CHARACTERIZATION OF VARIOUS ANTENNA STRUCTURES USED FOR DEVICE ILLUMINATION

by

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A thesis submitted to the Graduate Faculty of North Carolina State University in partial fulfillment of the requirements for the Degree of Master of Science

Electrical Engineering

Raleigh

May 2003

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A non-intrusive approach for personal surveillance is explored. A potential strategy for identification of electronic devices using electromagnetic illumination with the aim of identifying unique resonances and signatures is presented. The concept is that signatures can be compared to a known signature for a particular device. If discrepancies are detected, then the device may be flagged as a suspect item. Various antenna structures are analyzed to determine the ideal antenna characteristics needed for electromagnetic illumination. A double ridged horn antenna is designed, characterized and compared to a conical spiral right hand circular polarized antenna and a standard gain antenna. Various passive structures and electronic devices are illuminated with electromagnetic energy and responses analyzed.
To my dear wife Christine
BIOGRAPHY

Mark Buff was born in Warner Robins, Georgia, USA into an Air Force Family and has lived in numerous states and countries. He received his Bachelors Degree in Electrical Engineering from North Carolina State University in 1994, where soon after, he was employed by Alcatel Network Systems in Raleigh. There, he worked on fiber optic multiplexing transport systems and digital cross connect switches. He returned to North Carolina State University in the Fall of 2001 to pursue a Master’s Degree in Electrical and Computer Engineering.
ACKNOWLEDGEMENTS

I would like to express my deepest thanks and love to my wife who has sacrificed countless nights and weekends for the success of my education.

I take this opportunity to thank my advisor, Professor Michael Steer for his invaluable guidance and vivacity. I am looking forward to pursuing my doctorate under his direction. In addition, thanks goes to my thesis committee, Dr. Hamid Krim and Dr. Doug Barlage.

Thanks to my bright and talented lab friends whom all have acted as great sources of energy, ideas and camaraderie.

Thanks to my gracious parents for all of your support.
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Chapter 1

Introduction

1.1 Motivation

A new means for personal surveillance is explored. Existing techniques, such as those used in airport security systems and government workplaces, are intrusive and intimidating. While these systems present a certain deterrence, their use in the common public place is impractical. New means for individual surveillance in the public realm are required. While imaging at millimeter wave frequencies can be used to detect weapons, it cannot determine internal details of electronic and electrical devices. Electromagnetic (EM) illumination of electronic devices with the aim of identifying unique resonances and signatures is a scheme with a potential strategy for identification of such devices. The concept is that signatures can be compared to a known signature for a particular device. If discrepancies are detected, then the device can be flagged.

Illuminating a target with electromagnetic energy requires that an antenna with sufficient radiating properties be employed. The scattered response from various targets is considerably smaller than that of the incident wave, thereby requiring that ample power be transmitted. Furthermore, environment limitations such as multipath and noise, requires that high directivity antennas be used to help minimize undesirable reflections.
1.2 Thesis Overview

Chapter 2 discusses various surveillance systems in use today. System technologies are presented and their capabilities and limitations are analyzed. The analysis provides the justification for exploring electromagnetic device illumination for personal surveillance systems.

Chapter 3 discusses the theory of antenna operation and characterization. Important antenna parameters are identified and defined. This discussion emphasizes the antenna’s key role in successful device illumination and provides the groundwork needed for the analysis of the three antennas of interest here: the standard gain horn, the Right Hand Circular Polarized (RHCP) conical spiral, and a Double Ridged Horn (DRH) antenna.

The theory behind electromagnetic reflection and transmission at various interfaces is presented and how it applies to electromagnetic device illumination. Scattering by perfect electrical conductors and lossless dielectrics are analyzed.

Basic network calibration techniques are discussed as it pertains to this application.

Chapter 4 presents the actual results of the various antenna characterizations. The advantages and disadvantages of each antenna is exploited as it applies to electromagnetic illumination.

Chapter 5 presents actual results obtained by illuminating various structures. Microstrip targets are illuminated and the responses are analyzed. Actual measurements are compared to simulator results. Resonant characteristics of various width microstrips are examined. Multiple strip targets are analyzed and compared to individual responses.

Square metallic plates are illuminated and results compared to simulated results. Laptop computers from different vendors are illuminated and their signatures are compared. Illumination of various computer configurations are compared.

Chapter 6 presents an overall conclusion of the various antennas’ performance and their advantages and disadvantages. The Double Ridged Horn antenna’s contributions to successful device illumination is exploited.
The knowledge gained by illuminating various simple geometries and electronic devices are summarized.
Chapter 2

Existing Surveillance Technologies

2.1 Introduction

Existing surveillance systems are widespread in airports, ports of entry, government facilities, correctional facilities and customs. Sophisticated image processing techniques and millimeter wave technologies have taken these systems to a new level of detection capabilities. However, there still remains severe limitations in their applications and our surveillance system infrastructure as a whole.

2.2 X-Ray Imaging Systems

Current 3D imaging X-ray machines are capable of detecting concealed weapons, explosives, narcotics, currency and contraband by having the ability to penetrate articles with X-rays. Traditional 2D X-ray imaging systems produce a shadowgraph, or flat projection of an object under inspection. Individual components are superimposed on each other with no indication of their relative depth. Consequently, the resultant image can be very difficult to interpret. With contemporary X-ray scanning systems, 3D stereoscopic imaging is used providing the operator with a 3D image of objects under inspection. These visuals are in real time and decrease the probability of a missed threat. These systems have the capability to detect wires as small as 0.0048 inches in diameter and provide features such as material marking which automatically
Figure 2.1: Gilardoni Advanced Detection System (ADS) marking a potentially explosive material in orange.

highlights suspect items for extensive discrimination. This is achieved by providing the required material resolution to distinguish between explosives (organic) materials, narcotics and other materials.

In general, airport X-ray imaging systems are dual energy systems with an X-ray source capable of producing 140 to 160 kilovolts peak. X-rays pass through the target item and are received by a detector. Higher energy X-rays are passed through a filter on to a second detector. The two detectors are used in conjunction with one another to sort low and high absorption materials. Items are categorized into three main categories:

- Organic
- Inorganic
- Metal

Most explosive materials are organic. In Figure 2.1, a Gilardoni X-Ray system highlights suspect explosive materials in orange.

The Smiths Heimann X-Ray system is a leading edge X-Ray detection system that incorporates features such as an Advanced Contents Tracking (X-ACT) system. This feature provides an automated recognition system for detecting suspicious items. Materials of different origin are highlighted in one of three colors. The system marks possible explosive items in red, narcotics in green and high density items in blue. It has the ability of separating materials of interest from other items even if the suspect items are obscured or overlapped by others. Figure 2.2 shows images of each colored
Figure 2.2: Smiths Heimann X-ACT marking system framing various materials in different colors.

Figure 2.3: Smiths Heimann HI-SPOT automatic dense area detection analysis of a weapon covered by a steel plate.

frame. The Smiths Heimann system also incorporates a feature called HI-SPOT that automatically detects sections of high absorption. Once a high absorption area is detected (a piece of steel), then a special enhancement is placed on that area and then locally illuminated. The image evaluation of materials which are difficult to penetrate is improved without deteriorating the image information of the other image sections. Figure 2.3 demonstrates this ability to detect a weapon covered by a steel plate.

X-Ray systems are extremely advanced and do provide a significant force in detection of suspect items. The disadvantages of such systems are their extreme size, cost and their inability to be used on humans due to ionizing radiation. Figure 2.4 shows a typical X-Ray system. These systems can weigh over 3000 pounds and can cost hundreds of thousands of dollars. Their capability extends only to luggage and personal affects, therefore substantially limiting their applications.
2.3 Metal Detecting Systems

Generally, airport metal detectors use a pulse induction (PI) technology. A coil of wire on one side of the arch is used as the transmitter and receiver. Powerful bursts of current generates a brief magnetic field. When the pulse ends, the magnetic field reverses polarity and collapses suddenly, resulting in a sharp electrical spike. This spike constitutes the reflected pulse and lasts about 30 microseconds. The pulse frequency ranges from 25 pulses per second to over 1000 and depends on the manufacturer and model. Figure 2.5 shows a typical metal detecting archway system. When a metal object passes through the metal detector, the magnetic field induces an electrical current in the object, which in turn creates a magnetic field. When the magnetic pulse of the detecting system collapses, the magnetic field of the object arrives at the receiver at some later time. The sampling circuit monitors the length of the received pulse and determines if there is a metal object contributing to the received field. Figure 2.6 demonstrates the basic operation of a metal detecting system. Some metal detecting systems employ dual antenna or coil systems which provide multifield detection within the entry archway. These systems provide low false alarm rate circuitry and auto calibration routines.

While providing accurate metal detection capabilities, these systems provide no discrimination against simple innocuous items such as belt buckles, keys, coins, etc. and quite often create substantial bottlenecks in traffic flow.
Figure 2.5: Checkgate metal detector.

Figure 2.6: Metal detector operation.
2.4 Millimeter Wave Imaging

Systems utilizing three-dimensional millimeter wave imaging are emerging and can detect the presence of non metallic threats [13]. Passive systems do not illuminate the target and rely simply on black body radiation reception. A passive millimeter wave image is shown in Figure 2.7.

Active millimeter imaging systems, unlike X-ray systems, are nonionizing and therefore pose no known health hazard at moderate power levels. Active millimeter-wave imaging systems are capable of penetrating common clothing barriers to form an image of a person as well as any concealed weapons. Millimeter-wave systems can be high resolution due to the relatively short wavelength (1-10 mm). Figure 2.8 shows a 350 GHz reconstructed image of a Glock-17 9 mm handgun [13]. Although Figure 2.8 demonstrates a resolution of less than 1 mm, the scan required ten minutes to complete with a large number of samples, thereby making this impractical for a surveillance system. Practical weapon detection systems using millimeter wave technologies are expected to operate at 100 GHz or lower.

Practical weapon detection systems utilizing millimeter wave technology require
high speed scanning on the order of 3 to 10 seconds [13]. Sheen, McMakin and Hall achieve this by using a 27–33 GHz linear sequentially switched array and a high speed linear scanner. The system quickly switches transmitters and receivers over the large aperture to illuminate the target. Phase and magnitude information are captured at various frequencies and mathematically reconstructed by a computer to form a focused image of the target. Original holographic imaging systems operating at a single frequency do not allow targets with significant depth, such as a human body, to be reconstructed in complete focus. Figure 2.9 shows wide-band (27–33 GHz) images of a man carrying concealed handguns. The scans take approximately 1 s each and demonstrate the high image quality compared to that of single frequency systems. The first image in Figure 2.9 shows a handgun in the man’s beltline. The second shows a gun in the left-hand pants pocket. The third image shows a vinyl/paper checkbook in the left-hand pocket. The fifth image shows a leather wallet in the back right hand pocket. The advantage of these systems over X-Ray systems is that humans can be safely targeted. However, the millimeter wave systems are based on visual images that must be scrutinized for legitimate threats.
Figure 2.9: Wide-band images of a man carrying concealed handguns.
2.5 Chemical Trace Detection Systems

Detection of explosives and chemical agents can be detected by ultra-fast gas chromatography. The gas chromatograph separates ions accelerated by an electric field with charged plates [14]. The incoming sample is preconcentrated and injected into two separate detection systems. One system traps and analyzes the low volatility compounds such as nitroglycerine, AN, TNT, PETN and RDX. The second system traps and analyzes high volatile compounds such as EGDN, DMNB, O-MNT and P-MNT. One such system is the Smith Detection IONSCAN Sentinel II shown in Figure 2.10. This system directs a person into a portal where air is used to dislodge particles and vapors trapped on the body and clothing, where they are sampled and analyzed. Sensitivities of such systems extend down to the picogram level. Non portal based systems that require swiping a sample from electronic devices, documents, and personal items tend to create traffic bottlenecks, as do metal detection systems. These systems work well when explosives or chemical agents are located in baggage or clothing where air can transport enough of the sample to be detected, or if residual samples remain on personal items. However, carefully placed compounds in electronic devices may slip right through such detection systems.

While these systems are extremely advanced and capable, there remains a major
loophole in our surveillance and detection system infrastructure. Electronic devices make an appealing transportation medium for explosives and chemical agents. Due to the densely packed components in a relatively small package, X-Ray and millimeter wave imaging systems cannot sufficiently analyze internal components. Current security measures require security officials to verify functionality of these devices. X-Ray systems might flag explosive materials located in a suitcase but generally will fail to detect such materials encased in a metallic enclosure in a spare drive slot in a laptop.

The technology presented in this paper describes a potential strategy for comparing a device and its properties to known signatures for verification and validity of its composition. Initial work requires various antennas to be characterized and compared so a determination can be made as to which radiating structure is best suited for this particular application.
Chapter 3

Detection

3.1 Approach

In attempting to measure and analyze the scattering effects of various metallic structures and electronic devices, the characteristics of the antennas and the device-field interactions must be understood. Measuring such small backscattered power requires high directivity antennas, sufficient transmit power, a high dynamic range receiver and suppression of multipath and interference. A simplified diagram of the EM illumination concept is presented in Figure 3.1. Section 3.2 describes the theory and quantifying characteristics of various types of antenna structures.

3.2 Antenna Parameters

The antenna is a transitional structure between free space and a guiding device. It is used as a transducer that interfaces a circuit and space by transducing an electronic signal to a propagating electromagnetic signal. The radiation produced by any antenna is provided by the acceleration or deceleration of charge and is described by the basic radiation equation.

\[ \frac{dI}{dt} L = Q \frac{dv}{dt} \quad \text{(Am/s)} \]  

(3.1)
where

\( \frac{dI}{dt} \) is the time changing current, \( \text{A/s} \)

\( L \) is the length of the current element, \( \text{m} \)

\( Q \) is the charge, \( \text{C} \)

\( \frac{dv}{dt} \) is the time change of the velocity or acceleration

Consequently, accelerated charges radiate and time varying currents radiate.

The load that an antenna presents to a transmitter (or Thevenin equivalent impedance for a receiver) is

\[ Z_A = (R_L + R_r) + jX_A \]

where the load resistance \( R_L \) is used to represent the conduction and dielectric losses associated with the antenna structure and \( R_r \) is associated with the radiation resistance seen by the structure. The radiation resistance represents the loss due to power carried away in the propagating field. The reactive part \( X_A \) is the imaginary part and generally should be minimized. When the feeding structure is matched to the generator (or receiver) across the usable band and the dielectric and conduction losses are kept to a minimum, then maximum power can be delivered to the antenna, and therefore, space. If the antenna structure is not properly designed, then the structure can loose energy in ohmic losses or store energy rather than radiating it.
3.2.1 Radiation Pattern

An antenna radiation pattern is defined as being a mathematical or graphical representation of the radiation properties of the antenna as a function of space coordinates [10]. Radiation properties consist of power flux density, radiation intensity, field strength and directivity polarization [2]. For our concerns, the two dimensional spatial radiated electric field with respect to the spherical coordinate system will be used.

A directional antenna, unlike the isotropic, radiates energy more effectively in one direction. Omnidirectional antennas radiate uniformly in all directions. Directional antenna performance is captured in its radiation pattern which consists of lobes. These lobes include a main lobe, minor, side and back lobes. Lobes are defined as a portion of the pattern bounded by relatively weak radiation intensity [2]. The main lobe is considered to be the primary beam where directivity or gain is at its highest. Minor lobes include any lobes that are not a side or back lobe and is not part of the main lobe. The side lobes are defined as those that exist in the unintended direction of propagation. Back lobes exist at an angle of approximately 180 degrees from the main lobe. Figure 3.2 shows the various lobes of an antenna. In general, all lobes should be minimized except for the main beam. Lobes allow energy to be inadvertently coupled into or radiated from the antenna and can result in unintended behavior.

3.2.2 Field Regions

The area around an antenna is divided into three regions:

- the reactive near field

- the radiating near field (Fresnel zone)

- the far field (Fraunhofer zone)
The regions define the types of various fields that exist in different locations of the source of radiation (or reception) and are shown in Figure 3.3. The reactive field is the region defined as

\[
\text{that portion of the near field region immediately surrounding the antenna wherein the reactive field predominates [10].}
\]

For most cases, this region can be defined as follows:

\[
R < 0.62\sqrt{D^3/\lambda}
\]  

(3.2)

where \( D \) is the largest dimension of the antenna and \( R \) is the distance from the antenna. In this region, the fields are not completely solenoidal in character and can be complex in nature.

The radiating near field region is defined as

“that region of the field of an antenna between the reactive near field region and the far field region wherein radiation fields predominate and wherein the angular field distribution is dependent upon the distance from the antenna. If the antenna has a maximum dimension that is not large compared to the wavelength, this region may not exist. For an antenna
focused at infinity, the radiating near field region is sometimes referred to as the Fresnel region on the basis of analogy to optical terminology. If the antenna has a maximum overall dimension which is very small compared to the wavelength, the Fresnel region may not exist [10]."

The boundary for this region can be expressed as follows

\[ 0.62 \sqrt{D^3/\lambda} \leq R < 2D^2/\lambda \] (3.3)

This criterion is based on a maximum phase error of \(\pi/8\).

The far field region is defined as that region of the field of an antenna where the angular field distribution is essentially independent of the distance from the antenna. If the antenna has a maximum overall dimension \(D\), the far field region is commonly taken to exist at distances greater than \(2D^2/\lambda\) from the antenna. The far field patterns of certain antennas, such as multibeam reflector antennas, are sensitive to variations in phase over their apertures. For these antennas, \(2D^2/\lambda\) may be inadequate. For an antenna focused at infinity, the far field region is sometimes referred to as the Fraunhofer region on the basis of analogy to optical terminology [10].

In this region, the fields can be considered transverse and uniform.
3.2.3 Directivity, Gain and Antenna Efficiency

The directivity of an antenna is defined as the ratio of the radiation intensity in a given direction from the antenna to the radiation intensity averaged over all directions, where the average radiation intensity is equal to the total power radiated by the antenna divided by $4\pi$. The directivity is defined as

$$D_{max} = \frac{4\pi \times U_{max}}{P_{rad}}$$  \hspace{1cm} (3.4)

where:
- $D_{max}$ is the maximum directivity (dimensionless)
- $U_{max}$ is the maximum radiation intensity (W/unit solid angle)
- $P_{rad}$ = is the total radiated power (W)

For the isotropic source, the directivity is unity.

The gain of an antenna can be defined in various ways depending on the number and types of losses that are taken into account. The gain of an antenna will always be less than the directivity because of these losses. Such losses can include ohmic losses in the antenna, losses due to a radome, power that is lost due to heating of the antenna and feed mismatches. In most formal definitions of gain, mismatch losses are not to be incorporated in the gain calculation, but for these purposes, mismatch losses will be included.

The gain is related to the directivity by the antenna efficiency factor:

$$G = kD$$  \hspace{1cm} (3.5)

where $k$ is the efficiency factor ($0 \leq k \leq 1$), dimensionless. These losses can be of various forms and depends on the losses that are included in the gain measurements.

The gain of an antenna may be measured by comparing the power density of the antenna under test (AUT) with an antenna of known gain, such as a horn antenna or dipole.

$$\text{Gain} = G = \frac{P_{max}(\text{AUT})}{P_{max}(\text{ref. ant.})} \times G(\text{ref. ant.})$$  \hspace{1cm} (3.6)
3.2.4 Bandwidth

For the purposes of this research, antennas with broadband operation are important so that broadband responses from various targets may be measured without changing the test configuration. This is the motivation for designing the double ridged horn antenna described in Section 3.3.3). For broadband antennas, the bandwidth is described in terms of a ratio of the upper frequency of operation to that of the lower. For example, a 4:1 bandwidth could describe an antenna with a 1–4 GHz operational range. For narrow bandwidth designs, bandwidth is expressed as a percentage of the upper minus the lower frequency divided by the center frequency.

3.2.5 Input Impedance

The input impedance of the antenna should ideally match that of the driving transmission line throughout the band of interest. When the feed is not matched across this band, then mismatch losses become significant and steal from the radiated energy. The feed structure is commonly the band limiting component of the broadband antenna structure. Measurements of the input impedance can be expressed as the SWR and can be measured very easily across the band of interest using a network analyzer.

3.3 Antennas of Interest

3.3.1 Standard Gain Horn

The standard gain horn is probably one of the simplest and most widely used antenna structures of all. The horn is a flared out waveguide and produces a uniform phase front with a larger aperture than the normal waveguide, thereby providing greater directivity. The standard horn used in this effort minimizes aperture to free space discontinuities by using a gradual exponential taper in the flare. The standard gain horn antenna typically has a cutoff wavelength $\lambda_c$ of twice its broad dimension, $2a$. Due to its small bandwidth, the standard gain horn will be
used strictly for benchmarking purposes.

The directivity of the horn can be expressed in terms of an effective aperture \[1\],

\[
D = \frac{4\pi A_e}{\lambda^2} = \frac{4\pi \varepsilon_{ap} A_p}{\lambda^2}
\]  \hspace{1cm} (3.7)

where

- \(A_e\) is the effective aperture, \(m^2\)
- \(A_p\) is the physical aperture, \(m^2\)
- \(\varepsilon_{ap}\) is the aperture efficiency or \(A_e/A_p\)

For a rectangular horn, \(A_p = a_E a_H\). \(a_E\) and \(a_H\) are the physical dimensions of the width and height of the aperture, respectively and are assumed that each are at least 1\(\lambda\). Aperture efficiency has been experimentally calculated as being 0.6 for most general horn antennas by Alford \[1\]. Using \(\varepsilon_{ap} \simeq 0.6\), the directivity can be approximated as

\[
D \simeq \frac{7.5 A_p}{\lambda^2}
\]  \hspace{1cm} (3.8)

or

\[
D \simeq 10\log\left(\frac{7.5 A_p}{\lambda^2}\right) \text{ (dBi)}
\]  \hspace{1cm} (3.9)

or more accurately,

\[
D \simeq 10\log(7.5a_E a_H \lambda) \text{ (dBi)}
\]  \hspace{1cm} (3.10)

### 3.3.2 Conical Spiral

The Conical Spiral Antenna (CSA) or tapered helix, is a circular polarized antenna with little directivity and is considered to be a frequency independent structure. The conical spiral retains the frequency independent properties of the planar spiral while providing broad-lobed unidirectional circularly polarized radiation off of the small end of the antenna. The basis of operation of these antennas has been investigated by Rumsey \[11\] and is based on two principles: the angle principle and the truncation principle. The angle principle says that the performance of an antenna that is defined entirely by angles will be frequency independent. Antennas defined entirely by angles are infinite in size, so an additional consideration is needed. The truncation principle
Figure 3.4: Geometry of the conical spiral antenna.

says that the antenna must have an “active region” of finite size that is responsible for the radiation at a particular frequency. As the frequency is changed, the active region moves on the antenna in such a way that the electrical dimensions remain the same.

The two arms of the conical spiral are fed at the center point from a coaxial cable bonded to one of the arms, the spiral acting as a balun. In most cases, a dummy cable is bonded to the other arm and acts as a ground plane for the microstrip. The taper of the microstrip arms tends to provide an impedance transformation [12].

The bandwidth depends on the ratio of the base diameter ($\lambda/2$ at the lowest frequency) to the truncated apex diameter ($\lambda/4$ at the highest frequency). Therefore, the physical size of the structure determines the bandwidth. The bandwidth can be approximated as
For the RHCP conical spiral, the radiation is maximum in the forward direction and is RHCP while a small amount of left hand circular polarized (LHCP) radiation is emitted in the backward direction [11]. Upon receiving, the RHCP CSA receives the RHCP component of the incident field from the forward direction while LHCP radiation is not coupled.

\[
BW = \frac{f_{\text{max}}}{f_{\text{min}}} = \frac{\lambda_{\text{max}}}{\lambda_{\text{min}}} \approx \frac{D}{d} \tag{3.11}
\]

3.3.3 Double Ridged Horn

For this research, a high directivity and broadband antenna was needed. Very few radiating structures exhibit both of these qualities. In the Electromagnetic Compatibility and Susceptibility community, the Double Ridged Horn antenna is a favorite
antenna due to its radiation pattern and high gain. It was determined that designing a double ridged horn antenna would be best suited for this application.

Adding a double ridge to a waveguide, or horn antenna, lowers the cutoff frequency because of the capacitive effect at the center of the ridge spacing [3]. In principle, this cutoff can be lowered substantially by minimizing the ridge spacing, $d$, sufficiently which also lowers the input impedance.

The ridges also provide greater separation between modes than do typical waveguide horn antennas. The general design is shown in Figure 3.6.

The equations for the cutoff wavelengths of the various modes in ridged waveguides have been derived by Cohen [6] and Walton[7] and are:

$$\frac{B}{D} \tan \theta_2 - \cot \theta_1 + \frac{B_c}{y_{01}} = 0$$  (3.12)

$$\frac{B}{D} \cot \theta_2 + \cot \theta_1 - \frac{B_c}{y_{01}} = 0$$  (3.13)

where

$$\theta_1 = \frac{360}{\lambda_c} \left( \frac{A - S}{2} \right) \text{degrees}$$  (3.14)

$$\theta_2 = \frac{360}{\lambda_c} \left( \frac{S}{2} \right) \text{degrees}$$  (3.15)
These equations are solved and plotted in Figure 3.7.

The design of the double ridged horn [7] is based on three dimension ratios: $s/a$, $b/a$ and $d/b$. The dimensions chosen for the design used in this research effort are:

- $s = 1.9\text{cm}$ (width of ridge)
- $a = 5.5\text{cm}$ (inside width of throat)
- $b = 1.8\text{cm}$ (inside height of throat)
- $d = 0.2\text{cm}$ (spacing between ridges at throat)

Which gives the following ratios:

- $s/a = .35$
- $b/a = .33$
- $d/b = .11$
Using the graph in Figure 3.7, this equates to the following cutoff frequencies:

- $\lambda_{c,TE_{10}} = 1.09$ GHz
- $\lambda_{c,TE_{30}} = 10.9$ GHz

Thus the antenna is usable over the range 1.09 to 10.9 GHz for a bandwidth of 10:1 or 165%.

The cutoff for $\lambda_{c,TE_{30}}$ is relatively insensitive to changes in $d/b$, but $\lambda_{c,TE_{10}}$ changes significantly, where the maximum usable bandwidth increases very rapidly as the gap becomes smaller.

The design considerations for the ridges were such that a smooth taper transitions the 50 ohm feed to the impedance of free space. For ease of fabrication, the ridges were kept at a single width. With a constant ridge width, the overall length of the horn flare should be as long as possible to allow the impedance transformation to take place over at least half a wavelength at the lowest operating frequency.

The gap between the actual coaxial feed center conductor and the shorting plate must be less than one half of a wavelength at the highest frequency. The gap dimension $d$ for this design is 0.7cm, which accommodates a high cutoff frequency of 21 GHz.

The dimension $b$ must be kept less than a wavelength at the highest operating frequency to prevent propagation of the $TE_{02}$ mode. This dimension initially was 9cm, but was changed to 1.8cm with the deflector plates in the throat (see Figure 4.6). These plates were added to suppress higher order modes while keeping the dimension ratios such that the fundamental mode still propagates. Another critical dimension ratio is that of $s/a$. Since the ridge width is dictated by the 1.9cm thickness of the aluminum sheet used to build them, dimension $a$ has little flexibility. This was another motivation for the use of the deflector plates. The goal was to try and design a throat that had dimensions that varied in the direction of propagation, enabling a broader band of operation.

Final antenna performance results are presented in Chapter 4.
3.4 Reflection and Transmission of EM Waves

When an electromagnetic wave propagates through free space and illuminates a surface characterized by $\mu$ and $\epsilon$, energy will be reflected, transmitted or absorbed [5]. For simplicity, we will first consider our targets as perfect electric conducting (PEC) surfaces with $\epsilon_r = \epsilon' - j\sigma/\epsilon_0\omega = \infty$ as the conductivity $\sigma$, the reciprocal of resistivity, is infinite. This means that the skin depth will be zero because of the following:

$$\delta = \frac{1}{\sqrt{f\pi\mu\sigma}} \quad (3.16)$$

This implies that the fields inside the conductor are zero and that the electrons respond instantly to an applied electric field. However, because these electrons represent a charge density, their movement creates their own electric field, which constitutes the scattered field. These electrons move only when the net electric field is nonzero, so the field created by the electrons is in the opposite direction to the incident field. Therefore, when the scattered field is equal and opposite to the incident field, the total field on the conductor is zero and a net force is no longer acting to move the electrons. This is due to the boundary condition of a perfectly conducting metal having zero tangential electric fields.

Since the incident wave is time varying, the equilibrium does not last and the electrons must move in response to the changing incident field in order to maintain the zero tangential surface field. The time varying incident field causes a time varying charge separation on the conductor, in turn creating a current flow. This charge acceleration and deceleration (or time varying current) is the source for the scattered field. Figure 3.8 shows a uniform plane wave propagating in the $+z$ direction in air and normally incident on a perfect conducting surface at $z = 0$. The incident wave is chosen to be in the $x$ direction and therefore the magnetic field is in the $y$ direction. The incident phasor field representations are given below.

$$E_i(z) = \hat{x}E_i0e^{-j\beta_iz} \quad (3.17)$$

$$H_i(z) = \frac{\hat{y}}{\eta_0}E_i0e^{-j\beta_iz} \quad (3.18)$$
Figure 3.8: Plane wave incident normally on a perfectly conducting boundary.

The presence of the boundary conditions requires that the reflected electric field continues to be in the $x$ direction but traveling in the opposite direction of the incident field. This forces the reflected magnetic field to be propagating in the $-z$ direction with the field orientation in the $-y$ direction. The reflected phasor field representations are given below.

$$E_r(z) = \hat{x}E_{rO}e^{+j\beta_1z} \quad (3.19)$$

$$H_r(z) = -y\frac{E_{rO}}{\eta_1}e^{+j\beta_1z} \quad (3.20)$$

When an electromagnetic field is normally incident on a lossless dielectric medium, non-zero electromagnetic fields may exist within the dielectric and propagate. Since the dielectric constants of media 1 and 2 are assumed to be different, as shown in Figure 3.9, part of the incident energy will be transmitted into medium 2 where it will propagate in the $+z$ direction while part of the incident energy will be reflected in the $-z$ direction. The phasor fields for the incident, reflected and transmitted waves are given as

$$E_i(z) = \hat{x}E_{iO}e^{-j\beta_1z} \quad (3.21)$$
Figure 3.9: Plane wave incident normally on a lossless dielectric boundary.

\[ H_i(z) = \hat{y} \frac{E_{iO}}{\eta_1} e^{-j\beta_1 z} \]  
(3.22)

\[ E_r(z) = i \hat{x} E_{rO} e^{+j\beta_1 z} \]  
(3.23)

\[ H_r(z) = -\hat{y} \frac{E_{rO}}{\eta_1} e^{+j\beta_1 z} \]  
(3.24)

\[ E_t(z) = i \hat{x} E_{tO} e^{-j\beta_1 z} \]  
(3.25)

\[ H_t(z) = \hat{y} \frac{E_{tO}}{\eta_1} e^{-j\beta_1 z} \]  
(3.26)

These principles may be expanded into multiple dielectric boundaries with various angles of incidence to analyze the phenomena taking place with device illumination. Although this analysis is applicable only for simple boundary problems, advanced numerical method simulators can be used for more complicated illumination scenarios.
3.5 Network Analyzer Calibration

For most measurements, a normal short, open, load and through (SOLT) calibration procedure was used. The calibration reference plane was set at the antenna connector and not at the ideal location of the aperture of the antenna. Another calibration method that proved quite useful was simply subtracting the environment (including the target mount) from each measurement. The Agilent 8510C math feature was used to subtract the empty environment trace from the actual data trace. Although no phase information was preserved, the magnitude measurements are accurate and are very sensitive to small variations in the received field.

3.6 Summary

Electromagnetic illumination requires that high directivity antennas are utilized with sufficient radiating properties. In this chapter, the foundations for antenna characterization have been established as it applies to the research presented here. Furthermore, the theory behind the double ridged horn antenna has been examined. The basics of electromagnetic waves incident on metallic and lossless dielectric boundaries have been established. A brief discussion of network analyzer calibration procedures appropriate for this application has been presented.
Chapter 4

Antenna Characterization Results

4.1 Time Gating

The time gating feature of the Agilent 8510C Vector Network Analyzer (VNA) provided a means for gating out unwanted reflections from the test environment. Because an anechoic chamber was not readily available for the antenna characterization, it is shown that this type of gating improves measurement results.

The method used is as follows. The 8510C is calibrated using a standard SOLT calibration up to the feed of the antenna. Depending on the antenna and the number of adapters that cannot be included into the calibration (and the length of feed cable), the port extension feature is used to “dial out” the extra adaptors or cables. The transmit antenna is setup to radiate out into free space. A metallic reflector is placed at a known distance in the main beam of the transmit antenna. The measured distance is subtracted from the known distance, multiplied by two and divided by the speed of light. The resulting time value, which is typically several nanoseconds, is then entered into the network analyzer to accommodate for the electrical length. Figure 4.1 shows the test configuration for time gating calibration. This calibrates the range for any type of unknown cable lengths, adaptors, feeds, etc. Once the antenna range dimensions agree with the distance (or time) reported by the VNA, accurate time gating can be utilized. Figure 4.2 demonstrates the effect of the peaks created by a metal sheet (at 20 ns), a metallic table leg (at 13 ns) and a target of microstrips
(at 4 ns). Once it is determined which peak is associated with the target of interest, clutter reflections can easily be time-gated. This procedure has been used in the majority of the remaining measurements to create essentially a spatially filtered test range.

4.2 Standing Wave Ratio

The standing wave ratio (SWR) is defined as a measure of mismatch in a line and ranges from 1 to $\infty$, 1 being a perfect impedance match. The SWR can be defined as

$$\text{SWR} = \frac{V_{\text{max}}}{V_{\text{min}}} = \frac{1 + |\Gamma|}{1 - |\Gamma|}$$

(4.1)

4.2.1 Standard Gain Horn Antenna

The SWR for a Narda standard gain horn (model 642) is shown in Figure 4.3 and displays the usable bandwidth of the antenna as being between 5.3–8.2 GHz. Although the SWR is flat for this band, it is clear that this antenna will not be of much interest due to its lack of bandwidth. However, it provides a nice reference for the remaining
Figure 4.2: Time domain analysis of target with clutter.

Figure 4.3: SWR of Narda (model 642) 5.3–8.2 GHz standard gain horn antenna.
4.2.2 Conical Spiral Antenna

The conical spiral antenna has attractive features such as an SWR average of approximately 2 and a bandwidth of greater than 6:1. The disadvantage of this structure is the broad beamwidth and low directivity. A picture of the RHCP conical spiral antenna is shown in Figure 4.4. The SWR is relatively flat from 2–16 GHz except for a major peak at approximately 10.5 GHz and is shown in Figure 4.5. Although
the SWR remains under 2 up to 16 GHz, the actual band of operation has an upper frequency limit of approximately 10 GHz, (as discussed in Section 4.4.2) due to large variations in it’s transmitted response.

4.2.3 Double Ridged Horn Antenna

The double ridged horn antenna is shown in Figure 4.6. In the second picture, the deflector plates can easily be seen inside of the throat area. The overall dimensions of the antennas are:

Width of aperture = 35.7 cm
Height of aperture = 27.5 cm
Overall length, including throat assembly = 52 cm

The SWR measurements of each of the double ridged horn antennas are shown in Figure 4.7. Each antenna exhibits a unique SWR response with an average value of approximately 2. This equates to a reflection coefficient of 0.33 and a return loss of 9.5 dB. The antennas have a relatively consistent SWR envelope but display independent deficiencies across the band. Horn 1 has approximately 4.5 dB more return loss in the 7–9 GHz band compared to that of horn 2. Horn 2 has a peak in the SWR at 2 GHz of approximately 2.8, which equates to 6.4 dB of return loss.

Matching improvements in the feed structures could drastically improve the overall performance of these antennas.
4.3 Radiation Patterns

The radiation pattern is perhaps the most significant parameter of interest for device illumination. It is obvious that the broader beamwidths produce unwanted reflections. The radiation patterns in the following measurements were made keeping $\theta$ constant at 90 degrees and varying $\phi$ from 0 to 360 degrees. The data collection process was automated using a LabVIEW network analyzer controller shown in Figure 5.2.

4.3.1 Standard Gain Horn Antenna

Figures 4.8 and 4.9 show the electric field radiation pattern of the Narda standard gain horn antenna and demonstrates the usefulness of the time gating feature of the VNA. In Figure 4.8, the pattern was measured without the use of time gating. The size of the minor and back lobes are substantially larger than that of Figure 4.9, where time gating was utilized. Approximately 12 dB of false lobe suppression has been provided by time gating.

Although this horn has a narrow beamwidth of approximately 35 degrees with
Figure 4.8: Radiation pattern of the Narda 5.3–8.2 GHz standard gain horn antenna without time gating.

Figure 4.9: Radiation pattern of the Narda 5.3–8.2 GHz standard gain horn antenna with time gating.
very little side and back lobes, the low bandwidth (less than 2:1) is not sufficient for the research presented here.

### 4.3.2 Conical Spiral Antenna

The radiation pattern of the conical spiral antenna is expected to be hemispherical, but as shown in Figure 4.10, the antenna pattern has many deficiencies. This can be accounted for in the inherent lack of directivity and consequently, multipath phenomena from local surroundings. Although the time gating feature was used for this measurement, it does not completely filter reflections that are coupled in by local objects that lie within the time gated region (i.e., the floor). These reflections can cause phase shifts that ultimately lead to field cancellation. The low directivity and multipath problems were the main contributors to initial unsuccessful device illumination.
4.3.3 Double Ridged Horn Antenna

The electric field radiation pattern of the double ridged horn antenna in Figure 4.11 exhibits high directivity and is approximately 30 degrees wide. The high directivity provides passive gain to the transmit and receive antennas in the direction of the main lobe. The narrow main lobe allows the antenna to be pointed in the direction of the target and minimizes unwanted reflections. The pattern has similar characteristics to that of the Narda standard gain horn, but exhibits larger side lobes. These lobes may become problematic if the antennas are placed side by side with little distance between them.
4.4 Directivity and Gain

The two antenna method is used to calculate the gain of each of the antennas [9]. When choosing this method, it is assumed that the antenna sets have a high degree of dimensional stability and are identical in their construction, and therefore their performance. For our purposes, we will make this assumption, although it is clear from the SWR measurements made in Section 4.2.3 that this is not the case for the double ridged horns.

This method is based on the Friis transmission formula

\[ P_r = P_o G_A G_B \left(\frac{\lambda}{4\pi R}\right)^2 \]  

(4.2)

where

- \( P_r \) is the power received (W),
- \( P_o \) is the power transmitted (W),
- \( G_A \) is the power gain of antenna A,
- \( G_B \) is the power gain of antenna B, and
- \( R \) is the distance between the antennas (m).

Unlike most definitions of gain, the gain calculated here incorporates all sources of loss, including mismatch and polarization losses. With this calculation, it is assumed that there are negligible polarization losses and that the measurements are made satisfying the far field criteria. The far field criteria for each antenna set is as follows:

- Narda Horn: \( L = 24 \text{ cm}, \text{ far field at } 8.2 \text{ GHz } = 3.15 \text{ m } = 10.3 \text{ ft.} \)
- Conical Spiral: \( L = 13 \text{ cm}, \text{ far field at } 12 \text{ GHz } = 1.35 \text{ m } = 4.4 \text{ ft.} \)
- Double Ridged Horn: \( L = 52 \text{ cm}, \text{ far field at } 10 \text{ GHz } = 18.0 \text{ m } = 59.2 \text{ ft.} \)

Clearly, the far field criteria cannot be satisfied for the double ridged horn and the gain measurements are made at a distance of 14 ft. In logarithmic form, the Friis transmission formula can be written as

\[ (G_A)_{\text{dB}} + (G_B)_{\text{dB}} = 20\log\left(\frac{4\pi R}{\lambda}\right) - 10\log\left(\frac{P_o}{P_r}\right). \]  

(4.3)
Figure 4.12: Calculated directivity and measured gain of the Narda standard horn.

Based on the assumption that the antennas have identical gains, then equation 4.3 can be written as

$$(G_A)_{dB} = (G_B)_{dB} = \frac{1}{2} \left[ 20 \log \left( \frac{4 \pi R}{\lambda} \right) - 10 \log \left( \frac{P_o}{P_r} \right) \right].$$

(4.4)

This equation is used in the next few sections to calculate the gain of the antennas across their band of operation. This is accomplished by making an S21 measurement across the band for a given antenna set. S21 is squared to give a power ratio and used in Equation 4.4.

### 4.4.1 Standard Gain Horn Antenna

The directivity and gain of the standard horn antenna are compared in Figure 4.12. The directivity was calculated using Equation 3.10. It is apparent that the efficiency of this antenna is remarkably high in the operational band. At approximately 5.5 GHz, the gain is larger than the directivity. Clearly this is not possible and may be attributed to approximations made in the aperture efficiency term in Equation 3.10.
A value of 0.6 was used, but the aperture efficiency of these particular horns might be higher since they have an exponential taper.

4.4.2 Conical Spiral Antenna

The gain of the conical spiral antenna is shown in Figure 4.13. A directivity calculation was not attempted for this structure. From the gain, the operational bandwidth appears to be between 2 and 10 GHz. It is clear from Figure 4.13 that there are wide variations in the antenna response. Considering the radiation plot in Figure 4.10 and the SWR plot in Figure 4.5, deviations in the response could be attributed to reflections internal to the conical spiral feed or balun. The fluctuations in the passband gain above 6 GHz reach 10 dB and more.

4.4.3 Double Ridged Horn Antenna

The directivity and gain of the double ridged horn antenna are shown in Figure 4.14. The efficiency of these antennas are clearly lacking in portions of the band, especially
at higher frequencies. This is predominately due to the antenna feed mismatch that was shown in Figure 4.7). Excitation of higher order modes might also be taking place in various portions of the structure. However, the antenna displays a reasonable gain of 10 dB and higher across the operational band.

4.5 Summary

The antenna plays a key role in the device illumination system. It is vital to have a broadband radiating structure with high gain and a narrow main beam. While the conical spiral antenna is broadband, the lack of directivity makes this antenna useless for this work. The double ridged horn antenna exhibits the characteristics needed for accurate and reproducible illumination measurements.
Chapter 5

Device Illumination Results

There are very distinct signatures of simple and complex passive and electronic devices. The test environment and radiating equipment plays a key role in obtaining meaningful, reproducible and accurate results. The ability to time-gate out unwanted reflections and multipath residues greatly enhances accuracy of results when an anechoic chamber or an antenna test range is not available.

5.1 Illumination of Various Structures

5.1.1 Test Configuration

The test configuration used for measuring the backscattered power from various targets is shown in Figure 5.1. The transmit antenna is connected to test port 1 of the network analyzer and the receive antenna is connected to test port 2. All measurements were made in terms of $S_{21}$ and it was assumed that the ports were matched over the frequency band of the measurement. This implies that $V_1^-$ and $V_2^+$ are zero and that no power is lost due to mismatch and discontinuities in the connections.

The Agilent 8510 network analyzer has a system dynamic range of 93 dB and has convenient features for suppressing noise such as data averaging, trace math and trace smoothing. A LabVIEW control system was designed to collect, parse and plot data in a variety of formats. The control system encompasses an antenna radiation
Figure 5.1: Test configuration using the Agilent 8510C Vector Network Analyzer.

pattern utility that plots the radiation pattern in a polar plot. The graphical user interface is shown in Figure 5.2.

All measurements were made in the near field. A semi-anechoic chamber lined with microwave absorbing material was constructed to suppress environment noise and multipath. The chamber is approximately 4 feet square and has 5 sides. The open side is the area where the transmit and receive antennas were placed.

5.1.2 Microstrip Illumination

Metallic strips resonate when illuminated by an electromagnetic wave when the strip length is approximately $\frac{\lambda}{2}$. There exists multiple resonant frequencies for a single length microstrip at $\frac{\lambda}{2}$, $\frac{3\lambda}{2}$, $\frac{5\lambda}{2}$, etc. The microstrip has a peaked $S_{21}$ response when the wavelength of the incident wave is twice that of the electrical length of the target microstrip. Simulations were performed using the Ansoft Ensemble Method of Moments simulator. Simulation results are shown in Figure 5.3 displaying a large current density in the middle of the 4cm strip at 3.4 GHz. The remaining 2 and 3cm strips show approximately zero current densities. The current densities change
Figure 5.2: LabVIEW control system for the Agilent 8510C Vector Network Analyzer.

Figure 5.3: Current density at 3.4 GHz for a 4cm strip illuminated by a plane wave.
from each of the strips as the incident wave frequency increases and reaches twice the electrical length of the other strips.

In Figure 5.4, the measured and simulated response of a 5cm and 3cm long microstrip are shown. For the 5cm strip, there is agreement at the first resonance of approximately 3 GHz ($\lambda/2$) and at the second resonance of approximately 9 GHz ($3\lambda/2$). For the 3cm strip, there is agreement at the first resonance at approximately 5 GHz ($\lambda/2$) and at the second resonance of approximately 16 GHz ($3\lambda/2$). As the strip length becomes shorter and the width remains unchanged, the resonant frequency increases and the peak becomes more broad. The broadband nature of the response is caused by the increase in the surface resistance with increasing frequency, resulting in a decrease in the microstrip $Q$. The lower $Q$ results in a higher bandwidth response.

The discrepancies in magnitude are due to errors in converting from the simulated radar cross section (RCS) to $S_{21}$. The simulator results are in terms of RCS ($\sigma$) which is defined as follows.

$$\sigma = 4\pi R^2 \frac{P_{\text{scattered}}}{P_{\text{incident}}}$$  \hspace{1cm} (5.1)

RCS is defined at the target and the measured $S_{21}$ is defined at the calibration plane of the antennas. Therefore the conversion process must incorporate the difference in distance and is merely an estimate.

The resonating frequency of a microstrip decreases slightly as the width of the microstrip increases. As the geometry approaches a square, the induced current on the strip begins to travel between diagonal corners as well as opposing ends. This phenomena is shown in Figure 5.5. It is not until the width becomes close to the length, that a transverse resonance begins to appear. Transverse resonance occurs only when the incident wave contains both a horizontal and a vertical component.

Figure 5.6 shows the simulated and the measured response of a 3cm long microstrip of various widths.

As the strip becomes wider and approaches a square, the response begins to broaden out and the resonant peak disappears. The response becomes high pass in nature. Specular reflections become the dominating mechanism as the wavelength
Figure 5.4: Simulated and actual responses: (a) of a 5cm; and (b) a 3cm microstrip.
Figure 5.5: Current density vector plot of a 3cm long microstrip: (a) 0.1cm wide; (b) 0.4cm wide; and (c) 3.0cm wide.
Figure 5.6: Simulated (a) and actual (b) response of a 3cm long microstrip of various widths.
becomes smaller than the electrical size of the square. This is shown in the simulated results in Figures 5.7 and 5.8.

A square plate begins to reflect high frequencies in a specular fashion as the wavelength becomes smaller than the electrical size. There is substantial disagreement between the actual and simulated response for square conducting plates. This is attributed to the antenna response deficiencies at the higher frequencies.

Multiple microstrips on a single target respond in the same fashion as do the individual strips. Linear superposition appears to hold only when sufficient spacing is placed between the strips. This distance has been experimentally determined as approximately $\lambda/2$. Complex field interactions begin to dominate when the spacing becomes smaller. As the spacing continues to decrease, it responds as if the strips were a single metallic entity. This is shown in Figures 5.9 and 5.10. The inconsistencies between actual and simulated results may be due to the limitations of the simulator used. The simulator requires that targets be placed on an air box that is not quite large enough to allow sufficient $\lambda/2$ spacing between the air box boundaries and the neighboring strips.

A summary of the various microstrip length responses are shown in Figure 5.11. The lengths range from 1cm to 6cm in 0.2cm increments. The second resonances at $3\lambda/2$ begin to appear for the lengths starting at approximately 4cm.

### 5.1.3 Laptop Illumination

Unique signatures exist for each of the laptops illuminated. Each laptop exhibits peaks in different portions of the frequency spectrum. The magnitude of these peaks change as various components and drives are removed. It appears as if various mechanisms are contributing to the responses of compact electronic devices. Resonant behavior occurs for metallic traces that are aligned with the polarization of the incident field and have sufficient spacing with other scatterers. Specular reflections occur for metallic interfaces that are electrically larger than the incident wavelength. Dielectric boundaries will reflect electromagnetic energy and reflections depend on frequency, material properties and losses.
Figure 5.7: Simulated (a) and actual (b) response of various size square metallic plates.
Figure 5.8: Radar cross section (RCS) simulation results for a 3cm microstrip with widths varying from 0.1cm to 3.0cm.
Figure 5.9: Simulated and actual results for 4, 3, and 2cm microstrips.

Figure 5.10: Single and multiple $S_{21}$ strip response demonstrating linear superposition.
Figure 5.11: Actual $S_{21}$ response of various length microstrips ranging in length from 1.0 to 6.0 cm.
Figure 5.12: $S_{21}$ responses of Laptop 2 and Laptop 1 illuminated from the bottom side. Unique peaks for Laptop 1 located at 4.4 and 4.8 GHz. Unique peaks for Laptop 2 located at 5.8 and 7.7 GHz.

When illuminating two different laptops, it is clear that there are distinct signatures present for each device. The angle of the incident wave does have some bearing on the response magnitude but does not affect the location of the $S_{21}$ peaks. Figure 5.12 displays the $S_{21}$ signatures of Laptop 1 and Laptop 2 illuminated from the bottom side. It is evident that the Laptop 1 laptop has peaks at 4.4 and 4.8 GHz while the Laptop 2 does not. Conversely, Laptop 2 has peaks at 5.8 and 7.7 GHz while Laptop 1 does not. These peaks equate to electrical lengths of 3.4cm, 3.1cm, 2.6cm and 1.9cm respectively.

Many of the peaks in Figure 5.12 are in the same location for both devices. The antennas do not exhibit a flat response and therefore appear in device signatures. Although the calibration in Section 3.5 was used, the deficiencies in the band remain in the response. A quick glance at the signatures would lead one to conclude that there exists many unique peaks but in actuality, only a few exist. Without disassembling the laptops, it is not clear as to whether there are resonant objects responsible for the peaks or if the peaks exist because of more complicated boundary/field interactions.
A key contributor to the uniqueness in each response is the magnitude of the reflected power. Laptop 1 exhibits approximately 4 dB more power in portions of the band compared to that of Laptop 2. The boundaries that are presented by the devices’ printed circuit boards, the display, the plastic housing, the batteries and associated drive hardware all contribute to the variance in the scattered power. Power will be dissipated in lossy dielectric materials and scattered by other mediums. The scattering will be dependent on the intrinsic impedance \( \eta = \sqrt{\mu/\epsilon} \) of each medium which determines the reflection and transmission characteristics from each interface. The multiple boundary problem applies here. The system must be treated as a steady state boundary problem where the total reflected energy is comprised of reflection contributions from each interface. The thickness of each medium will have a significant impact on the amount of power being reflected and transmitted through each medium. If there are lossy materials, then \( \eta \) will be complex, thus contributing to a phase variation as well as magnitude variations. Clearly, there can be numerous interfaces in a electronic device that the incident wave encounters, each contributing to the multiple boundary problem.

The \( S_{21} \) response of Laptop 2 changes by approximately 6 dB in certain portions of the band by removing the drives and battery of the machine and is shown in Figure 5.13. The peaks change only in magnitude and not in frequency. This suggests that the laptop drives act more as a specular reflector than a resonant object. The drive and battery casings are constructed of metal and would explain the increase in reflected power. There exists one peak at approximately 5.7 GHz with the loaded laptop that does not exist for the stripped laptop. This frequency equates to a length of 2.6cm. Again, disassembly of the drives and battery would be required to determine if this peak has been created by a resonant object of this length.

5.2 Summary

Simulated and actual measurement results for simple and complex targets have been presented. It is clear that when devices are illuminated with electromagnetic energy, there exist a unique response for a given geometry or structure. The response
can be of a resonant nature, a specular nature or a combination of both. The signature is inherently dependent on material composition, geometry and boundary conditions.
Chapter 6

Conclusions

6.1 Conclusions

Detecting resonances and signatures of structures requires antennas with certain radiating properties. These properties consist of high gain, minimal side lobes and narrow beamwidths. The double ridged horn antenna designed for this application has proven to be beneficial due to its broadband operation and high gain. The conical spiral antenna, while broadband, has poor directivity and a broad main lobe which inherently contributes to unwanted reflections.

Simple and complex devices exhibit unique signatures when illuminated with electromagnetic energy. Microstrips that are an odd multiple of $\lambda/2$ in length exhibit a resonant response. As the strip becomes wider, the resonant peak begins to shift slightly lower. As the strip width approaches that of a square, the resonant peak diminishes and the response becomes high pass in nature.

When multiple microstrips are placed on a single target and illuminated, contributions from each of the strips are evident in the received response. These components only exist when the strips are spaced approximately $\lambda/2$ apart. As the spacing decreases, then complex field interactions begin to dominate and the response becomes unpredictable.

When illuminating complex devices, such as a laptop computer, the signatures become more complex in nature. Material thickness, metal power and ground planes,
microstrip traces and differences in various boundary intrinsic impedances all contribute to the resulting unique signature.

6.2 Future Research

There are many unexplored avenues in which this research could branch:

- Specular responses from targets need to be analyzed where the wavelength is much smaller than the physical size of the target. Utilizing an automated positioner would provide consistent and repeatable results.

- Devise an algorithm that unwinds the electrical distance out of the phase response of the target so that meaningful phase content may be extracted and analyzed.

- Develop an antenna calibration that provides a calibration plane at the radiating end of the antenna.

- In depth use of an electromagnetic simulator to explore theoretical responses of complicated structures and devices, and bypassing obstacles such as imperfections in the test environment and antenna deficiencies.

- In depth analysis of complex metallic and dielectric boundary scenarios.
Bibliography


