2-8.
The traveling wave antennas can present several benefits for the GPR application, especially the TEM horn antenna. First, such antenna demonstrates pretty low antenna dispersion. In reality, the TEM horn antennas terminated with resistors are very good option for GPR system, if the antenna dimension is not critical factor. The resistor termination is used to reduce, if not completely eliminate, the current reflected from the end of the antenna. Such technique can suppress the ringing phenomenon effectively, and thus are quite attractive for impulsive radar application. Another reward is that the traveling wave antennas like the horn antennas tend to have high gain (typically 10 – 15 dB), which can demonstrate good capability of penetrating.

One disadvantage is the larger and bulkier size of horn antennas when compared with the element antennas, which could limit their application. Besides, the cost of the horn as well as the difficulty of fabrication is higher than that of element antenna.
2.3.3 Frequency independent antenna
Frequency independent antennas have geometries that are specified by angles. Examples include equiangular spiral antenna, log-periodic antenna, and conical spiral antennas. The geometry varies from a smaller-scale portion to a larger-scale portion. The smaller-scale portion contributes higher frequency radiation, while the lower frequency radiation relies on the larger-scale portion of the antenna. One example of frequency independent antenna is shown in the Figure 2-9.

![Figure 2-9. Archimedean-spiral antenna [7]](image)

As the effective origin of signals radiated from the antenna moves with frequency, the frequency independent antennas tend to be dispersive. Therefore, they are not suitable to impulsive GPR application. However, due to their broad bandwidth, they can be applied in the step-frequency radar system, which will not be covered in this dissertation.

2.3.4 Reflector antennas
Usually, reflector antennas utilize different metallic shapes, including planar reflector, metal-cavity reflector as well as parabolic reflector to concentrate energy in a particular direction. In this way, the gain of such antenna can be increased significantly compared to the element antenna, and is comparable to the gain of horn antennas. Meanwhile, the
reflector antennas are structurally simpler than the horn antennas, and thus are easier and cheaper to fabricate and adjust.

One disadvantage of the reflector antenna is that the reflections from the metallic reflector plane or cavity will produce a much stretched-out and distorted waveform. Such ringing effect will harm the performance of the overall GPR system. One way to reduce the ringing is to load some absorber materials, including carbon-adulterated foam absorber and magnetically loaded rubber absorber [8 - 10]. Detailed discussion of reflector antenna design for GPR will be covered in the next chapter, Chapter 3.

For convenience, the different types of UWB antennas have been summarized in Table 2-1, as shown below.
Table 2-1, UWB Antenna Summary for GPR Application

<table>
<thead>
<tr>
<th>Types of antennas</th>
<th>Examples</th>
<th>Notes</th>
</tr>
</thead>
<tbody>
<tr>
<td>Element antenna</td>
<td>Monopoles, dipole, conical and bow-tie, ellipse antenna, rounded Bowtie antenna, bottom-fed planar rounded Bowtie</td>
<td>linear polarization, low directivity, small bandwidth, gain 3 -4 dB, Small size</td>
</tr>
<tr>
<td>Traveling-wave antenna</td>
<td>long wire antenna, the V antenna, the rhombic antennas, dielectric rod antenna, and helix, Horn antennas, Vivaldi antenna</td>
<td>linear polarization, low dispersion with resistor termination, high gain (10 -15 dB), bulky</td>
</tr>
<tr>
<td>Frequency independent antenna</td>
<td>equiangular spiral antenna, log-periodic antenna, and conical spiral antennas</td>
<td>circular polarization (Spiral), high dispersion</td>
</tr>
<tr>
<td>Reflector antenna</td>
<td>planar reflector, metal-cavity reflector, parabolic reflector, spherical reflector</td>
<td>linear polarization (10 dB), high gain, dispersive</td>
</tr>
</tbody>
</table>
2.4 UWB Antenna Requirements for GPR System

Unlike the modern UWB antenna design, which works for 3.1 – 10.6 GHz UWB systems as authorized by the FCC (part 15, Subpart F), the GPR antennas usually pursue different features. [11]

First, a GPR antenna should have maximal bandwidth. To maintain the transmit pulse waveform, the GPR antenna should demonstrate as wide of a frequency bandwidth as possible. In this way, there would be little important frequency component to be chopped off by the UWB antenna.

Second, instead of residing within a certain spectrum, the GPR antenna operates in lower frequency range than modern UWB system. By the nature of attenuation of electromagnetic (EM) wave in path, including transmission loss in the air and dielectric materials, signals of higher frequency are usually more strongly attenuated. As a result, the maximum depth of investigation decreases rapidly when the electromagnetic wave frequency increases. In reality, most sub-surface radar systems operate at frequencies less than 2 GHz or around 2 GHz. Typically, the maximum depths the GPR can penetrate rarely exceed 20 wavelengths. For most GPR used for mine detection, the frequency range is in KHz or MHz, so that the EM wave can go deeper.

The third property of GPR antenna is that it has to well-behaved and consistent across the antenna’s operational band, including the radiation pattern, gain, impedance matching and a requirement for low or no dispersion. Since the GPR pulse should be maintained as clean as possible when transmitted, low antenna dispersion is essential for good
performance of the radar system. This property imposes strong restriction on UWB antenna type and design. As a result, several kinds of antennas with very broad bandwidth cannot be used, including the equiangular spiral antenna, log-periodic antenna and reflector antenna.

Besides, the practical application for under-vehicle mounting also limits the option of UWB antenna. As the space is much confined, the GPR antenna in our design should be relatively low-profile. Also, the dimension of the antennas should not be too large and heavy, so that they can be assembled as an array for pavement condition assessment.

2.5 References


Chapter 3 Cavity-backed Rounded Bowtie Antenna Design

This chapter discusses a low-profile, high-gain reflector antenna, which is composed of a rounded bowtie antenna and a metal cavity as a planar reflector. By performing careful design based on parametric analysis and simulation, this antenna can operate at a broad bandwidth of 0.8 – 3.5 GHz in theory. Such good performance makes it a promising design for the under-vehicle GPR, but the ringing issue should be solved before it can be applied in subsurface radar system.

3.1 Introduction

Ground Penetrating Radar (GPR) is a non-destructive technology that uses electromagnetic radiation of RF/microwave band to detect or image the subsurface structures. It has extensive applications in a variety of fields, such as mine detection in geophysical research, and buried objects exploration in archaeology [1]. Due to its strong ability to precisely locate underground utilities, such as rebar and pipes, and to assess the quality and deterioration of asphalt or concrete pavement and bridge decks, GPR is an attractive non-destructive testing and evaluation method commonly used for health monitoring of roadways and bridge decks. Generally the GPR systems rely on ultra-wideband (UWB) pulse radar technology, in which very broadband radio frequency electromagnetic waves are transmitted and received by antennas. The properties of the subsurface structures or objects can be characterized according to the variation of the reflected signals. GPR antennas are usually ground-coupled and in contact with the
ground. Recently, air-launched GPR systems have been developed to operate above the road surface [2]. One of the difficulties for the air-launched GPR systems is to find a suitable UWB antenna with high gain (ideally flat across the frequency band of interest), unidirectional stable radiation pattern, small electrical size and low profile. The compact, rugged, low-profile physical characteristics are especially important for mounting the system in locations where space is a premium such as the underside of a vehicle.

A number of UWB antenna families have been explored in the past few years, such as monopole [3], resistively loaded dipole [4], planar bulbous elliptical dipole [5, 15], Y-shape bowtie antenna [13] and conical antenna [6]. Frequency-independent antennas include equiangular spiral, conical spiral, log periodic antenna, and other self-complementary structures [7, 8]. Vivaldi and Horn Antennas exhibit good performance at broadband frequency ranges, but the length or height is too large for the under-vehicle mounting [4, 9]. To achieve unidirectional radiation with high gain, the backed cavity was introduced in the wideband bowtie antenna design. Wu reported a resistor-loaded half-ellipse antenna with different backed cavity heights and the input impedance was 100 Ohm [10]. Qu investigated a cavity-backed triangular bowtie antenna and a cavity-backed folded triangular bowtie antenna, both differentially fed by a parallel stripline via a transition from a microstrip line [11, 12]. Such fragile mounting structures do not lend themselves well to applications where rugged use is important, such as a road-speed GPR application using a vehicle.

In this dissertation, a cavity-backed, trimmed-ellipse dipole antenna was designed. A methodical design procedure was employed to maintain the input impedance of this wideband bowtie antenna at 50 Ohm, so that it can be fed directly by a coaxial cable.
Careful parameter analysis during the numerical simulation led to a unidirectional, high-gain antenna with stable radiation patterns. Due to its simple structure, the low profile bowtie antenna is appropriate for application in the under-vehicle GPR systems.

3.2 Bowtie Antenna Geometry And Parametric Analysis
Figure 3-1 shows the schematic of the proposed bowtie antenna. A printed trimmed-ellipse dipole antenna rests on each side of the substrate, of which the thickness is t. The bowtie antenna is backed by a cavity with dimension of L x W x h. To get an optimization design, various design configurations were assessed using the Ansoft HFSS full wave simulator. A lumped port was adopted between the gap of the two bowtie arms to feed the antenna during the simulation.

![Figure 3-1. Geometry of the bowtie antenna design](image)

- A printed bowtie dipole antenna is placed on top of a conductor cavity. The thickness of the substrate is t. The dimension of the cavity is $L \times W \times h$. 

49
In realistic fabrication, a hole was drilled on the substrate to let the inner wire of the coaxial cable reach the upper bowtie arm, as shown in Figure 3-2. Meanwhile, the outer conductor of cable was connected to the lower bowtie arm. The other end of the cable went directly to the power supply through a hole on the bottom of the cavity, making the whole antenna design a simple structure. To prohibit the current flowing on the outer conductor, a clamp-on ferrite suppressor (FerriShield EMI/RFI Suppressors & Ferrites) was applied around the coaxial cable as a type of choke balun [14].

**Figure 3-2. Photo of the proposed bowtie antenna.** The antenna was fed directly by a 50 Ohm coaxial cable.

To achieve wideband 50 Ohm input impedance and good radiation performance, parametric analysis was conducted on several parameters of the antenna, including the dimension of the bowtie arms, the size of the substrate, as well as the height of the cavity. The substrate here was chosen as FR4 microwave epoxy with relative dielectric constant $\varepsilon_r=4.4$ and thickness $t=0.1$ inch (2.54 mm).
The simulated $S_{11}$ and gain varied with the depth of the cavity was shown in Figure 3-3 and Figure 3-4 respectively. In general we consider a good $S_{11}$ match to be ~ -10dB at the frequency of interest. From Figure 3-3 we found that $S_{11}$ decreased at lower frequency and increased at higher frequency as the cavity height increased. Therefore, the input impedance match for the lower frequency range improved when larger-depth cavity was applied. At the same time, the gain decreased with the cavity height, as shown in Figure 3-4. This was due to less reflection from the cavity bottom as its depth increased. To reach a optimization and compromise between impedance match and gain for this bowtie antenna design, the cavity height was chosen as $h=35mm$.

![Figure 3-3. The simulated S11 of the antenna varied with the height of the cavity.](image)

The effect of the dimension of the substrate as well as the cavity was also considered. The $S11$ varied with $L$ and $W$ as they were independently simulated; this is shown in
Figure 3-5 and Figure 3-6 for each of L and W, respectively. From Figure 3-5 it was found that the antenna had better performance below 1GHz as W decreased. However, the higher frequency range (around 2.5 GHz) possessed higher S11 value when W decreased. When W was fixed in Figure 3-6, we can see that the antenna performed better below 1 GHz as L increased. Meanwhile, the S11 figure of merit improved to around 2.5 GHz when L increased. Therefore, there must be tradeoff between high frequency and low frequency end for the dimension of the substrate. Final fabrication of the bowtie antenna was on a FR4 substrate with size of 200mm x 140mm.

Figure 3-4. The simulated gain of the antenna (at Phi=0°, Theta=0°) varied with the depth of the cavity.
Figure 3-5. The simulated S11 when L=200mm and W varied from 132mm to 148mm.

Figure 3-6. The simulated S11 when W=140mm and L varied from 192mm to 208mm.
3.3 Antenna Characterization
The S11 of the proposed bowtie antenna was measured with the Agilent PNA E8364A network analyzer. The experimental result was in good agreement with the simulation. The S11 < -10dB frequency range is 0.89-2.65GHz, as shown in Figure 3-7.

![Figure 3-7. Experimental S11 of the proposed antenna, in comparison with the simulation result.](image)

Apart of wideband input impedance match, the stable radiation pattern is also essential to the antenna application in the GPR system. Figure 3-8 showed the comparison of measured and simulated radiation pattern of the proposed bowtie antenna at 1 GHz. The gain was normalized to the maximum value of the radiation pattern. The experimental results agreed very well with the numerical simulation. The fluctuation at lower gain end was due to fact that the bowtie antenna turned back to the receiver in the anechoic chamber measurement, and the signal of became very weak.
Radiation pattern measurement was performed at various frequencies. Due to limits on anechoic chamber size and the absorber materials in use, it did not permit accurate measurements down to 1GHz frequency. Figure 3-9 presented the E plane and H plane radiation pattern of the proposed antenna at 1.0, 1.2, 1.5, 1.8, and 2.0 GHz. We can observe that the radiation pattern is pretty stable from 1.0 to 2.0 GHz. Simulation results also showed good performance of the antenna at lower frequency range, like 0.8 GHz. Due to experimental limitation, the lowest frequency that can be accurately measured is 1 GHz. The measured results showed that the HPBW for both E plane and H plane radiation pattern around 65°, which is quite appropriate for the under-vehicle GPR system application.
Figure 3-8. Measured and simulated radiation patterns of the proposed antenna at 1 GHz.

(a) Radiation pattern for E plane; (b) radiation pattern for H plane. The gain has been normalized to the maximum value in dB.
The gain of the cavity-backed bowtie antenna was measured in the anechoic chamber with two SAS-571 Double Ridge Guide Horn antennas as reference antennas, as shown in Fig 10. The gain went up from about 6 dB at 1.0 GHz to 10 dB at 2.0 GHz, which matched the simulation results well.

Figure 3-9. Measured radiation pattern at different frequency. (a) Radiation pattern for E plane; (b) radiation pattern for H plane.
3.4 Radar Testing
To test the reflector antenna prototype, two aluminum reflectors were placed side by side and connected to the radar board as transmit and receive antennas. The Antennas were held and scanned 5cm above the ground surface, as shown in Figure 3-11.
Figure 3-11. The reflector antenna with radar board scanned 5cm above the ground surface

Figure 3-12 shows one of the traces collected during the scanning. The curve in the Filter view of Figure 3-12 represents the trace after background removal.

Figure 3-12. One of the trace collected during the scanning
By plotting all the traces collected when scanning over some distance above the floor, it can be found that there might be some metallic bars or rods periodically buried under the floor. But the ringing effect blurred the image, and disguised the subsurface detection, as showed in Figure 3-13.

![Figure 3-13. B-scan 5cm above the floor](image)

3.4 Conclusions
In summary, a 50 Ohm coaxial cable fed cavity-backed bowtie dipole antenna was designed, fabricated and characterized. Anechoic chamber measurements showed good agreement with numerical simulation results, which confirmed our design. Experiments verified that a high gain antenna with stable radiation pattern from 0.8 to 2 GHz was obtained. Due to its simple structure, the low profile bowtie antenna is a good candidate for compact, under-vehicle GPR system application.

3.5 References


4.1 Dispersive antennas

Antenna Dispersion describes the phenomenon that the UWB signal waveform transmitted by an antenna is stretched out into a longer, more distorted waveform.[1] The antenna dispersion will create a ringing effect [2], that is, the radiated signal waveform will be distorted and last more time than its original waveform. So in this dissertation, the antenna dispersion and ringing effect are regarded equally and used alternatively.

In ordinary antenna applications, it is not necessary to consider the dispersion problem.[3] For impulsive radar, however, antenna dispersion is crucial to the system performance. [4] The main reason is that the impulsive GPR relies on time domain reflected signals of targets to perform its detection. Serious ringing effect will disguise the target signal, making it difficult or even impossible to complete the GPR detection task.

Physically, antenna dispersion can be interpreted by a concept of phase center. Phase center is the effective origin of signals radiated from an antenna, and when it moves as a function of frequency, the radiated waveform will be dispersive.[1] Several types of dispersive antennas are summarized below.

One type of dispersive antenna is a frequency independent antenna. Take the log-conical spiral antenna for example, which operates in the frequency range between 1 and 11 GHz. As shown in Figure 4-1. [1] The transmitted and received signals are demonstrated in Figure 4-2. A smaller scale portion radiates high frequency components and a larger scale
portion radiates lower frequency components of a signal at a later time. This is why usually the frequency independent antenna is not suitable to impulsive GPR system.

Figure 4-1. Log conical spiral antenna [1]

Figure 4-2. Transmitted (left) and received (right) voltage waveforms from a pair of Log conical spiral antenna [1]

The second type of dispersive antenna is the UWB antenna that possesses parasitic elements. For instance, the Ridged TEM Horn Antenna. The SAS-571 Double Ridge Guide Horn Antenna works in the frequency range of 700 MHz - 18 GHz, as shown in Figure 4-3. The rugged design of high gain and low VSWR makes this Horn excellent for both immunity and emissions testing. It is also used as a reference antenna for our antenna characterization in the anechoic chamber. The time domain response of the
Figure 4-3. SAS-571 Ridged TEM Horn Antenna

Figure 4-4. Transmitted and received voltage waveforms from a pair of SAS-571 Ridged TEM Horn Antenna. The distance between the two horns is 45cm.
ridged TEM Horn antenna is presented in the following Figure 4-4. It can be found that the ringing effect is obvious, due to the parasitic effect of the parallel rods on the two sides of the antenna. These rods serve as directors, but in the same time, they may introduce extra delayed signals to the original radiated signals.

Another important type of dispersive antenna is the reflector antenna. This type of antenna could also be classified as the antenna with parasitic elements. However, since the reflectors are so popular in practical application, they are cataloged separately. The cavity-backed rounded bowtie antenna is one example, where the total radiation is a combination of the direct signal and the inverted reflected signal from the backed-cavity. A reflected signal is subject to a time delay, and thus the total radiation waveform is distorted. In the next section of this chapter, more details will be provided on how to deal with the ringing in the cavity-backed rounded bowtie antenna configuration.

### 4.2 Theoretical Analysis of antenna dispersion

Under certain assumptions, it is possible to obtain the analytical solutions of radiated fields for some simplified antenna model. These models could also give us some hint to understand the time-domain behavior of the UWB antennas.

The first antenna model is infinitesimal electric dipole mentioned in chapter 2. It is an electrically small antenna, and the temporal variation of all signals is supposed to be negligible during the time for electromagnetic wave to travel across the antenna.

Similar to the calculation in chapter 2, the far field radiation can be represented as [5]

\[
E = \frac{\mu_0}{4\pi r} \dddot{p} \sin \theta \ \hat{\theta} \quad (4-1),
\]

\[
B(r, t) = \frac{\mu_0}{4\pi c r} \dot{p} \sin \theta \ \hat{\phi} \quad (4-2),
\]
where \([p] = p (t - r/c)\).

From the above equations (4-1) and (4-2), it can be found that the radiated electric and magnetic field are proportional to the second-order derivative of the electric dipole moment.

Therefore, if the input signal of electric dipole is a Gaussian function with characteristic time \(\tau\), as shown below, [5]

\[
p(t) = p(t)\hat{z} = p_0 e^{-(t/\tau)^2} \hat{z} \quad (4-3)
\]

then the radiated field will display a pulse waveform of second-order Gaussian:

\[
\ddot{p}(t) = \ddot{p}(t)\hat{z} = -\frac{2}{\tau^2} (1 - 2\left(\frac{t}{\tau}\right)^2)p_0 e^{-(t/\tau)^2} \hat{z} \quad (4-4)
\]

In the following the thin-wire dipole antenna will be discussed, where time of light traveling across the antenna length \(h\) should be taken into account. \(I_s(t)\) represents the source current.

First let us consider a traveling-wave dipole, which is loaded with reflectionless terminations at both ends, such as resistors. The radiation electric field of the traveling-wave dipole can be written as [5]

\[
E = \frac{\mu_0 c}{4\pi r} \left( \frac{\sin\theta}{1 - \cos\theta} \left[ I_s \left( t - \frac{r}{c} \right) - I_s \left( t - \frac{r}{c} - h/c(1 - \cos\theta) \right) \right] + \frac{\sin\theta}{1 + \cos\theta} \left[ I_s \left( t - \frac{r}{c} \right) - I_s \left( t - \frac{r}{c} - h/c(1 + \cos\theta) \right) \right] \right) \hat{\theta} \quad (4-5)
\]

When the current of source is a Gaussian function

\[
I_s(t) = I_0 e^{-(t/\tau)^2} \quad (4-6)
\]
the radiated electric field can be plotted as a function of looking angle $\theta$, as shown in Figure 4-5. The radiation is composed of three spherical wavefronts that are centered on the source (W1), on the top termination (W2), and on the bottom termination (W2').

Next is the dipole with total reflection of the charge at the ends, which can be named as standing-wave dipole. In this case, the radiated electric field is: [5]

$$
E = \frac{\mu_0 c}{2\pi r \sin \theta} \left\{ I_s \left( t - \frac{r}{c} \right) + I_s \left( t - \frac{r}{c} - 2h/c \right) - I_s \left( t - \frac{r}{c} - h/c(1 - \cos \theta) \right) - I_s \left( t - \frac{r}{c} - h/c(1 + \cos \theta) \right) \right\} \hat{\theta} \quad (4-7)
$$

The standing-wave dipole antenna can be interpreted as a combination of four basic traveling-wave elements, and the corresponding radiation is composed of four spherical wavefronts as shown in Figure 4-5.
Figure 4-5. Radiated electric field of traveling-wave and standing-wave dipole antennas (a) Left - radiated electric field of traveling-wave dipole antenna; (b) Right - radiated electric field of standing-wave dipole antenna. Both excitation signals are Gaussian pulse.[5]

Though books such as [1] and [2] talked about the non-dispersive antennas, the pure traveling-wave antenna without reflections cannot be realized in practice. It can only be constructed approximately.

4.3 Experimental characterization of antenna dispersion
Generally there are two approaches to characterize the antenna dispersion, one is in time domain and the other is in frequency domain.

In the frequency domain measurement, the network analyzer is used to sweep from close to zero to some GHz, and the complex response of the system is measured. The upper end
The configuration of frequency domain measurement for antenna dispersion.

**Figure 4-6. The configuration of frequency domain measurement for antenna dispersion**

The experimental setup in block diagram is shown in the Figure 4-6 [6]. The transmitter and receiver antennas are connected to the port 1 and port 2 of the network analyzer separately. The $H_{CH}(f)$ is the transfer function of the channel for signal transmission from transmitter to receiver antennas, which is usually the space of air between those antennas.

The transfer function of the total system is

$$H(f) = H_{TX}(f)H_{CH}(f)H_{RX}(f) = \frac{y(f)}{x(f)} = S_{21} \quad (4-8)$$

The impulse response of the system can be computed by IFFTs (Inverse fast Fourier transform).

$$h[t] = Re\{IFFT[H(f)]\} \quad (4-9)$$
Note that this calculation is for impulse response. To get response of arbitrary incident pulse, a convolution process should be performed.

The other approach is to measure the time domain response directly using a fast sampling oscilloscope. A narrow pulse is excited on the transmitting side, and a waveform on the receiving side will be captured using an oscilloscope. Meanwhile, the original transmitter signal elicited from a T-splitter is also captured in the oscilloscope for comparison. Figure 4-7 shows the experimental configuration to measure the time domain response of the UWB antenna.

![Figure 4-7. The schematic of experimental setup to test time domain response of UWB antenna](image)

This method is pretty straightforward, and it is used thoroughly in the antenna development and radar testing.

**4.4 Simulation and Modeling**

As discussed above, the frequency domain and time domain measurements provide good methods to characterize and verify the antenna performance on dispersion. Unfortunately
these methods cannot predict antenna time domain response for the UWB antenna design for radar system. From [6], it is found that no antenna practical geometry is considered.

The antenna dispersion characterization, however, do provides us some hint on how to foresee the antenna time domain response. By using computational electromagnetic methods in the time domain, such as FDTD (Finite-difference time-domain) method, it is possible to simulate the UWB antenna time domain response for arbitrary excitation pulse waveform. Also, the direct observation and comparison with experimental results is possible, which provides a powerful aid for the impulsive radar antenna design.

In our design process, the CST Microwave studio software is utilized, since it uses the time domain computational EM method. As an example, the time domain response of the rounded bowtie antenna is simulated using CST, as shown in Figure 4-8. Meanwhile, the time domain response of the bowtie antenna with ridged edges are also demonstrated here as a comparison, as shown in Figure 4-9.

![Figure 4-8. Time domain response modeling of the rounded bowtie antenna](image)
Figure 4-9. Time domain response modeling of the bowtie antenna with ridged edges

From Figure 4-10 and Figure 4-11, it is clearly shown that the rounded bowtie antenna has much lower ringing compared to the ridged-edge bowtie antenna.

Figure 4-10. Time domain response of the rounded bowtie antenna
In the simulation, the third order Gaussian pulse is used as the exciting signal, as shown in Figure 4-12.

**Figure 4-12. Exciting signal for face-to-face time domain response simulation**

### 4.5 Brief Overview on Improvement of Cavity-backed Rounded Bowtie Antenna Design

Operating as one critical component of a GPR system, the ultra-wideband (UWB) antenna is used to transmit and receive short pulses with small late-time ringing. The ringing effect mainly arises from the multiple reflections between the antenna ends and the feeding point. It is essential to prohibit such effect, since it may mask the reflected signal of the target. Situations become even worse when the buried object is shallow underneath the ground surface, for it is hard to distinguish the target signal from the
direct coupling between antennas and ground surface reflection. Therefore, the time
domain response is crucial to the performance of the GPR system in addition to the gain,
stable radiation pattern, and appropriate dimensions of the UWB antenna.

A variety of UWB antennas have been explored in the past few decades, including the
Horn antenna, Vivaldi antenna [7], frequency-independent Archimedean-spiral, and log-
spiral antenna. The spirals possess large bandwidth but their ringing is serious, due to a
phase delay between different frequency components of the radiated field [8]. The bowtie
antenna is the most popular planar dipole used in GPR applications using ground-coupled
antennas. The electromagnetic energy can be effectively coupled to the ground for this
broadband antenna, as the bowtie can be easily placed close to the ground surface.
Usually the resistive loading technique is used to suppress the late-time ringing [9, 10].
The bowtie antenna is connected to a ground plane through lumped resistors to damp the

Lestari et al. made use of a volumetric microwave absorber as resistive loading and
narrow slots on the antenna surface as capacitive loading to improve the pulse radiation.
This technique allows transmission of short transient pulses with very small late-time
ringing and relatively high radiation efficiency, but the antenna which worked above
0.5 GHz possessed a large dimension of 50 cm long with a flare angle of 90° [12].
Nishioka studied the use of a ferrite coating for resistor-loaded bowtie antennas [13]. The
antennas were covered with a rectangular conducting cavity of which inner walls were
coated partially or fully with ferrite absorber. The effects of the ferrite absorber on the
GPR characteristics were theoretically investigated using surface impedance boundary condition (SIBC) for the frequency-dependent ferrite sheet backed by a perfect conductor. From the finite-difference time-domain (FDTD) results the paper concluded that significant improvement of the resistor-loaded antenna characteristics cannot be achieved by ferrite absorber. Uduwawala et al. also modeled a GPR antenna system using the three-dimensional FDTD technique [14]. The antenna was resistor-loaded and shielded by a rectangular conducting cavity. The effect of adding a wave-absorbing coat to the shield was also studied. They used a perfectly matched layer (PML) absorbing boundary condition (ABC) to simulate the absorbing layer in order to investigate its applicability together with planar dipoles. Their simulation concluded that adding a wave-absorbing layer to the inner surfaces of the antenna cavity cannot improve the performance of planar dipoles.

In this section, a low-profile, cavity-backed bowtie antenna was fabricated and tested for a ground penetrating radar system. Since resistive loading would visibly reduce the radiation efficiency, only ferrite-rubber absorber was attached to the inner surface of the cavity. Anechoic chamber measurements indicated that the proposed antenna exhibited broadband impedance match, and displayed stable radiation pattern at the wide frequency range. Time domain response measurement and theoretical analysis verified that the attached ferrite-rubber absorber can effectively reduce the dispersion while maintaining the relative higher signal amplitude, compared to the cavity-backed bowtie antenna without absorber. GPR experiments using the proposed antenna were also performed, and the rebar buried in the sand box was detected successfully with antennas 305mm above
the sand surface. With a simple structure, the low-profile bowtie antenna coated with absorber is a good candidate for an air-coupled under-vehicle GPR system application.

4.6 Characterization of cavity-backed Bowtie Antenna

4.6.1 Design Summary

The bowtie antenna was fabricated on FR4 microwave epoxy with a relative dielectric constant $\varepsilon_r=4.4$ and thickness $t=0.1$ inch (2.54 mm), as shown in Figure 4-13. The size of the substrate was $200 \, \text{mm} \times 140 \, \text{mm}$, and the height of the metal cavity on the bottom of the substrate was chosen as $h=39 \, \text{mm}$. To achieve wideband input impedance and good radiation performance, parametric analysis was conducted on several parameters of the antenna in the Ansoft HFSS full wave simulator, including the dimension of the bowtie arms, the size of the substrate, as well as the height of the cavity. This antenna geometry was optimized for the performance of impedance matching, gain, and time domain response.

![Figure 4-13. Photo of the proposed cavity-backed bowtie antenna.](image)

The antenna was fed by a 50 Ohm coaxial cable. A ferrite-rubber absorber was attached on the bottom of the cavity.
This antenna was fed directly by a 50 Ohm coaxial cable. A hole was drilled on the substrate to allow the inner wire of the coaxial cable to reach the upper bowtie arm. Meanwhile, the outer wire of cable was connected to the lower bowtie arm. The other end of the cable went directly to the signal source through a hole on the bottom of the cavity, making the whole antenna design a simple structure.

In Figure 4-13, a ferrite-rubber absorber was attached on the bottom of the cavity to prohibit the ringing effect for this cavity-backed bowtie antenna. Before the exploration of the time domain response for the UWB antenna, the characterization of this antenna was conducted in an anechoic chamber.

![Figure 4-14](image.png)

**Figure 4-14. Experimental $S_{11}$ of the cavity-backed bowtie antenna coated with ferrite-rubber absorber.**

Figure 4-14 illustrates the $S_{11}$ of the cavity-backed bowtie antenna coated with ferrite-rubber absorber. The $S_{11}$ was measured with the Agilent PNA E8364A network analyzer. The $S_{11}<-10$ dB frequency range covered from 0.61 up to 2.85 GHz, which can
contribute positively to the GPR at low frequency performance. Parameters of ferrite-rubber absorber provided by manufacturers were imported to HFSS to simulate the behavior of the proposed antenna, but it is difficult to match experimental test results perfectly due to the modeling inaccuracies. Therefore, only experimental results were demonstrated here.

4.6.2 Radiation pattern considerations

In addition to matching the broadband input impedance, a stable radiation pattern is also essential to the antenna application in the GPR system. Figure 4-15 shows the measured E plane and H plane radiation patterns at 1.8 GHz for the cavity-backed bowtie antenna coated with ferrite-rubber absorber. The corresponding results of the cavity-backed bowtie antenna without absorber were also presented for comparison. It’s shown that the E plane radiation pattern (Figure 4-15 (a)) remained the same after the ferrite-rubber absorber was applied. The half power (-3dB) beam width (HPBW) was 66° for the E plane radiation pattern. On the other hand, the H plane radiation pattern (Figure 4-15 (b)) for the bowtie with ferrite-rubber absorber became broader, compared to the results of bowtie without absorber. The HPBW of bowtie without absorber evolved from 60° to around 66° for the bowtie with absorber. As a result, the radiation pattern of the cavity-backed bowtie antenna coated with ferrite-rubber absorber became more symmetric for E plane and H plane radiation.

E plane and H plane radiation patterns for the cavity-backed bowtie with absorber at different frequencies were also measured, as shown in Figure 4-16. We find relatively
stable radiation patterns from 1.2 to 2 GHz. The measured results showed that the HPBW for both E plane and H plane radiation pattern were around 66°, which is quite appropriate for the under-vehicle GPR system application. In both Figure 4-15 and Figure 4-16, the gain was normalized to the maximum value of the radiation pattern. The fluctuation at lower gain end was due to experimental limitation in the anechoic chamber.

Figure 4-15. Measured radiation patterns of the cavity-backed bowtie antenna coated with/without ferrite-rubber absorber at 1.8 GHz. (a) radiation pattern for E plane; (b) radiation pattern for H plane. The gains have been normalized to the maximum value in dB.
The gain of the cavity-backed bowtie antenna with ferrite-rubber absorber was obtained in the anechoic chamber with two SAS-571 Double Ridge Guide Horn antennas as reference antennas, as shown in Figure 4-17. The gain of the bowtie with absorber ranged from 0.42 dB at 1.2 GHz to 2.1 dB at 1.8 GHz, lower than the corresponding values of the cavity-backed bowtie without any absorber. This is due to the fact that the microwave absorbers on the inner faces of the antenna cavity reduced the multiple reflections effectively.

Figure 4-16. Measured radiation patterns of the cavity-backed bowtie antenna coated with ferrite-rubber absorber at 1.2, 1.5, 1.8 and 2.0 GHz.
(a) radiation pattern for E plane; (b) radiation pattern for H plane. The gains have been normalized to the maximum value in dB.
Figure 4-17. The measured maximum gain of the cavity-backed bowtie antenna with/without ferrite-rubber absorber from 1.2 to 2.0 GHz.

4.7 Time Domain Response Measurement

The dispersion (ringing effect) is critical for the UWB antenna application in GPR system, since it may mask the targets by reflections from the ground or the antenna structure itself. A good way to assess the dispersion is to look at the time domain response to the signals emitted by an antenna. If the received signals are significantly distorted compared to the transmitted signals, then we can conclude that the dispersion is large. The setup of our experiment to measure the time domain response of the UWB antenna has been seen in Figure 4-7.

During the testing, a Vivaldi antenna (Figure 4-18) was chosen as transmitter (Tx), since it showed good time domain response as shown in Figure 4-19. The measurements were performed when the distance of Transmitter and Receiver (Rx) antenna was 30 cm. The input pulse was shown at the bottom of Figure 4-19 for comparison. Figure 4-19
demonstrated time domain responses using different antennas as receivers, including a Vivaldi, bowtie (no cavity or absorber were applied), and the cavity-backed bowtie antenna with/without ferrite-rubber absorber. It is observed that the dispersion is the smallest when no cavity was attached to the bowtie antenna. The utilization of a metal box as cavity introduced much ringing for the bowtie antenna. Meanwhile, the application of absorber reduces the dispersion effectively, due to the fact that they can reduce the surface current on the cavity walls, leading to fewer reflections from the antenna structure itself.

Figure 4-18. The Vivaldi antenna as a reference of transmitter (Tx) antenna in the time domain response measurement

It is interesting to find that the maximum value (0.262 mV) of the voltage amplitude for the cavity-backed bowtie antenna without absorber was greater than the value (0.163 mV) of the cavity-backed bowtie antenna with ferrite-rubber absorber, which approximated the value (0.162 mV) of the bowtie only case (when no cavity or absorber were applied).
Figure 4-19. Time domain responses with different receiver antenna configurations, including Vivaldi, bowtie (no cavity or absorber were applied), and the cavity-backed bowtie antenna with/without ferrite-rubber absorber.
4.8 Theoretical Analysis
Here a simplified model was constructed to verify that the measured voltage amplitude in the time domain response measurement for different receiver antenna configurations are consistent. Shown in Figure 4-20 is a dipole antenna with a distance d above the infinite conducting plane. The conducting plane served to model a reflector like the metal cavity. In computation, the effective origin of the reflected signal was an image antenna at a distance d below the conducting plane [1].

![Diagram of dipole antenna above conducting plane](image)

**Figure 4-20. A dipole antenna above a conducting plane to model the backed-cavity bowtie antenna with/without absorber.**

For simplicity, consider the incident wave as harmonic wave. A pulse signal can be interpreted as a superposition of harmonic waves at different wavelengths. When no reflector plane was applied, the radiation field of a dipole antenna can be written as:

\[
\bar{E} = V_0 \sin \theta \frac{e^{-jkr}}{r} \hat{\theta}, \quad (4-10)
\]

where \(V_0\) is constant, and \(k = 2\pi/\lambda\) is the wave number of electromagnetic wave in the vacuum, \(\lambda\) the wavelength in free space. The maximum directivity of the dipole is \(D = 1.76\) dB.
As the conducting plane is applied, a signal from the image dipole must travel additional length \(l = 2d \cos \varphi\), which introduces a phase delay of \(\theta_0 = \omega l / c\), where \(\omega = kc\) is the speed of light in free space. Therefore, the total radiation is a combination of the direct signal and the inverted reflected signal:

\[
\vec{E}_t = (1 - e^{-j\theta_0})\vec{E}, \quad (4-11)
\]

Here the relative direction of the electric field to the ground plane was ignored for simplicity.

The total radiation field can be larger than the direct component of the dipole antenna when

\[
|1 - e^{-j\theta_0}| = \sqrt{(1 - \cos \theta_0)^2 + \sin^2 \theta_0} > 1 \quad (4-12)
\]

For the experiment \(\varphi = 0\). The above inequality leads to \(\lambda < 12 \cdot d\). The height \(d = 39\)mm corresponds to a maximum wave length of 0.468m, i.e., a minimum frequency of 0.64 GHz. Therefore, the observed voltage amplitude for the cavity-backed bowtie antenna should be greater than that of the bowtie only case. But at the same time, the ringing effect of the cavity-backed antenna became worse due to the phase delay introduced by the reflector. From the closeness between the measured voltage amplitude 0.163mV for the cavity-backed bowtie antenna with ferrite-rubber absorber and that of bowtie only case (0.162mV), it can be concluded that the absorber worked effectively in reducing the surface current on the inner wall of the cavity, and hence alleviated the ringing.
A more detailed calculation shows that the directivity of the horizontal dipole above an infinite perfect electric conductor can be written as [3]

\[
D_t = \begin{cases} 
\frac{4\sin^2(kd)}{R(kd)}, & kd \leq \frac{\pi}{2} (d \leq \lambda/4) \\
\frac{4}{R(kd)}, & kd > \frac{\pi}{2} (d > \lambda/4)
\end{cases}, \tag{4-13}
\]

where

\[
R(kd) = \frac{2}{3} - \frac{\sin(2kd)}{2kd} - \frac{\cos(2kd)}{(2kd)^2} + \frac{\sin(2kd)}{(2kd)^3}. \tag{4-14}
\]

The maximum value of the directivity is 7.5 dB for small values of d when d is less than \(\lambda/4\), which can match our experiments well when considering the simplicity of the model. The physical shape of the bowtie antenna as well as the cavity as a wrapped-up conducting plane should be taken into account when more accurate analytical study is needed to perform.

### 4.9 Ground Penetrating Radar Experiment

In Figure 4-21 one pair of the cavity-backed bowtie antenna attached with ferrite-rubber were tested with the ESS (Earth Science System) GPR. The antennas were positioned 12 inches above the surface of the sand box with one rebar spanning the middle of the sandbox. The metal bar was placed 2 inches beneath the sand surface. During the testing process, the antennas took one scan per inch, and the process lasted 34 scans. Hyperbolas were visible in the plotted data representing the metal bar in the sand, as shown in Figure 4-22.
Figure 4.21. One pair of the cavity-backed bowtie antenna attached with ferrite-rubber were tested with the ESS (Earth Science System) GPR. The wood shelves 12 inches above the sand box were used to hold the antennas during the scanning process. A metal bar was buried 2 inch underneath the sandbox surface and a metal sheet was placed on the bottom of the sand box.

In Figure 4-22, a GPR profile (B-scan) was plotted from the scanning data composed of 34 traces showing a hyperbola image. From the image we can find that the effect of antenna direct coupling, sand surface reflection, metal sheet reflection on the bottom of sand box, and the metal bar in the sand are clearly demonstrated.
Figure 4-22. Hyperbola image plotted from the scanning data. In the image the effect of antenna direct coupling, sand surface reflection, metal sheet reflection on the bottom of sand box, and the metal bar in the sand were demonstrated.

4.10 Conclusions
A low-profile, cavity-backed bowtie antenna attached with ferrite-rubber absorber was fabricated and tested for ground penetrating radar system. Anechoic chamber measurements indicated that the proposed antenna exhibited broadband impedance match from 0.6 up to 2.85 GHz, and displayed stable radiation pattern at the wide frequency range. Time domain response measurement and theoretical analysis verified that the attached ferrite-rubber absorber can effectively reduce the dispersion while maintaining
the relative high signal amplitude, compared to the cavity-backed bowtie antenna without absorber. GPR experiments using the proposed antenna were also performed, and the rebar buried in the sand box had been detected successfully. Due to its simple structure, the low profile bowtie antenna coated with absorber is a good candidate for air-coupled under-vehicle GPR system application. Further research will be conducted to reduce the ringing effect in the UWB antenna, as dispersion is an essential factor for the resolution of the GPR system.

4.11 References


Chapter 5 Compact Bowtie Slot antenna Design for Ground Penetrating Radar Application

5.1 Introduction

Ground Penetrating Radar (GPR) uses electromagnetic radiation to detect or image the subsurface structures of interest. It is an attractive non-destructive instrument for mine detection and buried objects exploration in archaeology. In the civil engineering field, GPR has been employed in probing subsurface conditions in asphalt, concrete and other materials, as well as structural assessment such rebar and water pipes [1], mainly in the low GHz frequency band.

Generally GPR systems rely on impulsive radar technology, in which ultra-wideband (UWB) antennas are needed to transmit and receive broadband radio frequency electromagnetic waves. Such technology presents significant restrictions on the antenna selection and development. To attain a high resolution and good penetration, the antenna is expected to demonstrate broad bandwidth, high gain across the frequency band of interest, unidirectional radiation pattern, and low antenna dispersion (for impulsive radar systems). Considering the space-confined application, a low-profile antenna is needed, which usually means a small electrical antenna with low gain. Furthermore, In order to develop the air-coupled GPR which operates above the road surface, the requirement on the antenna development is more critical, since the radiated energy will be reflected mostly on the ground surface. As a result, the resolution and penetration of the air-coupled GPR system will be decreased seriously.
A number of UWB antennas have been investigated for the GPR application in the past few years. Horn Antenna and Vivaldi (planar Horn) antenna exhibit good performance at broadband frequency ranges, but their dimensions are too bulky for portable application [2]. Bowtie antenna is also widely used in GPR systems due to its low profile and small antenna dispersion. The radiation gain of such antenna, however, is usually less than 4 dBi, which could limit the penetration capability of the radar system. Lestari et al developed a bowtie antenna with very small late-time ringing and relatively high radiation efficiency, using volumetric microwave absorber as resistive loading and narrow slots on the antenna surface as capacitive loading. However, the antenna displayed relatively low gain, less than 4 dBi at most of the operation frequency band from 0.5 to 3.5 GHz, and it possessed a large dimension of 50 cm long with a flare angle of 90° [3]. Frequency-independent antenna, such as Archimedean-spiral and log-spiral antenna, exhibits serious antenna dispersion, which is not suitable for transmission of short transient pulses with very small late-time ringing.

Slot antennas have also been proposed for UWB application [4, 5], but their difficulty in achieving low-dispersion limits their application in radar system.[6] Barnes developed a promising continuously taped magnetic slot antenna with typical gain of 5-6dBi. It is reported to exhibit good performance in the Time Domain Corporation’s RV1k through-wall UWB radar, but no more details have been illustrated [7]. Sagnard et al have presented a bowtie slot antenna including triangular metal stubs at the centre for impedance matching to work in a wide frequency band 0.4 – 1.5 GHz for GPR
application. The antenna geometry displays a length close to 50 cm and a width around 22 cm. Testing on a sand box made of a 25 cm thick sand layer was performed, and it reported that further measurements need to be done to pursue the validation of the bowtie slot antenna designed [8].

The purpose of this research is to develop a highly compact bowtie slot antenna for impulsive GPR system, especially for air-coupled application. The proposed bowtie slot antenna is eyes-shaped and fed by 50 Ohm coplanar waveguide (CPW) structure, which is quite suitable for system integration. The operation frequency range of the bowtie slot antenna is 1.1 – 3.5 GHz. The compact, rugged, low-profile physical characteristics are especially important for mounting the system in locations where space is a premium such as the underside of a vehicle.

Careful parameter analysis during the numerical simulation led to a unidirectional, high-gain antenna with stable radiation patterns. Due to its simple structure, the low profile bowtie antenna is appropriate for application in the under-vehicle GPR systems.

5.2. Antenna Configuration

The geometry of the proposed bowtie slot antenna is illustrated in Figure 5-1. Figure 5-1 (a) shows the antenna schematic with size of 106.7 mm x 68mm. When assembled with two metallic backed-boxes for GPR test, the total dimension of one antenna is 10.7 cm x 7 cm x 5 cm, as shown in Figure 5-1 (c).

The bowtie slot antenna actually possesses eye-shaped slots, as shown in Figure 5-1(a) and (c). The eye-shaped slot is rounded from a quadrilateral with two diagonals.
=15mm and $l_a + l_b = 34\text{mm} + 19\text{mm} = 53\text{ mm}$. The width of the edge conductor is $l_e = 0.35\text{mm}$.

The Coplanar-Waveguide (CPW) feed structure of the proposed antenna is zoomed in and shown in Figure 5-1(b). The width of the trace is $s = 1.6\text{mm}$. The total length of CPW trace is $w_a = w_b + w_c = 14\text{mm} + 20\text{mm} = 34\text{mm}$, while $w_c = l_t + l_c = 13.3\text{mm} + 6.7\text{ mm}$. The CPW feed structure is made of three components. CPW trace of length $w_b$ is connected directly to the antenna’s feed point. CPW trace of length $l_c$ is touched directly by the standard 50 Ohm SMA connector at the edge of Printed Circuit Board (PCB). The trace of length $l_t$ is a taped CPW, and the widths of slot gaps are $d = 1\text{mm}$ and $g = 0.25\text{mm}$.

Figure 5-1(c) shows the assembly of bowtie slot antenna for GPR test. The proposed antenna is fabricated with low-cost FR4 PCB of thickness 0.76 mm. The bowtie slot antennas are backed with two metal cavities. Layers of carbon-adulterated foam absorber are used to prohibit the notorious reflection which will introduce signal distortion for impulse GPR system. For the testing convenience, the two bowtie slot antennas are hold in a larger plastic box.
Figure 5-1. Bowtie slot antenna geometry. (a) Proposed bowtie slot antenna schematic. $l = 106.7$ mm, $w = 68$ mm, $ws = 15$ mm, $la = 34$ mm, $lb = 53$ mm, $le = 0.35$ mm, $wa = 34$ mm, $wb = 14$ mm, $wc = 20$ mm. (b) the Coplanar-Waveguide (CPW) for antenna feeding. $lt = 13.3$ mm, $lc = 6.7$ mm, $d = 1$ mm and $g = 0.25$ mm. The width of the trace is $s = 1.6$ mm. (c) The assembly of bowtie slot antenna for GPR test with dimension of 10.7 cm x 7 cm x 5 cm. The Two antennas are backed with metal cavities, and hold in a larger plastic box.
5.3 Antenna Parameter Analysis and Characterization

To achieve good performance of the bowtie slot antenna, a methodical design procedure is employed to attain broadband impedance match, suitable radiation pattern, and relatively high gain. Low antenna dispersion is also required for impulsive GPR application. A variety of parameters as illustrated in Figure 5-1 are optimized to arrive at a promising compact GPR antenna design.

Considering the tradeoff between the penetrating ability and resolution of electromagnetic waves, the target central frequency of the GPR system is set to be around 2 GHz. The bandwidth of the GPR antenna is supposed to be as wide as possible. Therefore, the good impedance match is pursued as one of the key factor for the antenna design.

As a first step to develop an impedance-matched antenna over a wide frequency band, a high gain slot antenna is inspected as a prototype of the final design. Then a low insertion loss taped CPW feed structure as shown in Figure 5-1(b) can be attached to one side of the prototype, which can maintain the high radiation efficiency of the proposed antenna. Subsequently, an exhausted investigation of antenna parameters is conducted to reach optimized performance.

The overall dimension of the antenna can make significant influence on the impedance matching. Figure 5-2 and Figure 5-3 demonstrate the simulated S11 varies with the length and width of the bowtie slot antenna. From Fig. 3 it is found that the proposed antenna with smaller length exhibits wider bandwidth. The length of the antenna, however, cannot be shorten than $2 \times (la + lb)$, since there will be gap between the upper and lower side of
the bowtie slot antenna. It will change the antenna impedance dramatically, as the case $l = 105\text{mm}$ in Figure 5-2.

![Graph showing simulated S11 vs. frequency for different substrate lengths](image)

**Figure 5-2.** Simulated S11 of the bowtie slot antenna varies with the length of the substrate $l$. 
Figure 5-3. Simulated S11 of the bowtie slot antenna varies with the width of the substrate $w$ while maintaining $wa = 34\text{mm}$.

The return loss (described by S11) of the antenna behaves differently at lower and higher frequency end. Smaller width can introduce better impedance match in the lower frequency end, while acts conversely in the higher frequency end as shown in Figure 5-3. During the investigation the length of CPW feed structure $wa$ is constant.

In addition, the thickness of the FR4 substrate imposes substantial impact on the antenna performance. S11 curve in Figure 5-4 and the dependence of gain on substrate thickness in Figure 5-5 demonstrate that thinner substrate can enhance the antenna performance with broader bandwidth and higher gain. Thus the substrate thickness $t_{\text{sub}} = 0.762 \text{ mm}$ is selected.
Figure 5-4. Simulated S11 of the bowtie slot antenna varies with the substrate thickness $t_{\text{sub}}$.

Figure 5-5. Simulated gain of the bowtie slot antenna varies with the substrate thickness $t_{\text{sub}}$. 
The experimental S11 of the proposed bowtie slot antenna is obtained with the Agilent PNA E8364A network analyzer, as illustrated in Figure 5-6. The antenna dimension of length $l = 106.7\text{mm}$, width $w = 68\text{mm}$, $wa = 34\text{mm}$ and FR4 substrate thickness 0.762mm are chosen for the final antenna manufacture. The experimental result is in good agreement with the results of Ansoft HFSS simulation. The S11 $<-10\text{dB}$ frequency range is 1.4 - 3.5GHz.

![Figure 5-6](image)

**Figure 5-6. Experimental S11 of the proposed antenna, which is in good agreement with the simulation result.**

The gain of the cavity-backed bowtie slot antenna was shown in Figure 5-7. Due to the limitations of our the anechoic chamber and other instrumental conditions, only radiation gain of three frequency points are measured with two SAS-571 Double Ridge Guide Horn antennas as reference antennas. From Figure 5-7 it is found that the experimental results can agree with the simulation well. The maximum gain is expected to be about 6 dB at 2.75GHz.
Figure 5-7. Experimental measured gain of the proposed antenna at 1.5, 1.75 and 2.0 GHz, which is in good agreement with the simulation result. The maximum gain is close to 6 dB at 2.75GHz.

5.4 Antenna Dispersion

Since the property of antenna dispersion is critical to the GPR performance, the time domain response of the Bowtie slot antenna is simulated and experimental characterized. The GPR radar excitation signals have been extracted from the GPR circuit board, as shown in the Figure 5-8. Ideally, the pulse received on the receiver end should be the same as the transmitter side, but in reality, there is always some degradation from each electric component or connector. Since the total system performance is concerned, the extracted excitation pulse will be used as the input to the simulation model.
By importing the Radar excitation signal directly to the simulation model, it provides a useful and powerful method for the UWB antenna design and verification. Figure 5-9 is the excitation pulse used in the simulation model, and the face-to-face Bowtie slot antenna model for time domain response characterization is shown in Figure 5-10.

Figure 5-8. Radar excitation signal extracted from the GPR circuit board

Figure 5-9. Excitation pulse used in the simulation model for Bowtie slot time domain response
The simulation and experimental results of Bowtie slot antenna time domain response are shown in Figure 5-11, together with the original excitation signal. The transmitter excitation signal has been rescaled in order to plot it together with the waveform captured on the receiver side.

It can be seen that the simulation and experimental results match very well. Compared to the original excitation pulse, it can be found that the received signal demonstrates waveform distortion and stretching slightly. But practical testing with GPR shows that this design is pretty successful. It displays good resolution while maintaining relatively compact size.
Figure 5-11. Simulation and experimental results of Bowtie slot antenna time domain response, together with the original excitation signal.

Also, there is some method to prohibit the ringing effect. One way is to reduce the common mode current distribution on the outer surface of the coaxial feed by ferrite choke, as shown in Figure 5-12. Such current comes from the unbalanced transition between the coaxial cable feed and the CPW transmission line.
The experimental results of Bowtie slot antenna time domain response are shown in Figure 5-13 when ferrite chokes are used. When one ferrite choke is used, the ringing is reduced slightly. The effect becomes much obvious when two ferrite chokes are wrapped around the coaxial cable. It can be found that the waveform distortion in the later time of the received signal mainly arises from the common mode current. Such current will radiate as well, and cause unwanted delayed oscillation.
Figure 5-13. The experimental results of Bowtie slot antenna time domain response when ferrite chokes are used

The GPR testing with the Bowtie slot antenna will be discussed in the next chapter together with different UWB antennas developed for VOTERS GPR system.

5.5 References


Chapter 6 Antenna testing with Compact Ground Penetrating Radar Systems

This chapter serves as a summary of different types of antennas developed for the VOTERS GPR (ground penetrating radar) project, including rounded bowtie antenna, bowtie slot antenna, and Vivaldi antenna (planar horn antenna). New models of Vivaldi antenna will also be presented with GPR testing results. More importantly, how the GPR system works with the UWB antennas will be covered in detail, which is supposed to provide useful reference and guidance for antenna design for practical impulsive GPR system.

6.1 Background

This background section will discuss the FCC constraints, antenna specifications, a review of the state-of-the-art in air-coupled GPR antennas for civil engineering applications [1-2], and the general design approach taken [3].

FCC Constraint

Apart from considerations of antenna performance, the FCC (Federal Communications Commission) imposes regulations on UWB systems that also have to be taken into account. Since the FCC 02-48 specification exerts strict limits on maximum average emission above 960 MHz, especially at frequency range of 960-1610 MHz. This specification restricts the power level to as low as -65.3 dBm as illustrated in Figure 6-1. The frequency band above 2 GHz was considered to be the key frequency of operation [4]. Furthermore, in order to maintain good penetrating ability of the radar, the lower frequency band below 900 MHz was also considered as a secondary option.
Antenna Specifications

Antenna specifications have been derived based on the mandate of creating an air-coupled, road-speed GPR system, to operate across multiple channels spanning the width of the vehicle, and will be FCC compliant to eventually receive certification. A few key parameters include:

(1) Frequency range – the antenna frequency range must match the GPR electronics and together should meet/exceed the FCC requirements presented above;
(2) Polarization – linear – the choice is between reduced complexity (see size below) and increased performance. Considering the physical features that are common for road survey, the linear polarization was chosen;

(3) Input matching – the $S_{11}$ of the antenna must be $|S_{11}| > 10$ dB for the frequency band chosen in (1) above; this is to ensure that 90%+ of the power delivered to the antenna radiates towards the target of interest;

(4) Gain – the antenna gain provides amplification to the signal delivered by the GPR transmitter and provides a boost for the signal returned from the target; acceptable gains were seen as $\sim 3 – 10$ dB;

(5) Beamwidth – the -3dB beamwidth relates to the physical footprint of the radiated energy where half of the total energy is delivered. Beamwidth and gain are inversely related (higher gain results in narrower beamwidth) and for the purposes of the design a beamwidth of 40 – 100 degrees was desired;

(6) Size – the antenna physical dimensions should be as low as possible to allow for compact packaging/mounting on a road-vehicle; the goal of 20 x 10 x 5 cm as the maximum antenna dimension was set as a benchmark for the designs.

State-of-the-art

A number of UWB antenna families have been explored in the past few years for GPR applications. Frequency-independent antennas include equi-angular spiral, conical spiral, log periodic antenna, and other self-complementary structures [5], but their poor time domain responses have restricted their application to GPR [6]. Horn antennas are commonly used in single channel GPR systems, because they exhibit good performance
at broadband frequency ranges [7], but the dimensions are too large for the under-vehicle mounting or deployment in a dense array.

To achieve unidirectional radiation with high gain, the backed cavity was introduced in the wideband bowtie antenna design. Wu et al. [8] reported a resistor-loaded half-ellipse antenna with different backed cavity heights. Also, Qu et al. [9,10] investigated a triangular bowtie and a folded triangular bowtie antenna, both cavity-backed and fed by a parallel stripline via a transition from a microstrip line. Such designs were too fragile to lend themselves to rugged applications in road-speed GPR systems.

**General Design Considerations**

In order to achieve an antenna design that meets the stringent goals delineated in the GPR system specifications and to realize the performance required for air-coupled operation, we follow the design procedure illustrated in Figure 6-2. The iterative design procedure allows for the use of a number of commercially-available tools for the development of antenna in concert with the front end of the GPR system. This simultaneous development allows for a reduced number of iterations required to obtain the final desired performance specifications.
6.2 Antenna Development

Based on the requirements, constraints, and specifications set previously, two sets of UWB antennas were designed; the first set includes both a cavity-backed rounded bowtie antenna and a Vivaldi antenna fed by microstrip and pins (Vivaldi-1) which covers both frequency bands of interest (below 0.96 GHz and above 1.61 GHz). The second set consists of a cavity-backed slotted bowtie antenna and another Vivaldi antenna fed only by microstrip (Vivaldi-2) and this set is particularly optimized for >2 GHz operation. Pictures of each of the antenna prototypes are shown in Table 6-1.

During the antenna design, parameters and data extracted from manufacturer data sheets were introduced and analyzed for each antenna model. Low-cost FR4 microwave epoxy with a relative dielectric constant εr=4.4 was used as substrate for all proposed antennas. Various design configurations were synthesized in the Ansoft HFSS full wave simulator.
to get an optimization design, including the antenna’s profile, the size and thickness of the substrate, as well as the feeding for each antenna. Frequency domain simulation techniques were applied for impedance optimization, gain evaluation, and antenna radiation pattern determination, while time domain simulations were performed for the time-domain response. Finite Difference Time Domain (FDTD) commercial software was used to examine the antenna dispersion, a parameter which is essential to be minimized for good resolution in a GPR system.

After the design of the antennas, manufacturing CAD/Gerber files are generated for circuit board fabrication. The PCB manufacturer is provided with the Gerber files along with a specification for the material to be used in the fabrication. Further, a list of parts is generated; which includes such additional components as cables, absorber, and connectors. The PCBs are built with standard PCB processes and assembled in the lab.

Metal boxes acting as backing cavities were used for the rounded/slotted bowtie antennas in order to achieve unidirectional radiation. However, the cavity also introduces undesirable ringing effects (UWB antenna dispersion) [11]. Therefore, absorbers were attached to the inner surface of the cavity [12]. For rounded bowtie antenna, the magnetically loaded rubber absorber from Cumming Corporation was used. For the slotted bowtie antenna, low-cost AN-75 carbon loaded foam absorber was applied. Time domain measurements verified that the attached ferrite-rubber absorber can effectively reduce the dispersion while maintaining the signal amplitude compared to the cavity-backed bowtie antenna without absorber. Owing to the improved design using absorber
material and backed cavity, it was found that the slotted bowtie antenna retains higher gain (around 6 dB at maximum) than the rounded bowtie antenna, while the former also maintains much smaller dimensions.

6.3 Antenna Characterization

We measured the antenna parameters using a network analyzer for the S11 parameters within an anechoic chamber. The equipment used for testing exceeds the bandwidth specifications of the antennas that were tested and are shown in Table 6-1. External equipment parameters were de-embedded post-measurement in order to extract only the antenna parameters.

In Table 6-1 the four antennas developed are illustrated together for comparison, while the fifth antenna, a commercial, high-cost Vivaldi antenna (Imego brand), is used as a reference.

Vivaldi-1 and Vivaldi-2 have a similar profile, but Table 6-1 shows that the Vivaldi-2 provides a higher gain (about 9 dB at maximum) than Vivaldi-1, and higher resonance frequency (at 3.2 GHz), which better fits with the FCC regulations illustrated in Figure 6-1. All antennas exhibit broadband impedance matches when tested in an anechoic chamber, and displayed stable radiation pattern over the desired wide frequency range. The average 3 dB radiation beam-width is around 60 degrees, which is very suitable for GPR mounted beneath the vehicle.
Table 6-1. Five types of UWB antennas for GPR, with individual experimental S11 and Gain curves presented.

<table>
<thead>
<tr>
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</thead>
<tbody>
<tr>
<td>Rounded Bowtie antenna (cavity-backed, with magnetically loaded rubber absorber) 20 cm x 14 cm x 3.9 cm</td>
<td><img src="image1.png" alt="Image" /></td>
<td><img src="image2.png" alt="Image" /></td>
<td><img src="image3.png" alt="Image" /></td>
</tr>
<tr>
<td>Bowtie Slot antenna (cavity-backed, with carbon-adulterated foam absorber) 10.7 cm x 7 cm x 5 cm</td>
<td><img src="image4.png" alt="Image" /></td>
<td><img src="image5.png" alt="Image" /></td>
<td><img src="image6.png" alt="Image" /></td>
</tr>
<tr>
<td>Vivaldi-1 antenna 15 cm x 14 cm (length)</td>
<td><img src="image7.png" alt="Image" /></td>
<td><img src="image8.png" alt="Image" /></td>
<td><img src="image9.png" alt="Image" /></td>
</tr>
<tr>
<td>Vivaldi-2 antenna 10 cm x 18 cm (length)</td>
<td><img src="image10.png" alt="Image" /></td>
<td><img src="image11.png" alt="Image" /></td>
<td><img src="image12.png" alt="Image" /></td>
</tr>
</tbody>
</table>
When considering the antenna tradeoffs, the Vivaldi-2 antenna provide higher gain and narrower beam-width performance while the slotted bow-tie antenna benefits from the lowest physical footprint. These requirements are very important when considering the air-coupled, space-confined (on-vehicle) applications.

6.4 Introduction to Ground Penetrating Radar Measurement
When testing the antenna with GPR, usually the data collected in our experiment have the form:[2]

\[ f(x, z) = A(x_i, y_j, z_k) \]  \hspace{1cm} (6-1),

with i = 1 to M, k = 1 to N, and j = constant. The GPR is moved along the x-axis direction, and the reflection signals in the z-axis direction are gathered. Such data collection form is called B-scan, and the schematic of B-scan is shown below in Figure 6-3. When
measuring one trace at each scan, this type of data collection form is called A-scan. In reality, when one trace is selected in the Figure 6-3, it can be considered as one A-scan.

![Figure 6-3. B-scan schematic [2]](image)

One example is given in Figure 6-4. to illustrate the B-scan measurement. The antennas along with the GPR board are moved in the x-axis direction, and the z-axis is supposed to be normal to the ground/floor plane.

![move in x-axis direction](image)

![Figure 6-4. B-scan with GPR and UWB rounded bowtie antennas](image)
When B-scanning the pavement in the experiment, hyperbola curve and flat continuous lines are the most common patterns observed in images. These images are plotted directly from the B-scan data. The flat continuous lines arise from the reflection of the air/ground interface, or the interface between each pavement layer.

The origin of hyperbola curve is different. The hyperbola usually comes from reflections of rebar or pipeline buried underneath the road, as shown in Figure 6-5. Usually the reflection cannot be as continuous and homogenous as the reflection from air/ground or pavement layer interface.

![Figure 6-5. the detection of rebar by GPR](image)

Assume a target is buried under the dielectric material surface by depth d. In the homogenous and isotropic materials, the electromagnetic wave propagation speed is

\[ v_r = \frac{c}{\sqrt{\varepsilon_r}} \]  \hspace{1cm} (6-2)

The depth of the target can be calculated as

\[ d = v_r t / 2 \]  \hspace{1cm} (6-3)
where \( \varepsilon_r \) is the dielectric constant and \( t \) is the transit time to and from the target.

As the GPR moves along x-axis in the B-scan in Figure 6-6, the pulses reflected from the rebar will be recorded. When plotted in the x-z plane, a hyperbola envelope of the rebar reflection can be found, as demonstrated in Figure 6-6.

Figure 6-6. Hyperbolic spreading function, which arises from the signal envelop of rebar-reflected pulses [2]

The depth of the rebar target can be written as

\[
    d_0 = \frac{v_r t_0}{2},
\]

(6-4)

And other depths shown in the Figure can also be derived as

\[
    d_n = \sqrt{\frac{x_{n-1}^2 t_n^2 - x_n^2 t_{n-1}^2}{t_{n-1}^2 - t_n^2}},
\]

(6-5)
Detailed discussion on GPR B-scan measurement will be given in the next sections.

### 6.5 Antenna Testing With GPR System

**Experimental Setup**

Tests were designed with the intent to compare the antenna designs (Table 6-1) in a practical setting and at the GPR frequencies of interest. To achieve the goals, realistic testing dimensions were established. The test bed was approximately 1.2 m (4 ft) by 0.6 m (2 ft) with nearly 20 cm (8 in) of sand. The antennas were placed approximately 30 cm (12in) above the sand surface, and the height was maintained throughout the experiment. This test setup attempted to approximate the operation of an air coupled GPR system attached under a vehicle (Figure 6-7).

![Sand Box test setup](image)

**Figure 6-7. Sand Box test setup**

The antennas were connected to an experimental prototype of a GPR system [13,14] with shielded RF cables for testing. The GPR electronics generate a pulse with a center
frequency around 2.5 GHz and a bandwidth of approximately 3 GHz, as shown in Figure 4(b). The signals are attenuated by 20 dB in a loop back experiment, due to the fact that amplitude of direct measurement without antennas would get saturated. From Figure 6-8(a), it is evident that the time-domain source pulse is about 1.2 ns wide. During the testing, the electronics allow the user to increase the gain and change the time between sample points. The settings of the electronics were kept constant throughout the experiments for the ability of comparing data sets.

Figure 6-8. Pulse from a loop back test with a 20dB attenuator. (a) Pulse in time domain; (b) Pulse in frequency domain. (Amplitude(t) is in arbitrary units.)

**Experimental Results**

Since each step was a static experiment, we averaged 100 traces at the location in order to improve the signal to noise ratio for each trace. Then the stacked traces for each location were compiled into a single B scan. The resulting B scans, one for each antenna from Table 6-1, are presented in Figure 6-9.
Figure 6-9. B scans one for each of the tested antenna designs (Table 6-1). (a) Rounded Bowtie, (b) Bowtie Slot, (c) Vivaldi-1, (d) Vivaldi-2, (e) Pacman Antenna (miniaturized Rounded Bowtie antenna), and (f) commercial IMEGO Vivaldi covering approximately 3ft of the test bed, with the x-axis showing the number of traces. The illustration on the left shows the experimental setup.

In Figure 6-9 the B-scans show 3 feet of antenna travel. The x-axis is the number of scans recorded. Data is collected and displayed at the rate of one scan per inch for all antennas. The rebar is visible in the B-scans as a hyperbola between two bands representing the top and bottom of the sand layers. For data processing, a clear well defined hyperbola is desirable. The gain and sampling settings were kept constant for all the antenna tests.

**Trace Analysis**

Further study on those traces from B scans can shed some light on the antenna performance and evaluation on the GPR system. The time domain trace, with antennas situated above the rebar, was extracted and shown in the first column of Table 6-2. It is expected to find some amplitude peaks in the curves, which can be ascribed to the
reflections at those interfaces illustrated in Figure 6-9. Also, the scattering effect of the rebar can be detected from these curves. The peaks labeled with (a), (b), (c) and (d) in the first column of Table 6-2 correspond to the direct coupling of the transmitter-receiver antennas, the reflections of sand surface, the scattering of the rebar, and the reflections of the metal plate on the bottom of the sandbox, respectively.
Table 6-2. Traces extracted from B scans (Figure 6-9) for each antenna. In this table, Amplitude(t) and Amplitude(f) are in arbitrary units. The labels correspond to the direct wave (a), to the sand reflection (b), to the reflection from the top of the buried pipe (c), and the metal plate below the sandbox (d).

<table>
<thead>
<tr>
<th>Antenna Type</th>
<th>Time-domain Trace with antennas sitting on the shelf right above the rebar</th>
<th>Frequency-domain Trace for each antenna shown in frequency domain</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rounded Bowtie antenna</td>
<td><img src="image1" alt="Graph" /></td>
<td><img src="image2" alt="Graph" /></td>
</tr>
<tr>
<td>Bowtie Slot antenna</td>
<td><img src="image3" alt="Graph" /></td>
<td><img src="image4" alt="Graph" /></td>
</tr>
<tr>
<td>Vivaldi-1 antenna</td>
<td><img src="image5" alt="Graph" /></td>
<td><img src="image6" alt="Graph" /></td>
</tr>
<tr>
<td>Vivaldi-2 antenna</td>
<td><img src="image7" alt="Graph" /></td>
<td><img src="image8" alt="Graph" /></td>
</tr>
<tr>
<td>Pacman antenna</td>
<td><img src="image9" alt="Graph" /></td>
<td><img src="image10" alt="Graph" /></td>
</tr>
</tbody>
</table>
6.6 Antenna Summary

Our goal is to reduce the size of the antenna as well as enhance the radiation performance of the antennas. The low cost of fabrication is also pursued through the use of more streamlined designs and comprehensive validation before the antenna manufacture. After numerous simulations and experiments, finally three types of antennas were identified for further study and development, as summarized in Table 6-1. The three antennas are Pacman (miniaturized rounded bowtie antenna, 10 x 7.2 x 5 cm), Bowtie Slot antenna (10.7 x 7 x 5 cm), and Vivaldi -2 (10 x 18 cm).

The radiation gain, radiation pattern, impedance match and return loss of the antennas have been tested in the anechoic chamber at Northeastern University (NEU). The characterization verifies that the expected design can match the experimental result well.
Table 6-3. New Versions of Antennas designed and manufactured

<table>
<thead>
<tr>
<th>Feature</th>
<th>Pacman (miniaturized Rounded Bowtie)</th>
<th>Bowtie Slot</th>
<th>Vivaldi -2</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gain (dBi)</td>
<td>3-4</td>
<td>5-6</td>
<td>8-9</td>
</tr>
<tr>
<td>Bandwidth (GHz)</td>
<td>&gt;2</td>
<td>&gt;2</td>
<td>2.5</td>
</tr>
<tr>
<td>Beam width (degree)</td>
<td>~50</td>
<td>~60</td>
<td>~50</td>
</tr>
<tr>
<td>Low frequency cut (GHz)</td>
<td>1.0</td>
<td>1.1</td>
<td>1.6</td>
</tr>
<tr>
<td>size</td>
<td>10 cm x 7.2 cm</td>
<td>10.7 cm x 7 cm</td>
<td>10 cm x 18 cm</td>
</tr>
<tr>
<td>ESS GPR v3 Test</td>
<td>good</td>
<td>good</td>
<td>fair</td>
</tr>
</tbody>
</table>

The three types of ultra-wideband antennas have also been assembled and packaged for Ground penetrating radar (GPR) measurement. They have been integrated and tested with the ESS radar v3 for road-characterization, as shown in Figure 6-10. Sandbox and Concrete slab test have been conducted to validate and demonstrate the performance of each antenna. Currently, the electric bowtie antenna shows the best resolution for the rebar detection in the subsurface sensing. The experimental results will be displayed in section 2.
6.7 Further Antenna testing with the ESS Radar V3 System

6.7.1 Sandbox testing
Figure 6-11 illustrates the experimental setup in the lab for sandbox testing. One rebar is buried 6 inches below the surface of the sandbox. The cables seen in Figure 6-11 are connected to antennas on one end, and connected to the ESS radar on the other side.
Figure 6-11. Experimental setup in the lab for sandbox testing

Figure 6-12 shows the B scans from sandbox testing for six antennas. Compared to the commercial Imego Vivaldi design, the designed antennas do demonstrate good performance and possess smaller dimensions. The measured performance of the antennas along with the prototype GPR system illustrates the feasibility of using them for air-coupled, vehicle-based GPR applications.
Figure 6-12. B Scans from sand box testing. a) Commercial IMEGO Vivaldi, b) Pacman, c) Bowtie Slot, d) Vivaldi-2. Antennas are approximately 1ft above the sandbox surface.

6.7.2 Concrete Slab Testing
Figure 6-13 illustrates the experimental setup in the lab for Concrete Slab testing on the NEU Burlington campus. The radar system is 1 feet above the slab surface. The distance between the slab bottom and the ground surface is 6 inches.
Figure 6-13. Experimental setup in the lab for Concrete Slab testing

Air coupled Mode

Figure 6-14. B Scans from Concrete slab testing (air-coupled). a) Commercial IMEGO Vivaldi, b) Pacman, c) Bowtie Slot, d) Vivaldi-2.

Figure 6-14 shows the B scan of air-coupled GPR above the concrete slab. It can be found that the Electrical Bowtie antenna performs the best resolution. On the other hand,
the higher gain antenna, commercial Imego and Vivaldi -2 exhibit better Penetrating capability.

**Ground Coupled Mode**

![Figure 6-15. B Scans from Concrete slab testing (ground-coupled). a) Commercial IMEGO Vivaldi, b) Pacman, c) Bowtie Slot, d) Vivaldi-2.](image)

The ground-coupled Concrete slab testing shown in Figure 6-15 clearly shows the very good resolution of Electrical Bowtie designed in NEU. The performance of Magnetic Bowtie is also good, compared to the performance of Imego and Vivaldi-2 antennas.
6.8 Conclusions

In this chapter, several types of antennas were designed, manufactured, and evaluated for use in an air-coupled GPR system operating at road speeds. A methodical design procedure was employed to maintain the broadband impedance of the UWB antennas as well as the tradeoff between good performance and reduced size. Careful parameter analysis during the numerical simulation led to unidirectional, high-gain antennas with stable radiation patterns. The measured performance of the antennas along with their integration with a prototype GPR system illustrate the feasibility of using them for air-coupled, vehicle-based GPR applications.

From the testing with ESS GPR v3 system, it can be concluded that the Pacman (miniaturized rounded Bowtie) antenna demonstrates the best resolution. The Bowtie slot antenna is also a good candidate considering its compact size and relatively high gain. Also, its planar structure with CPW feed makes it convenient for radar system integration and antenna array development.

Regarding the cost and performance of the antennas designed for VOTERS project, the Pacman (miniaturized rounded Bowtie) antenna and the Bowtie slot antenna will be used to develop antenna array, as discussed in chapter 7.
6.9 References


Chapter 7 Antenna Array Development for Impulsive Ground Penetrating Radar

The focus of this chapter is the design of an antenna array with 2 or more antennas to achieve higher gain [1, 2] while maintaining the good resolution and compact size. [3, 4]

As summarized before and shown in the Table 1 of Chapter 6, the Pacman and Bowtie slot antennas demonstrate good resolution in both ground-coupled mode and air-coupled mode. Therefore, these two types of antenna are proposed for antenna array development.

The following sections will mainly discuss the antenna array modeling and simulation for both Pacman and Bowtie slot antenna.

7.1 Bowtie slot Antenna Array and Pacman Antenna Array

7.1.1 Bowtie slot antenna array
The following Figure 7-1 shows the proposed Slotted Bowtie antenna array. The size of this antenna array is 82mm x 260mm x 50mm. This array is composed of two bowtie slot antennas, and fed by a 3-port power divider.

![Figure 7-1. The proposed Bowtie slot antenna array](image)
The frequency range of $S11 < -10\text{dB}$ is $1.6 – 3.5 \text{GHz}$, as shown in Figure 7-2.

![Figure 7-2. S11 of Bowtie slot antenna array](image)

The gain of this array can be improved more than 3dB, reaching as high as 9dB, as shown in Figure 7-3.

![Figure 7-3. The gain of the proposed Bowtie slot antenna array](image)

The radiation pattern of antenna array at 2 GHz is shown below in Figure 7-4. It can be found that the radiation slightly deviates from the normal of antenna plane. Parameter analysis shows that the width of upper ground plane and the length of feed line and lower
ground plane have great influence on the antenna radiation pattern. When the length of the upper and lower ground plane is relatively equal, then the major radiation is in the normal direction. Ideally, if the upper and lower ground planes are symmetric, the major radiation direction would be purely forward and backward.

**Figure 7-4. The radiation pattern of Bowtie slot antenna array at 2 GHz**

The antenna dispersion performance is performed using model in Figure 7-5, with a pair of Bowtie slot antenna array face-to-face 30 cm away. The excitation signal of third order Gaussian pulse is shown Figure 7-6.

**Figure 7-5. Model for antenna dispersion characterization**
The excitation signal of third order Gaussian pulse can be found in Figure 7-6. The time domain response of the Bowtie slot antenna array is shown in Figure 7-7. It can be found that the array displays heavier ringing effect, in comparison with the case when only single Bowtie slot is used in the GPR.

Figure 7-7. The time domain response of the Bowtie slot antenna array.
Also, as the improper reflections occur in the 3-port power divider, it is difficult to alleviate the ringing in such antenna array configuration. Thus, the following discussion will focus on the Pacman antenna development.

7.1.2 Pacman antenna array
The array of two Pacman antenna is also proposed, as shown in Figure 7-8. The size of this antenna array is 140 mm x 93mm x 50mm. The bi-directional radiation pattern when no shielding box used is demonstrated on the right side of Figure 7-9, which is quite suitable for Ground Penetrating Radar application. In this case the two elements (Pacmans) are fed with equal in-phase currents [2].

Figure 7-8. The proposed Pacman antenna array
Figure 7-9. Radiation pattern of the proposed Pacman antenna array

The S11 (return loss) of the 2-Pacman array is displayed in Figure 7-10, with S11 < -10dB in the frequency range of 1.3 – 3.5 GHz. While the Pacman array maintains the broadband width of the signal Pacman antenna, the gain of this array can be improved more than 4dB, as shown in Figure 7-6.

Figure 7-10. The S11 of the proposed Pacman antenna array
Figure 7-11. The gain of the proposed Pacman antenna array

The time domain response of the Pacman antenna array is shown in Figure 7-12, in comparison with the case when only single Pacman is used in the GPR. It can be found that the Pacman antenna array could demonstrate good time domain response. It just becomes slightly worse, compared to the single Pacman case. Note that in this simulation, the model is constructed with two transmitter antennas, and one receiver antenna, as shown in Figure 7-13.
Figure 7-12. The time domain response of the Pacman antenna array.

Figure 7-13. The model used for time domain response simulation of Pacman antenna array

7.2 Design of Pacman Antenna Array Prototype
To verify the Pacman antenna array performance, an array prototype was built. When constructing the antenna array, the insertion loss should be as low as possible, and phase synchronization is critical to get small antenna dispersion over a broad frequency
bandwidth. Thus the Mini Circuit ZB2PD-63+ 2 way-0\(^0\) power splitter is used in the experiment, which has low insertion loss of 0.5 dB over 0.6 – 6 GHz.

During the time domain response measurement, three types of antenna configurations are applied.

1) one-one case, in which one Pacman used as transmitter antenna, and one Pacman used as receiver antenna.

2) one-two case, in which one Pacman used as transmitter antenna, and two Pacman used as receiver antenna.

3) two-two case, in which two Pacman used as transmitter antenna, and two Pacman used as receiver antenna.

The two-two case is shown in Figure 7-14 below.

Figure 7-14. The two-two Pacman antenna array setup for time domain response measurement
The time domain response of Pacman antenna array is plotted in the Figure 7-15, for the three types of antenna array configurations: one-one, one-two, and two-two. From the experimental results of Figure 7-15, the one-two antenna array dispersion performance can match the simulation well. Meanwhile, the one-two configuration demonstrates low ringing effect, and thus is probably suitable for antenna array development.

![Figure 7-15](image)

**Figure 7-15.** The time domain response of Pacman antenna array, for the three types of antenna array configurations: one-one, one-two, and two-two.

### 7.3 Pacman Antenna Array Test with Ground Penetrating Radar

Further tests were conducted to explore the potential of Pacman antenna array for GPR application, since it exhibits higher gain while maintaining antenna dispersion.
7.3.1 Sandbox Testing in Air-coupled Mode

Figure 7-16 illustrates the experimental setup in the lab for sandbox testing. One rebar is buried 6 inches below the surface of the sandbox. The 1- and 2- ports of the power divider are connected to the two Pacman in the antenna array. The port 3 of the power divider is joined to the receiver side of the ESS GPR v3 circuit board.

![Experimental setup in the lab for sandbox testing](image)

**Figure 7-16. Experimental setup in the lab for sandbox testing**

Figure 7-17 shows the B scans from sandbox testing for Pacman array of one-one and one-two configurations. It is interesting to observe that the one-two configuration collects much stronger reflected signals. This is consistent with the expectation, since the Pacman antenna array demonstrates higher gain than a single Pacman.
Figure 7-17. Sandbox Testing of Pacman Array in Air-Coupled Mode
(a) one - one, (b) one - two

7.3.2 Concrete Slab Testing in Air-coupled Mode
Figure 7-18 illustrates the experimental setup in the lab for Concrete Slab testing in Burlington Campus, NEU. The radar system is one feet above the slab surface. The distance between the slab bottom and the ground surface is 6 inches.
Figure 7-18. Experimental setup in the lab for Concrete Slab testing

Figure 7-19. Concrete Slab Testing of Pacman Array in Air-Coupled Mode
(a) one - one, (b) one - two

Figure 7-19 shows the B scan of air-coupled GPR above the concrete slab for the one-one and one-two antenna array configurations. Due to the higher gain in the two- Pacman
antenna array, the one-two configuration collects stronger reflected signals, and thus could observe more details from the inside of the subsurface. From these experimental results, it is verified that one-two antenna array configuration does enhance the penetrating capability of the GPR system.

7.3.3 Pavement layer identification in air-coupled mode
The antenna array is also used to identify the pavement layers in air-coupled mode, as shown in Figure 7-20. The antenna array of the radar system is one feet above the ground surface. The top layer of the pavement is made of asphalt, and its thickness is around 2 inches.

![Experimental setup for pavement layers identification in air-coupled mode](image)

Figure 7-20. Experimental setup for pavement layers identification in air-coupled mode

Figure 7-21 shows the B scan of air-coupled GPR above the pavement for the (a) one-one and (b) one-two antenna array configurations. It can be found that the one-two configuration demonstrates clearer and stronger signals due to the higher gain as well as relatively low antenna dispersion. The B-scan images of one-two display delayed signal detection because of the longer transmission line is used in the experimental setup.
Figure 7-21. Pavement layers identification in air-coupled mode for Pacman antenna array configuration of (a) one-one, (b) one-two

7.4 The Ground Reflection Cancellation in Antenna Array
In the one-two antenna array configuration discussed above, the ‘two’ Pacmans is arranged always as Figure 7-22 (a). In this way, the benefit of enhanced gain is obtained through the array, and the corresponding B-scan with the metal bar on the ground surface demonstrates the results of Figure 7-24(a).
Figure 7-22. Parallel vs anti-parallel antenna arrangement in one antenna array (a) the ‘two’ Pacmans arranged parallel in the one-two antenna array configuration; (b) the ‘two’ Pacmans arranged antiparallel in the one-two antenna array configuration.

Figure 7-23 shows the practical example of transmitter and receiver Pacman antennas used in the measurements. Figure 7-23(a) describes the same antenna configurations as Figure 7-22(a), and Figure 7-23(b) represents the same antenna configurations as Figure 7-22(b).
Figure 7-23. Transmitter and receiver Pacman antennas used in the measurements. (a) Two Pacman elements in the antenna array (RX) are in parallel arrangement; (b) Two Pacman elements in the antenna array (RX) are in anti-parallel arrangement.

When the two Pacmans are arranged antiparallel in the one-two antenna array configuration in Figure 7-23 (b), the B-scan in the testing with the metal bar on the ground surface shows the results of Figure 7-24 (b). It is interesting to found that the ground reflection has been cancelled effectively, and the direct coupling as shown in Figure 7-24(a) has been reduce significantly.
Figure 7-24. B-scan when the metal bar on the ground surface for the one-two antenna array configuration (a) the two Pacman elements in the array arranged parallel; (b) the two Pacman elements in the array arranged antiparallel

The reason behind such phenomenon originates from the effective current distribution for radiation in the Pacman array. When the two Pacmans are in parallel direction, the surface currents on the two Pacmans are in-phase and contribute to the radiation positively, and thus the gain will be enhanced. On the other hand, when the two Pacmans arrange anti-parallel, the currents on the two antennas are in opposite phase and cancel the radiations of each other in the normal direction of the array [2]. As a result, very slight or little signal would be detected. The current distributions of these two cases are shown in the following Figure 7-25.
Meanwhile, there is always phase difference when the electromagnetic wave is not normal incidence to the target. As a result, the hyperbola outline could always be plotted when the anti-parallel Pacman array is used in the one-two antenna configuration, while the signal strength of the vertex point is still quite faint. A schematic has been shown in Figure 7-26. When the distance between the rebar and the ground surface becomes zero, it represents the testing setup of Figure 7-24. Such method could provide a useful technique for the ground reflection cancelation.
Figure 7-26. The hyperbola outline could always be plotted when the anti-parallel Pacman array is used in the one-two antenna configuration, while the signal strength of the vertex point is still quite faint.

7.5 GPR Testing for Antenna Array with Anti-parallel Pacman Antenna Element Arrangement

In this section, the experimental setup of the GPR testing with sandbox and concrete slab is the same as the setup in section 7.3.1 and 7.3.2. The only difference is that the antenna element in the array has been positioned anti-parallel. For comparison, the B-scan results of both parallel and anti-parallel cases in the one-two antenna array configurations are demonstrated.

7.5.1 Sandbox Testing in Air-coupled Mode

Figure 7-27 shows the B-scan images of sandbox testing. The antenna array configuration is of one-two case. It can be found that the direct coupling and the reflection from the sand surface have been reduced effectively. What’s more, the reflection from the floor...
ground surface, which is represented by the horizontal line around sample interval number of 280, has been depressed significantly. The rebar buried in the sand, however, can still be detected clearly through the hyperbola shape in Figure 7-27 (b).

![Figure 7-27. B-scan of sandbox testing for one-two antenna array configuration](image)

(a) the two Pacman elements in the array arranged parallel; (b) the two Pacman elements in the array arranged antiparallel

### 7.5.2 Concrete Slab Testing in Air-coupled Mode

Figure 7-28 shows the B-scan images of concrete slab testing. The antenna array configuration is also of one-two case. Due to the vibration of the cart when crossing the concrete slab edges, there are some inconstancy and fluctuations in the images.
The reflections from the concrete slab surface have also been reduced effectively.

Meanwhile, the rebar buried in the concrete can be identified distinctly through the hyperbola shape in Figure 7-28 (b).

![Figure 7-28. B-scan of concrete slab testing for one-two antenna array configuration (a) the two Pacman elements in the array arranged parallel; (b) the two Pacman elements in the array arranged antiparallel](image)

7.6 Conclusions
The compact antenna arrays enhance the GPR penetrating capability via higher gain while maintaining a low level of ringing effect. When two Pacman antennas in the array are positioned in parallel, the gain of the array can be augmented up to 4 dB when compared to the original single Pacman application.

Anti-parallel Pacman arrangement in the antenna array has also been proposed and tested, which could eliminate the ground reflection and direct coupling effectively due to the
1800 phase difference between the two Pacmans in one antenna array. This technology can be very beneficial for rebar detection applications.

Field test has been conducted in different scenarios, including rebar detection buried in sand box, as well as rebar identification in reinforced concrete slab. Both the parallel and anti-parallel antenna arrangement in the antenna array has been tested with ESS GPR system on sandbox and concrete slab to verify the enhanced antenna array performance. Multi-layer structures in the reinforced concrete slab have been identified with great clarity. It is also found that anti-parallel connection is able to reduce ground reflection significantly by around 70%, while the rebar can still be observed clearly in the experiment, which provides a promising way to map and image the radar as well as other subsurface objects.

7.7 References


Chapter 8 Summary

For this dissertation I completed a full project cycle to develop Ultra-Wideband (UWB) antennas for impulsive ground penetrating radar (GPR) system at Northeastern University. As an important component of the versatile roaming sensors system, the GPR plays an essential role in subsurface pavement and bridge deck condition assessment.

With extensive survey and investigation on the UWB antenna specifications and requirements for GPR system, three highly compact UWB antennas were designed for vehicle-mounted impulse subsurface radar. The UWB antennas included electric small antenna (rounded Bowtie antenna), magnetic small antenna (Bowtie slot antenna), and planar horn antenna (Vivaldi antenna).

The UWB antennas for VOTERS project at Northeastern University were pursued intensively with low-profile and compact size. Numerical simulations were conducted exhaustively for antenna performance optimization. Higher gain and broader bandwidth were strived for using different optimization techniques, such as trimming antenna geometry, and designing efficient feeding structure.

The UWB antennas were manufactured, assembled and characterized in anechoic chamber. A series of measurements were also performed for the antennas collaborating together with the VOTERS GPR system, including sand box testing and concrete slab testing. Real-world field testing was also conducted to indentify the pavement layers.
During the development of GPR antennas, the antenna dispersion problem was studied, clarified and summarized with theoretical analysis, simulation models and experimental characterization. Simulation Modeling was developed to aid the UWB antenna design for good overall GPR system performance, as it always provides an effective way to reduce the ringing effect.

In the end of the dissertation, the UWB antenna array was explored to achieve stronger penetrating capability while maintaining good resolution for the GPR system. Simulation and experiment verified successfully that it is feasible to utilize antenna array to go deeper into the subsurface and attain stronger reflection signals clearly. What’s more, experiments, theoretical and simulation analysis proved that Anti-parallel Pacman arrangement in the array could eliminate the ground reflection and direct coupling effectively, which could be very beneficial for rebar detection applications.

In the future work, more effort can be exerted to explore and develop compact antenna array, which exhibits great potential in enhancing the overall GPR system performance.