NOVEL MATERIALS, FABRICATION TECHNIQUES AND ALGORITHMS FOR MICROWAVE AND THZ COMPONENTS, SYSTEMS AND APPLICATIONS

by

Min Liang

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ABSTRACT

This dissertation presents the investigation of several additive manufactured components in RF and THz frequency, as well as the applications of gradient index lens based direction of arrival (DOA) estimation system and broadband electronically beam scanning system. Also, a polymer matrix composite method to achieve artificially controlled effective dielectric properties for 3D printing material is studied. Moreover, the characterization of carbon based nano-materials at microwave and THz frequency, photoconductive antenna array based Terahertz time-domain spectroscopy (THz-TDS) near field imaging system, and a compressive sensing based microwave imaging system is discussed in this dissertation.

First, the design, fabrication and characterization of several 3D printed components in microwave and THz frequency are presented. These components include 3D printed broadband Luneburg lens, 3D printed patch antenna, 3D printed multilayer microstrip line structure with vertical transition, THz all-dielectric EMXT waveguide to planar microstrip transition structure and 3D printed dielectric reflectarrays.

Second, the additive manufactured 3D Luneburg Lens is employed for DOA estimation application. Using the special property of a Luneburg lens that every point on the surface of the Lens is the focal point of a plane wave incident from the opposite side, 36 detectors are mounted around the surface of the lens to estimate the direction of arrival.
(DOA) of a microwave signal. The direction finding results using a correlation algorithm show that the averaged error is smaller than 1° for all 360 degree incident angles.

Third, a novel broadband electronic scanning system based on Luneburg lens phased array structure is reported. The radiation elements of the phased array are mounted around the surface of a Luneburg lens. By controlling the phase and amplitude of only a few adjacent elements, electronic beam scanning with various radiation patterns can be easily achieved. Compared to conventional phased array systems, this Luneburg lens based phased array structure has a broadband working frequency and has no scan angle coverage limit. Because of the symmetry of Luneburg lens, no beam shape variation would occur for the entire scanning range. Moreover, this alternative phased array requires much less system complexity to achieve a highly directional beam. This reduction in system complexity allows the electronic scanning system to be built at much lower cost than traditional phased arrays.

Fourth, the characterization of carbon based (Graphene and carbon nanotube) thin films on different substrates via Terahertz time-domain spectroscopy are presented in this dissertation. The substrate permittivity is first characterized. The film under test is then treated as a surface boundary condition between the substrate and air. Using the uniform field approximation, the electromagnetic properties of the film can be extracted. To improve accuracy, precise thickness of sample substrate is calculated through an iteration
process in both dielectric constant extraction and surface conductivity extraction. Uncertainty analysis of the measured thin film properties is performed.

Fifth, a coded transmitter TDS near field imaging system by employing photoconductive antenna (PCA) array is reported. Silicon lens array is used to couple and focus the femto-second laser onto each PCA. By varying the bias state of each PCA element, the ON/OFF state or power level for different PCAs can be controlled independently. The sample object is placed 10 μm away from the PCA array to measure the THz near field image. A Hadamard matrix is applied to code the 2x2 antenna array to improve the SNR. Measured results clearly indicate an improved SNR compared to individual antenna measurement. In addition, Multiphysics COMSOL and a FDTD algorithm combined with HFSS time domain simulation is used to model the physics of TDS photoconductive antenna and optimize the performance of TDS transmitter and receiver. Good agreement between simulation and experiment is obtained.

Finally, a design of a Principal Component Analysis (PCA) based microwave compressive sensing system using reconfigurable array is presented. An iterative beam synthesis process is used to realize the required radiation patterns obtained from PCA. A human body scanning system is studied as an example to investigate the compressive sensing performance using PCA generated radiation patterns. Optical images are used as surrogates for the RF images in implementation of the training PCA dictionary.
Compared to random patterns based compressive sensing system, this PCA based 
compressive sensing system requires fewer numbers of measurements to achieve the 
same performance.
CHAPTER 1 INTRODUCTION

1.1 Introduction to Additive manufacturing

Additive manufacturing (AM), often called “3D printing” or “rapid prototyping”, is an automated fabrication technology to make 3-dimensional physical object directly from digital data. Contrary to the subtractive manufacturing which realizes a product by subtracting material from a larger piece of material such as cutting out a screw from a piece of metal, it makes a product layer by layer additively.

AM was originated in the United States and was first commercialized in the late 1980s [1]. At that time it was called “rapid prototyping (RP)” or “generative manufacturing” [2] and these terms are still occasionally in use presently. In the early 1990s, several different AM processes including Laser Sintering (LS) [3] and Fused Deposition Modeling (FDM) [4] were developed and became available commercially. In the mid of 1990s, another 3D printing process which creates an object by jetting a liquid binder onto a bed of powder and doing post processing to solidify the whole structure was invented [5]. After that, through the rest of the 1990s, further research and development were mainly focused on materials such as various thermoplastics [6] and elastomeric polymers [7] in different forms to enable AM techniques to be used in more
applications [1]. As the new century begins, the focus was shifted back to improving the
AM technology by developing new printing processes. New techniques such as the Laser
Melting (LM) and Electron Beam Melting (EBM) processes were developed. These
techniques allow various alloyed materials to become available in the AM process. Over
subsequent years, more and more AM companies are founded from all over the world and
starting to develop their own printable materials and AM systems. Many new types of
materials and systems become available as the demand for AM increases. It is realized
that these techniques are not just for rapid prototyping, instead, they can be applied as a
new form of manufacturing technology. Therefore, from then on, the name “Additive
Manufacturing” has been coined. Recently, AM has received much attention with
impressive demonstrations ranging from musical instruments [8], to vehicles [9], to
housing components [10] or even entire buildings [11]. Many different structural
materials such as metal [12], polymer [13], ceramics [14], concrete [15] and even
bio-compatible materials [16] have been incorporated in various 3D printing technologies.
Due to its ability to realize desired structures with arbitrarily designed spatial distribution,
3D printing technology has been argued to be the future of manufacturing as it offers
huge potentials to revolutionize both the design and manufacturing procedures.

The technical realization of AM is based on layer by layer processes and therefore it
is also called “layer based technology” or “layer oriented technology”. The working
principle of the layer based techniques is to create a 3D physical structure from many slices with the same thickness. Each slice is fabricated according to the information from the corresponding 3D model and placed on top of the previous layer. A schematic illustration of typical AM procedure is shown in Figure 1-1. The process starts with a 3D computer aided design (CAD) model which represents the 3D object to be printed. This CAD model can be created directly from CAD software or by digital 3D scanning of a real structure [17]. After the CAD model is obtained, specialized software is used to slice the model into layer by layer cross sections. As a result, a series of layered slices with equal thickness are generated. The information of these slices including position, layer thickness and layer number is sent to a machine that could print each layer and bond it to the previous one. The printing and bonding of the layers can be done in many ways based on different physical phenomena. By printing the object layer by layer, the entire structure is built from bottom up.

These basic steps are the same for almost all variety of AM equipment available today. The differences of different equipment are how they generate the layers, how the adjacent layers are joined together to form the final part, and the corresponding built materials.
Compared to conventional manufacturing methods (such as injection molding, casting, and stamping and machining), AM approach has the following advantages.

1. Arbitrary complexity

AM approach has the ability to create 3D objects with arbitrary shape and complexity. The cost of the 3D printed components is only related to the volume of the parts, there is no additional cost or lead time for making the structure more complex. Also, with multiple printing heads, it is possible to cohesively integrate different materials at the
same location simultaneously. Therefore, AM may revolutionize product designs because of the much more flexible object geometry and material property distribution it offers. For example, 3D structures with arbitrary EM property distribution may be printed relatively easily.

2. Digital manufacturing

After an object is designed, the whole 3D printing process is accurately controlled by a computer with very little human interaction needed to realize the design. This automatic 3D printing process means that the time between design iterations can be dramatically reduced compared to conventional manufacturing methods.

3. Waste reduction

A 3D printed component is created bottom up via layer by layer processes so that only materials needed for the design are used. Therefore, material waste in AM process will be much less than conventional subtractive manufacturing techniques.

Various 3D printed microwave & THz components have been reported taking advantages of the AM technology. Components of different structures such as horn antennas [18], patch antennas [19], meander line antennas [20], multilayer microstrip [21], gradient index (GRIN) lens [22], electromagnetic band gap structure [13], THz microstrip to waveguide transition structure [23] and THz reflect-array [24]; made of different material such as all dielectric structure [25], all metal structure [26] and dielectric metal
combined structure [19, 21, 23, 24]; working at different frequencies from GHz to THz have been realized using different 3D printing techniques. In the following section, an overview of various AM techniques relevant to microwave and THz application is provided and the pros and cons for each are discussed.

1.2 Overview of 3D Printing Techniques

At the present time, there are several kinds of 3D printing techniques, all of which follow the basic steps of AM discussed in the previous section, for example, generating individual physical layers and combining them together. Various materials such as metal, plastic, ceramics or even bio-compatible materials can be used in the generation of the physical layers. According to the methods of generating physical layers and bonding adjacent layers together to form an object, five basic categories of AM processes are commercially available [2], including selective sintering and melting, powder binder bonding, polymerization, extrusion and layer laminate manufacturing (LLM). Key aspects of these five processes are discussed and some commercially available 3D printers as well as printed examples are reviewed in the next section.
1.2.1 Selective sintering and melting

The 3D printing technique using a laser to selectively sinter or melt powdered material is called selective laser sintering (SLS) [3] or selective laser melting (SLM) [27]. If an electron beam is used instead of laser, the process is called electron beam melting (EBM) [28].

A SLS printer usually includes a building chamber to be filled with powdered built material and a laser beam on top that can be scanned precisely in the XY (horizontal) plane. The bottom of the chamber is moveable in the Z (vertical) direction. During the printing process, the entire chamber is heated to a high temperature close to the melting point of powder so that they are at an optimal temperature for melting. To prevent oxidation, the chamber is often filled with shielding gas (e.g., nitrogen). The scanning laser beam is then used to fuse the powders at designated locations. As the laser beam travels in the XY plane, the melted powders cool down and solidify. After the scanning of an entire layer at designated positions, a solid layer with designed pattern is achieved. After one layer is printed, the powder bed is lowered by the amount of one layer thickness and an automated roller adds a new layer of powdered built material on the top of the previous layer. Then the selective melting process repeats until the entire object is printed. The remaining unsolidified powders are then removed after printing. The SLS technique is quite versatile since it can be used to print several classes of materials,
including plastics, metals and ceramics.

Typically, SLS fabricated metal parts such as steel and titanium are dense. They may be post-processed by cutting or welding, depending on specific materials involved. Plastic parts such as nylon and polystyrene fabricated using SLS have properties similar to those made by plastic injection molding. As an example, Figure 1-2 shows a SLS printer (EOS P800) and a metallic part made by using SLS.

Selective laser melting (SLM) is developed in particular to process metal parts that need to be very dense (> 99%). In this case, the laser beam melts the metal powders completely into liquid phase which results in a close to 100% density part after

Figure 1-2. (a) Photo of a selective laser sintering printer (SLS) (Model EOS P800; Size: 2.25 m x 1.55 m x 2.1 m) [29]. (b) A SLS printed metallic object [30].
re-solidification. SLM can be used to print many metals including stainless steel, carbon steel, CoCr, titanium, aluminum, gold and a large variety of alloys.

EBM is a similar 3D printing process in which metal powders are melted or fused by applying an electron beam under a high voltage (typically in the range of 30 ~ 60 KV) instead of a laser beam. To avoid oxidation, the process is performed in a high vacuum chamber. Because the electron beam penetrates much deeper than a laser beam, EBM allows a higher scanning speed. In addition, deeper penetration can be used for powder preheating so that the printing process works at elevated temperatures compared to the laser case. As a result, mechanical stress and distortion of printed objects are reduced and greater strength can be achieved. Figure 1-3 shows an example of an EBM 3D printer and a 3D printed object using EBM technique.

Sintering and melting processes are very suitable for applications requiring high strength and / or high temperature. Antennas printed by SLS or EBM can be very dense, void-free and very strong. The disadvantages of selective sintering and melting techniques are that the printing resolution is limited by the size of the powders (i.e., tens of microns) and a high vacuum chamber or shielding gas is needed to avoid oxidation [31].
Figure 1- 3. (a) Photo of an electron beam melting (EBM) printer (Model Arcam Q20; Size: 2.3 m x 1.3 m x 2.6 m) [32] and (b) an EBM fabricated part. [33]

1.2.2 Powder binder bonding

Powder binder bonding is another 3D printing technique that implements layer by layer bonding of powdered materials by selectively injecting a liquid binder onto the powder bed. This process was first developed in the mid-1990s. Currently, various materials such as plastics, metals and ceramics can be printed using this technique.

A typical powder binder bonding printer is very similar to a selective laser sintering printer with a piston at the bottom of chamber to adjust the height and a roller to recoat the powders. The printing process starts with depositing small drops of liquid binder onto a layer of built material powders at designated locations. The powders forming the
designed structure are bounded together while the surrounding loose powders support the next layer of the structure to be printed. The printing process is then repeated for each layer until the entire structure is completed. Compared to the sintering or melting process, this process is performed at much lower temperature. Therefore, no preheating, shielding gas or vacuum chamber are needed.

At the end of the printing process, the residual powders are removed and an infiltration process may be performed for enhanced durability. For plastic part, wax or epoxy resin can be used in the infiltration process. If this technique is used to print a metallic antenna [34], a subsequent high temperature process is needed to provide strength and durability. For example, to print a bronze object, the printed part needs to be infused into bronze powder and heated up to more than 1000 °C to replace the binder with bronze [35]. This process can also be used to print alloy materials by changing the sintering temperature and time during the infiltration process [36]. Figure 1-4 shows a powder binder bonding 3D printer and an example printed using this technique. Similar to the sintering and melting processes, the resolution of this technique is also limited by the size of the powders. For the currently available printer on the market, the minimum feature size is 0.1 mm.
Figure 1-4. (a) Photo of a powder binder bonding 3D printer (Model ProMetal S15; Size: 3.1 m x 3.4 m x 2.2 m) [37]. (b) An example printed using the powder binder bonding technique [38].
### 1.2.3 Polymerization

Polymerization is a process that selectively solidifies liquid resin using ultraviolet radiation or other power sources. Typically, photosensitive polymers are used as built material. There are several kinds of AM methods based on polymerization process. Their differences are mainly in how the photon energy is applied and how the layers are created.

Stereolithography is the most accurate polymerization process which employs an ultraviolet laser to solidify a liquid ultraviolet curable polymer. To print each layer, a laser beam scans over the surface of a liquid polymer reservoir to cure the cross section according to the designed pattern. The curing thickness can be adjusted by the laser power and laser scanning speed. After one layer is printed, the building stage descends a distance of one layer thickness. Then, a blade sweeps across the surface of the printed part, recoating it with fresh liquid polymer before the next layer is printed on top. It is possible to incorporate different materials in the printing process, thus achieving multiple material stereolithography [2]. In this case, the resin needs to be drained and replaced by the new material. After an object is printed, it is cleaned and moved into a UV chamber for a final post curing process to make it more stable. Figure 1-5 shows a stereolithography 3D printer example and an object realized by stereolithography.
Compared to other AM techniques for 3D printing of antennas, stereolithography process can achieve a very good surface smoothness and finer resolution. In fact, two-photon stereolithography process has been reported to obtain sub-micron printing resolution [39]. However, the strength of a 3D printed part by stereolithography is weaker than other techniques such as sintering, melting or powder binder bonding.

Figure 1-5. (a) Photo of a laser stereolithography 3D printer (Model 3Dsystems iPro™ 8000; Size: 1.26 m x 2.2 m x 2.28 m) [40]. (b) A sample fabricated using stereolithography technique [41].

If photosensitive polymer is applied by printer heads, the AM process is called polymer jetting. During printing, the printer head deposits photo-sensitive polymers onto a stage with designed patterns. Upon jetting, the printed photosensitive polymers are
immediately cured by an ultraviolet lamp on the printer head and unlike stereolithography, no post curing process is needed. The thickness of each layer of this process can be on the order of 20 µm, which provides a very smooth surface. Moreover, multiple types of polymers can be printed simultaneously using multiple printer heads. A gel type of polymer can be used as support material to print overhanging structures and released (e.g., water soluble support materials can be washed away) after the printing process. A schematic drawing of the polymer jetting procedure is shown in Figure 1-6. The polymer jetting method can only be applied to print polymers, limiting its applications to all dielectric antennas. Additional metallization process would be required to incorporate conductor part. It has a better resolution than sintering and powder binder bonding techniques. However, similar to stereolithography, parts printed by polymer jetting are not as strong as some of the other AM techniques.
Figure 1-6. (a) Schematic picture of the polymer jetting technique [42]. (b) Photo of a polymer jetting 3D printer (Model Stratasys Eden350V; Size: 1.3 m x 1 m x 1.2 m).

1.2.4 Extrusion technique

Extrusion, often called fused deposition modeling (FDM), is an AM process that prints an object by extruding thermoplastics through a heated nozzle. A FDM printer includes a feeding roll, a heated extrusion head and a building platform. The building materials are usually thin thermoplastic filaments which are wound up and stored in a cartridge. The thin thermoplastic filament is guided into the extrusion head by the feeding roll. During the printing process, the heated extrusion head melts down the filament and extrude it through the nozzle at designated locations on the building platform. When the extruded thermoplastic reaches the building platform, it cools down and hardens. After one layer is completed, the platform lowers down by one layer thickness and is ready for printing of the next layer. Figure 1-7 shows the schematic of a FDM printer and a printed example.

There are a number of available built materials for FDM including polycarbonate (PC), acrylonitrile butadiene styrene (ABS), polyphenylsulfone (PPSF), etc. The advantages using this technique to print antennas are the relatively simpler processing (i.e., no post processing needed) and lower printer cost compared to other AM techniques. The
disadvantage of FDM is lower resolution (about 0.25 mm [10]).

1.2.5 Layer by layer bonding

Layer by Layer bonding is an AM technique that creates a 3D structure by cutting pre-fabricated sheet or foil into designed contour and subsequently bonding a number of layers together. It is often called Laminate object manufacturing (LOM). A LOM printer consists of a building platform that can move in the z-direction, a foil supply system to supply and position the foil and a cutting device to create the contour. The LOM procedure is as follows: First, the foil is positioned and adhered to the building platform.
by a heated roller. Second, the cutting tool scans on the foil to create the designed contour and perform cross cutting on the non-model area to make it into small pieces for easier removal after printing. After one layer is printed, the platform moves down and the roller positions the next layer of foil on top of the previous layer. Then the platform moves up into position to receive the next layer and the process repeats until the entire 3D object is printed completely. A photo of a LOM 3D printer using paper material is shown in Figure 1-8 together with a 3D printed example.

The foil built materials for the LOM technique can be paper, plastic or metal [2]. The cutting tool can be a scanning laser, a knife or a milling machine. To bond adjacent layers, different methods such as gluing, soldering, ultrasonic or diffusion welding can be used depending on the material properties. Compared to other AM techniques, the advantages for using the LOM in antenna printing include lower material cost and faster building speed for large objects. The disadvantages are less accuracy (e.g., 0.3 mm for the Solidimension SD300 3D printer shown in Figure 1-8 [45]) and some material waste depending on the geometry.
Figure 1-8. (a) Photo of a LOM 3D printer (Model Solidimension SD300; Size: 450 mm x 725 mm x 415 mm) [45]. (b) An object made of paper printed using the LOM method [46].

1.2.6 AM techniques summary

Most of the AM processes currently available can be classified by the above mentioned five basic categories. Table 1-1 summaries the key features of these techniques.
In chapter II and chapter III, several kinds of microwave and THz components fabricated using different AM techniques will be reported. Compared to conventional microwave and THz components, these 3D printed components can be fabricated with much lower cost and the fabrication process is more convenient and faster. Moreover, with AM techniques, complicate structures which are very difficult or even impossible to fabricate using conventional method can be easily achieved using AM approach. Therefore, much more flexible object geometry design and material property distribution can be utilized.
1.3 Gradient index device and Luneburg lens

Gradient index (GRIN) components are EM structures that exhibit spatially-continuous variations in their index of refraction \( n \). The appeal of GRIN components comes from the fact that small, continuous variations of \( n \) along a macroscopic path can frequently be more efficient in terms of EM effects than traditional discontinuous index changes – resulting in smaller, more effective components.

Luneburg lens is an attractive GRIN component used as antenna for wide angle radiation scanning because of its broadband behavior, high gain and the ability to form multiple beams. It has a superior performance compared with conventional lenses made of uniform materials. Every point on the surface of an ideal Luneburg lens is the focal point of a plane wave incident from the opposite side. Usually, for a lens made of non-magnetic \( (\mu_r = 1) \) material, the index of refraction \( n \) distribution of a spherical Luneburg lens is given by Equation (1-1) [47]:

\[
\frac{n(r)^2}{\varepsilon_r(r)} = 2 - \left(\frac{r}{R}\right)^2
\]

(1-1)

where \( \varepsilon_r \) is the relative permittivity, \( R \) is the radius of the lens and \( r \) is the distance from the point to the center of the sphere.

Historically, many theoretical and experimental investigations have been done after Luneburg’s work in 1944 [47]. Eaton relaxed the restriction that the incident rays need to
be parallel to the symmetry axis [48,49]. Brown and Gutman designed lenses with the focal point being interior to the lens [50, 51]. Peeler et al. [52] and Slager et al. [53] investigated using only a part of the lens together with some reflecting surface to reduce the size and weight. Kay presented a procedure for finding the index variation for a spherically symmetric lens which could provide, with some restrictions, any desired beam pattern [54], while the solution of this problem was later generalized by Morgan [55].

Manufacture of the constantly changing radial permittivity profile is impossible for a spherical lens. The permittivity profile must be approximated by discretized steps, which is usually achieved/implemented as an onion like concentric spherical layers of thin molded hemispherical layers, which are both difficult to produce with acceptable permittivity and shape accuracy. During assembly of such a spherical lens, care must be taken to avoid air gaps between the layers. [56,57,58,59,60]. Many commercial manufacturers such as Emmerson & Cumming, Mayurakshi Equipments, LuneTech, Konkur, Thomson, Matsing and Rozendahl can fabricate this kind of Luneburg lens. Other methods of building the Luneburg lens using materials with variable effective dielectric constant have also been reported, including changing the thickness of dielectric plate in a waveguide [61,62,63], drilling holes on a dielectric plate to control the effective permittivity by the hole density [64,65,66], varying the effective permittivity using
complementary metamaterials [67] and adjusting the metallic patterns on a printed circuit board [68], [69]. However, these methods are mostly used for building 2-D lenses because of either intrinsic or fabrication limitations. In addition, the conventional method for building a 3-D spherical Luneburg lens are prone to tolerance issue, time consuming and have a complex fabrication process because a large number of layers need to be fabricated separately and assembled carefully.

In this work, additive manufacturing technique is used to realize a 3D broadband Luneburg lens. The desired gradient index (or, permittivity due to the relative permeability being 1) is realized by controlling the filling ratio of a polymer / air based unit cell. Efficient and accurate fabrication of the designed 3-D lens is enabled by an AM technique called polymer-jetting rapid prototyping [13]. The effective permittivity of each unit cell is designed independently based on its distance to the center of the sphere. Since the refractive index of the lens is independent with frequency as long as the long wavelength condition holds (to guarantee the accuracy of the effective medium approximation), it could operate in a quite wide frequency range. Compared to traditional Luneburg lens fabrication techniques, this 3-D Luneburg lens can be fabricated with lower cost and the fabrication process is more convenient and faster using the rapid prototyping technique.

This demonstrated 3D Luneburg lens can be applied to several exciting applications
such as direction of arrival estimation and electronically beam scanning by mounting a number of detectors / transmitters around the lens. The details of these applications will be reported in chapters 4 and 5.

1.4 Carbon based nano-material characterization

Carbon nanotube (CNT) that are self-assembling nano-structures constructed of sheets of hexagonally-arranged carbon atoms rolled up into cylinders [70] with a diameter on the order of nano-meters, may offer great potential for next generation high frequency (microwave to THz) devices. A tube consisting of one or more sheets is called a single-walled or multi-walled nanotube (SWNT or MWNT), respectively. Depending on its geometry, a SWNT can be either metallic (with much higher conductivity than conventional conductors such as copper and gold) or semiconducting (with its band gap determined by geometry), which makes it very attractive for next generation ICs to replace silicon ICs since the semiconducting tubes can be used as active devices while the metallic tubes can be used as interconnects and other passives, thus leading to all-carbon ICs. Active SWNT devices have been extensively studied in recent years and a variety of electronic and optoelectronic devices have been demonstrated, including field-effect
transistors (FET) [71], diodes [72], light emitters, detectors [73] and electro-mechanical resonators [74,75]. On the other hand, passive nanotube components are also very important, for they may provide ideal solutions for the difficult interconnecting / interfacing problems within or among nano-scale ICs. For example, according to the International Technology Roadmap for Semiconductors (ITRS), by 2014, the current density required for wiring materials will exceed $10^7$ A/cm$^2$, which is the maximum for copper [76]. With current density exceeding $10^9$ A/cm$^2$ and ballistic transport property, metallic nanotubes could be a promising solution to overcome issues such as reliability, scattering and signal loss in nano-scale interconnecting wires [77]. Due to their high kinetic inductance, metallic nanotubes may also be used as transmission lines with high impedance and very small phase velocity and wavelength (i.e., 100’s times smaller than conventional transmission lines) [78], thus readily matching to intrinsically high impedance active nano devices [79]. Moreover, the smaller wavelength could lead to miniaturized high frequency components such as electrically small antennas [80]. This is particularly suitable for future integrated circuits and systems for which wireless capabilities are highly desirable, while conventional antennas have a footprint of half a wavelength (0.5 mm to 0.5 m for microwave frequencies) and are difficult to be fully integrated with nano-ICs without affecting the achievable circuit density. Furthermore, because of the unique mechanical and electrical properties of nanotubes, a large variety
of sensors may be achievable [81]. It is possible that complete sensing, signal processing and transmitting and receiving functions could be achieved by a single integrated nano-system based on CNT.

Recently, isolated graphene, two-dimensional flat monolayer (or few layer) honeycomb lattice composed of carbon atoms, has been discovered and appears to have some advantages over CNT because of its unique electrical and mechanical properties and its potential to be fabricated macroscopically (~ cm wide graphene sheets) with newly developed techniques [82]. Graphene is also believed to have interesting nonlinear effects such as tunable properties and frequency multiplication in the THz frequency range [83,84,85,86].

Terahertz (THz) research involving the spectrum from 100 GHz to 10 THz has been an exciting forefront of modern technology and experiencing rapid growth in recent years. Rapid developments in nanotechnology in recent years have offered exciting possibilities for revolutionary discoveries in many branches of human endeavor. Nano-materials and associated devices are being widely studied and developed for applications in electronics, optics, biology, energy, etc. Though it has been suggested that nano-materials (such as CNT and graphene) based devices may work well in the THz range [78,87], most of the previous measurements were done at DC, low frequencies and optical frequencies [88,89,90]. The characterization techniques of nano-materials at THz frequency are
important for both fundamental research and practical applications before proposed components such as antennas, interconnections and circuit building blocks [80,91,92] can be realized. In chapter 6 of this dissertation, the characterization method using THz time-domain spectroscopy (TDS) for various carbon nanotubes (CNT) and graphene samples is presented.

1.5 THz time domain spectroscopy (TDS) near field imaging

Terahertz (THz) research involving the spectrum from 100 GHz to 10 THz has been experiencing rapid growth in recent years. The growth is application-driven and involves wide-ranging topics including astronomic and atmospheric spectroscopy and sensing, defense and security screening [93], chemical and biological detection and imaging [94], material research, semiconductor and pharmaceutical industry quality control [95], and next-generation communication networks and radars [96].

Terahertz Time-domain Spectroscopy (THz-TDS) is a very useful tool in various THz applications such as material characterization and identification, biomedical imaging and nondestructive detection. A pair of photoconductive antennas (PCA) is usually served as transmitter and receiver in a TDS system. For a common far-field TDS imaging setup,
the sample object is located in the far field region of the PCA. In this case, the resolution of the system is limited by the diffraction limit of the THz signal. This limit can be overcome by applying a near field imaging setup. With near field setup, a resolution of much smaller than wavelength can be achieved [97].

Most of the previous near-field imaging related works focuses mainly on detection mode [97], where the sample is placed very close to the THz detector. Very limited work is done in the emission mode, and is only done with single emitting antenna [98]. In this work, we apply a PCA array structure (2×2 stripline antenna array) as the THz emitter. With this configuration, many exciting applications can be applied, including Hadamard multiplexing method to improve the SNR [99] and compressive sensing method to decrease the number of measurements [100].

In chapter 7 of this dissertation, a photoconductive antenna (PCA) array used as THz emitter in a THz near field imaging setup is reported. The sample object is placed in the near field region of the antenna array (the antenna-sample distance is about 10 μm). A microlens array is used to couple and focus the fs-laser onto each antenna. By varying the bias state of each PCA element, the ON/OFF state or power level for different PCAs can be controlled independently. The response of the sample with quartz-gold edge is measured for preliminary study on the near field imaging resolution of the antenna. A Hadmard matrix coding is applied to the transmitter array to improve the SNR.
1.6 Compressive sensing

Compressive sensing is a novel sampling/sensing paradigm that enables significant reduction in sampling and computation cost for signals with sparse or compressible representation. It has been experiencing rapid growth in recent years and attracted much attention in electrical engineering, optics, statistics and computer science. The fundamental idea behind compressive sensing is that rather than sampling at high rate first and then compressing the sampled data, it would be much better to directly sample the data in an appropriate compressed format [101]. For example, efficient sampling protocols can be designed to capture small amount useful information of the signal in a sparse domain. After sampling, the full length signal from the small amount of sampled data is reconstructed using numerical optimization algorithm.

In ref [102], compressive sensing technique was applied to a microwave imaging system in which a guided wave metamaterial aperture is used to generate different radiation patterns for compressive sensing. The reconstruction of compressive images at 10 frames per second was achieved at K-band. However, the radiation patterns generated by the metamaterial aperture are basically random and the sampling protocol for this
system is not optimized to capture the signal information. In chapter 8 of this dissertation, a microwave imaging system for human body scanning is investigated. Principal component analysis (PCA) method is used to optimize the measurement radiation patterns for compressive sensing and a reconfigurable array is employed to realize the obtained patterns. Compared to random patterns based compressive sensing system, fewer numbers of measurements is required for this PCA based system to achieve the same performance.

1.7. Dissertation Organization

This dissertation presents the investigation of several additive manufactured components in RF and THz frequency, as well as the applications of gradient index lens based direction of arrival (DOA) estimation system and broadband electronically beam scanning system. Also, a polymer matrix composite method to achieve artificially controlled effective dielectric properties for 3D printing material is studied. Moreover, the characterization of carbon based nano-materials at microwave and THz frequency, photoconductive antenna array based Terahertz time-domain spectroscopy (THz-TDS) near field imaging system, and a compressive sensing based microwave imaging system
are discussed in this dissertation. The dissertation is organized as follows.

In Chapter 2, the theoretical design and experimental measurement of several 3D-printed components in microwave frequency are reported. These components include 3D printed broadband Luneburg lens, 3D printed patch antenna and 3D printed multilayer microstrip line structure with vertical transition. The designs are simulated by full-wave finite-element simulation software HFSS and the results are compared with measurement. The fabrication is implemented using different AM techniques such as polymer jetting rapid prototyping and fused deposition modeling. The operating frequency varies from several GHz to tens of GHz. Good agreement between simulation and experimental results are obtained.

In Chapter 3, several kinds of 3D printed components at THz frequency are reported. These components include THz all-dielectric EMXT waveguide to planar microstrip transition structure and 3D printed dielectric reflectarrays for low-cost high-gain antennas. The fabrication process is combing the polymer jetting rapid prototyping technique and gold plating. Polymer jetting creates the dielectric part of the components and gold plating creates the metallic part of the components. Insertion loss of 6 dB for the designed back-to-back structure is achieved around 110 GHz indicating that this kind of waveguide to planar transition structures will be useful for THz characterization and potential integrated Micro-systems involving small integrated circuits. 3 different types of
3D-printed reflector arrays operating at 100 GHz are characterized and compared to simulation results. Good agreement between simulation and measurement is achieved.

In Chapter 4, a direction of arrival estimation system employing an additive manufactured 3D Luneburg lens is reported. Using the special property of Luneburg lens that every point on the surface of the Lens is the focal point of a plane wave incident from the opposite side, 36 detectors are mounted around the surface of the lens to estimate the direction of arrival of a microwave signal. The system is demonstrated at 5.6 GHz. The direction finding results using a correlation algorithm show that the averaged error is smaller than 1° for all 360 degree incident angles.

In Chapter 5, a novel broadband electronic scanning system based on AM printed Luneburg lens structure is investigated. The radiation elements of the phased array are mounted around the surface of the Luneburg lens. By controlling the phase and amplitude of different transmitting elements, continuous beam scanning with various radiation patterns can be achieved without any mechanical rotation. Compared to conventional phased array systems, this Luneburg lens based phased array structure can operates at a very broadband frequency range and has no scan angle coverage limit. Also, because of the symmetry of Luneburg lens, no beam deformation effect would occur when scanning to different directions. Moreover, this structure requires much less system complexity to achieve a highly directional beam. This reduction in system complexity allows the system
to be built with a much lower cost compare to traditional phased arrays.

In Chapter 6, the characterization of carbon based (Graphene and carbon nanotube) thin films on one side and both sides of substrates via Terahertz time-domain spectroscopy are discussed. The substrate electromagnetic properties are first characterized. The film under test is then treated as a surface boundary condition between the substrate and air. Using the uniform field approximation, the electromagnetic properties of the film sample can be precisely extracted. To improve accuracy, precise thickness of sample substrate is calculated through an iteration process in both dielectric constant extraction and surface conductivity extraction. Uncertainty analysis of the measured thin film properties is performed. The SWNT results show consistent surface conductivities for samples on different substrates and with different film thicknesses. The measured graphene Terahertz conductivity is comparable to the values reported in literatures at DC and optical frequency. This characterization method has been successfully applied as a means to evaluate metallic content of SWNT samples to verify a metallic SWNT removing process using high power microwave irradiation.

In Chapter 7, a coded transmitter TDS near field imaging system by employing photoconductive antenna (PCA) array is discussed. Silicon lens array is used to couple and focus the femto-second laser into each PCA. By varying the bias state of each PCA element, the ON/OFF state or power level for different PCAs can be controlled
independently. The sample object is placed 10 μm away from the PCA array to measure the THz near field image. A Hadamard matrix is applied to code the 2x2 antenna array to improve the SNR. Measured results clearly indicate an improved SNR compared to independent antenna measurement. In addition, Multiphysics COMSOL and a FDTD algorithm combined with HFSS time domain simulation are used to model the physics of TDS photoconductive antenna, optimize the performance of transmitter and predict the near field scanning results. Good agreement between simulation and experiment is obtained.

In Chapter 8, a design of a Principal Component Analysis (PCA) based microwave compressive sensing system using reconfigurable array is reported. The required radiation patterns obtained from PCA is realized by employing an iterative beam synthesis process. A human body scanning system is studied as an example to investigate the compressive sensing performance using PCA generated radiation patterns. Optical images are used as surrogates for the RF images in implementation of the PCA training dictionary. Compared to conventional microwave imaging system or random patterns based compressive sensing system, this PCA based compressive sensing system requires fewer numbers of measurements to achieve the same performance.

Finally, Chapter 9 presents a summary of achievements and contributions in this dissertation.
CHAPTER 2 3D PRINTED COMPONENTS IN MICROWAVE FREQUENCY

2.1. 3D printed broadband Luneburg lens

2.1.1 Introduction

Luneburg lens is an attractive gradient index component for wide angle radiation scanning because of its broadband behavior, high gain and the ability to form multiple beams. In this paper, a 3-D printable Luneburg lens antenna is designed, printed and characterized for X-band (8.2 – 12.4 GHz) operation. The desired gradient index (or, permittivity due to the relative permeability being 1) is realized by controlling the filling ratio of a polymer / air based unit cell. Efficient and accurate fabrication of the designed 3-D lens is enabled by a polymer-jetting rapid prototyping technique developed previously [13]. The effective permittivity of each unit cell is designed independently based on its distance to the center of the sphere. Since the refractive index of the lens is independent with frequency as long as the long wavelength condition holds (to guarantee the effective medium approximation), it could operate in a quite wide frequency range. For example, although designed for X-band operation, this lens should work at lower frequencies with reduced gain and at higher frequencies until the effective medium approximation breaks down (roughly 20 GHz for the prototype in this work). The
fabricated Luneburg lens antenna is tested using an X-band WR-90 waveguide as the feed. Antenna properties such as radiation patterns and efficiency as a function of frequency and feed location have been studied. The measurement results show that the half-power beam width of the $4\lambda_0$ (12 cm) diameter lens is 15 degrees at 10 GHz and the gain of the antenna is 18.5 dB. The measured results agree very well with design simulations. The demonstrated 3-D Luneburg lens antenna can be used to easily realize 3-D switched beams and can be very useful for a number of communication and sensing applications. We have realized several very interesting applications such as direction of arrival estimation and low-cost electronic beam scanning utilizing this kind of 3-D lens antenna. Moreover, the polymer jetting rapid prototyping technique used in this work is quite convenient, fast and inexpensive, and can be applied to realize other interesting gradient index and metamaterial based components.

The chapter is organized as the following. Section 2.1 reviews the polymer jetting rapid prototyping technology in printing the lens. Then, the Luneburg lens design procedure is discussed in section 2.2. Next, numerical simulation studies of the Luneburg lens are reported in section 2.3. The experimental setup and measured antenna properties are then presented in section 2.4 before the final conclusions are given in section 2.5.

2.1.2 Polymer Jetting Rapid Prototyping

The polymer jetting rapid prototyping technique used here allows fast printing of
polymer components with arbitrary shapes and complexity [13,103,104]. A commercial rapid prototyping machine Objet Eden 350 is employed to print the Luneburg lens. The printing process of the prototyping machine is as follows. First, the CAD file of a designed 3-D object is converted into a series of layered slices, with each slice representing a 16-μm thick region of the designed model. The data describing the slices are then sent to the prototyping machine one by one. Once the data for each slice is received by the prototyping machine, a series of print heads, just like the print head of an ink-jet printer, deposits a thin layer of polymer made of two different ultraviolet-curable materials onto the construction stage. The model material regions of the slice receive an uncured acrylic polymer while the support material regions receive an uncured water-soluble polymer. Upon jetting, both of the materials are immediately cured by the ultraviolet lamps on the print head. After one layer is completed, the construction stage is lowered by 16 μm, and then the next layer is printed on top. When the entire structure is printed, the water-soluble support material of the structure is washed away using a high pressure water spray, leaving just the model material in the designed region. For the convenience of washing away the support material in the center of the Luneburg lens, the lens is divided into 22 different layers and each layer was printed separately. After washing away the support material of each layer, they are assembled together very easily using the predefined alignment holes and rods.
The Objet Eden 350 polymer jetting printer states a droplet size of 84 μm x 42 μm x 16 μm, which is more than sufficient for fabricating the X-band Luneburg lens. Moreover, large structures with a size of up to 30 cm x 30 cm x 30 cm could be printed, so that low volume batch printing of a large number of components is possible. With this polymer jetting technique, the printing process is relatively fast, convenient and inexpensive. The total printing time for the 12-cm diameter lens is less than 4 hours and the material cost is less than $200.

2.1.3 Luneburg Lens Design

Figure 2-1. The front view of the Luneburg Lens design. The left figure is the discrete polymer cubes with different sizes to control the dielectric constant distribution of the lens. The right figure includes the thin rods used to support the whole structure and connect all the cubes together. The inset at the center is a schematic of the cubic unit-cell which has an
The designed Luneburg lens is composed of discrete polymer cubes with different size $b$ as shown in Figure 2-1. The cubic unit cell of the lens is shown as the center inset with an overall size of 5 mm and a dielectric cube with a variable dimension $b$. The whole structure is mechanically supported by thin rods that go through each of the unit cell and connect all the discrete cubes together. The front views with (right in Figure 2-1) and without (left in Figure 2-1) the thin connecting rods of the designed Luneburg lens are shown in Figure 2-1. The dimension of these connecting rods is very small (diameter of 0.8 mm) compared to the unit cell size so that they have negligible impact on the EM properties of the Luneburg lens structure. The gradually variable dielectric constant is controlled by the polymer cube size in each unit cell. When the polymer cube size is larger, the filling ratio between polymer and air void is larger, leading to greater effective permittivity of the unit cell. To achieve the permittivity distribution given by Equation (1-1), the polymer cube size (or the filling ratio) should be maximized at the center of the lens, and gradually decreasing with the radius and finally reaches zero at the surface of the lens. The relative permittivity of the polymer used in the prototyping printer is measured to be 2.7. Since the desired variable relative permittivity in Equation (1-1) is from 2 to 1, it is easy to realize the desired permittivity distribution by changing the
The diameter of the Luneburg lens is designed to be 12 cm, which is $4\lambda_0$ ($\lambda_0$ is the wavelength in free space) at the frequency of 10 GHz. The discrete unit cell size is 5 x 5 x 5 mm$^3$, which is $1/6$ of $\lambda_0$ at 10 GHz. Each unit cell is essentially a polymer cube in the center with air voids around it. Because the effective permittivity of the unit cell is not perfectly linear with the filling ratio as the simple effective medium theory would predict, we cannot just use the approximated average permittivity from the filling ratio as the effective permittivity to design the Luneburg lens. In order to determine a more accurate relationship between the effective permittivity and the polymer cube size (or, the filling ratio) of a unit cell, finite-element simulation software ANSYS HFSS is applied to obtain the effective permittivity of the unit cell and optimize the design.

The effective permittivity simulation setup in HFSS is shown in Figure 2-2. A polymer cube with its supporting rods is placed in a waveguide. PEC and PMC

<table>
<thead>
<tr>
<th>Cube size (mm)</th>
<th>0.5</th>
<th>2</th>
<th>2.5</th>
<th>3</th>
<th>3.5</th>
<th>4</th>
<th>4.25</th>
<th>4.5</th>
<th>4.75</th>
<th>5</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\varepsilon_r$ (simulation)</td>
<td>1.013</td>
<td>1.103</td>
<td>1.1735</td>
<td>1.274</td>
<td>1.473</td>
<td>1.752</td>
<td>1.9</td>
<td>2.1</td>
<td>2.37</td>
<td>2.7</td>
</tr>
<tr>
<td>Tan $\delta$ (simulation)</td>
<td>0.002</td>
<td>0.0027</td>
<td>0.004</td>
<td>0.006</td>
<td>0.0087</td>
<td>0.0117</td>
<td>0.0133</td>
<td>0.0147</td>
<td>0.0174</td>
<td>0.02</td>
</tr>
<tr>
<td>Filling ratio</td>
<td>0.001</td>
<td>0.0588</td>
<td>0.1148</td>
<td>0.1984</td>
<td>0.343</td>
<td>0.512</td>
<td>0.6141</td>
<td>0.729</td>
<td>0.8574</td>
<td>1</td>
</tr>
<tr>
<td>$\varepsilon_r$ (filling ratio)</td>
<td>1.0017</td>
<td>1.1</td>
<td>1.1952</td>
<td>1.3372</td>
<td>1.5831</td>
<td>1.8704</td>
<td>2.044</td>
<td>2.239</td>
<td>2.4575</td>
<td>2.7</td>
</tr>
</tbody>
</table>
boundaries are used to set up the periodic structure. From the $S$-parameter simulation of the waveguide and applying the standard retrieval method [105], the effective permittivity of the unit cell can be extracted for different polymer cube size from 0 mm to 5 mm. The extracted results are shown in Table 2-1.

![Diagram of effective permittivity simulation setup](image)

Figure 2-2. The effective permittivity simulation setup, in which $h$ is the thickness of the unit cell slice, $a$ is the length of the unit cell, and $b$ is the size of polymer cube.

Also, the approximated average permittivity is calculated from the filling ratio using Equation (2-1):

$$\varepsilon_r = \varepsilon_p \cdot (f) + \varepsilon_0 \cdot (1 - f)$$

in which $\varepsilon_p$ is the permittivity of the polymer material and $f$ is the polymer filling ratio of
the unit cell. It can be seen in Table 2-1 that the effective permittivity extracted from the
HFSS simulation and calculated from the filling ratio using Equation (2-1) agrees well
when the polymer cube size is less than 2 mm, but differs for larger cube sizes. To
determine the required cube size for the desired permittivity in Equation (1-1), an
exponential fitting was applied to the extracted results. As shown in Figure 2-3, the black
curve is the effective permittivity extracted from the S parameter results of HFSS
simulation; the red curve is the exponential fitting of the extracted results; and the blue
curve is the approximated average permittivity calculated by Equation (2-1). After this
curve fitting, the spatial polymer cube size distribution is then obtained from the desired
permittivity using the fitted exponential function (the red curve in Figure 2-3):

\[ b = 5.5593 - 590974 \varepsilon_r^{-0.07958} - 9.54823 \varepsilon_r^{0.95537} \]  

(2-2),
in which \( b \) is the polymer cube size and \( \varepsilon_r \) is the desired permittivity in Equation (1-1).
2.1.4 Simulation Results of The Designed Luneburg lens

To understand the design parameters and evaluate the performance of the Luneburg lens, the entire lens structure is simulated using the HFSS software. The schematic of a 120 mm diameter lens is shown in Figure 2-4. It is composed of 7497 unit cells of polymer cubes with different sizes. At the center of the spherical lens, the polymer cube size is the largest (4.38 mm) and at the surface of the lens, the polymer cube size decreases to zero. In the simulation, the dielectric constant of the model region (polymer cubes) is set to 2.7, and the loss tangent is set to 0.02, according to previously measured
material properties. The feeding source is a rectangular waveguide – WR-90/WR-62 (for different frequency range) placed on the surface of the Luneburg lens, which is the same as the experimental configuration that will be discussed later. Figure 2-5 plots the simulated radiation patterns in the H-plane (XY plane) at different frequencies in the X-band. Figure 2-6 plots the simulated radiation patterns in the H-plane at Ku-band. It can be seen that the Luneburg lens works as a narrow beam antenna in a broad frequency band as predicted.

Figure 2-4. Simulation setup of the Luneburg Lens in HFSS with a rectangular waveguide as the excitation.
Figure 2- 5. H-plane radiation patterns from HFSS™; excitation was a WR-90 flange as per Figure 2-4.

Figure 2- 6. H-plane radiation patterns from HFSS™; excitation was a WR-62 open waveguide.
Figure 2-7. Simulated antenna directivity and gain (left) and HPBW (right) in the H-plane at frequencies from 8.2 GHz to 12.4 GHz with WR-90 waveguide as excitation.

Figure 2-8. Simulated antenna directivity and gain (left) and HPBW (right) in the H-plane at frequencies from 10 GHz to 20 GHz with WR-62 waveguide as excitation.

The simulated antenna gain and directivity versus frequency are shown in Figure 2-7 (left) and Figure 2-8 (left) for X-band waveguide feed and Ku-band waveguide feed, respectively. Both the gain and directivity increases with frequency due to the increase of
the effective aperture size, as expected. At 8.2 GHz, the simulated directivity and gain of the 12-cm diameter Luneburg lens antenna are 18.3 dB and 17.7 dB, respectively. At 20 GHz, they are 25.76 dB and 24.25 dB, respectively. The H-plane half-power beam width (HPBW) of the lens antenna is from 7.88 degrees at 20 GHz to 19 degrees at 8.2 GHz (Figs. 2-7 and 2-8, right). To find out the reason for the increased loss at higher frequencies, the antenna gain with 0 loss tan material is also simulated. The results show that the antenna loss still increases at high frequencies. Therefore, the increased loss of the lens at higher frequencies is not only due to the material loss but also due to the finite size of the polymer cubes.

To evaluate the Luneburg lens antenna size effect, lenses with different diameters ranging from 45 mm to 125 mm are simulated. In these simulations, the excitation of the lens antenna is an ideal lumped-gap source placed on the surface of the lens. The lumped gap has a dimension of is 0.4 mm x 0.4 mm.

Figure 2-9(a) illustrates the simulation results of the far field radiation patterns in the H–plane with lumped port feed. Different colors in the figure represent the antenna patterns with different lens size. The simulation is analyzed at 10 GHz. Figure 2-9(b) compares the radiation pattern of a 125 mm lens with waveguide feed and lumped port feed. It can be seen that the maximum gain value has about 3 dB differences which is due to the reason that the lumped port feed radiates to all 360 degrees and therefore half of
the power will not be captured by the lens. The simulated antenna gain and HPBW versus lens diameters ranging from 45 mm to 125 mm with lumped port feed are plotted in Figure 2-10. The gain of the Luneburg lens antenna increases with the increasing of the lens diameter and the HPBW of the antenna decreases with the increasing of the diameter, as expected. When the diameter is at 45 mm, the antenna gain is 8.9 dB and the HPBW is 38 degrees. When the lens diameter increases to 125 mm, the antenna gain increases to 17.3 dB and the HPBW decreases to 13 degrees.
Figure 2-9. (a) Simulated far field H-plane gain patterns of the Luneburg lens antenna for different lens diameter with a lumped port feed at 10 GHz. (b) Simulated radiation pattern of a 125 mm lens with a waveguide feed and Lumped port feed at 10 GHz.

Figure 2-10. Simulated gain and HPBW in the H plane with different lens diameters at 10 GHz. The feeding source is a lumped gap port on the surface of the Luneburg lens.
2.1.5 Fabrication And Experiment

A Luneburg lens with the same size as the simulation is fabricated using the polymer jetting rapid prototyping technique as described previously in section 2.1.2. It is composed of 7497 discrete polymer cubes with different sizes. All the cubes are mechanically connected and supported by thin posts that go through each of the cubes in the X, Y and Z direction. The cross section dimension of the post are 1 mm x 1mm. Figure 2-11(a) is a photograph showing the cross-section cut through the center of the lens and Figure 2-11(b) is a photograph of the entire lens structure. It can be seen from Figure 2-11(a) that the cube size is the largest at the center and gradually decreases to zero at the surface of the lens to achieve the required dielectric constant distribution in Equation (1.1). The posts outside the Luneburg lens in the X, Y and Z direction (thus the lens looks cubical instead of spherical) are just for the convenience to fix the lens antenna on the measurement stage. They do not significantly influence the electromagnetic properties of the lens antenna since the post volumes are very small compared to the unit cell size of the designed Luneburg lens such that the effective relative permeability in that region is still very close to 1 (dielectric constant of the free space). As shown in Figure 2-3, when the cube size is at 0.5 mm, the effective relative permittivity of the unit cell is just 1.013. In chapter 4, we also printed a spherical shape 24 cm diameter lens without the outside posts in order to have enough space to mount a number of detectors around
the lens. The performance of that spherical lens is quite similar to the performance of this cubic lens except higher gain due to the larger diameter which verified that the outside posts do not influence the electromagnetic properties of the lens.

![Image of the Luneburg lens](image1)

Figure 2-11. Photographs of the fabricated Luneburg lens: (a) the cross-section cut through the center of the lens; (b) the entire lens.

![Image of experiment setup](image2)

Figure 2-12. Experiment setup with the Luneburg lens fed by a X-band waveguide.
(WR-90) mounted on the surface of the Luneburg lens.

![Graph](image)

Figure 2-13. Simulated results of the radiation gain patterns when an X-band feeding waveguide is at some distance (from 0 mm to 30 mm) away from the lens surface.

In the experiment, the Luneburg lens antenna is fed by an X-band / Ku-band waveguide mounted on the surface of the lens as shown in Figure 2-12. To evaluate the tolerance of the feeding position, the far field pattern of the Luneburg lens is simulated with an X-band waveguide placed at various distance (0 to 30 mm) from the lens surface. The result is shown in Figure 2-13 in which different colors represent the far field patterns with waveguide feeding at different distance away from the lens surface. It can be seen that the gain of antenna is almost the same when the waveguide is within a 10 mm distance away from the lens surface. As the waveguide feed moves a relatively larger
distance away from the lens surface (i.e., 20 mm and 30 mm), the gain of antenna begins to decrease up to 3 dB.

The lens antenna radiation patterns at X-band are measured using a vector network analyzer (HP 8720) in an anechoic chamber. A standard gain horn antenna at X-band is used to calibrate the Luneburg lens antenna gain. The measured H-plane gain pattern of the antenna at 10 GHz is show in Figure 2-14, together with HFSS simulation results. The measured gain of the Luneburg lens is 18.7 dB and the HPBW is 15 degrees at 10 GHz, agreeing well with the simulation results. The measured side lobe is about 25 dB lower than the main peak. The agreement between the simulation and measurement results are reasonable.

In Figure 2-15, the measured radiation gain patterns in the H-plane at different frequencies from 8.2 GHz to 12.4 GHz are plotted. Directional beams around 0 degree can be seen for all frequencies, indicating that this Luneburg lens works well as a directional antenna in a broad frequency band.

The measured and simulated antenna H-plane gains versus frequency from 8.2 GHz to 12.4 GHz are compared in Figure 2-16(a) together with lines for different aperture efficiencies. The gain of the Luneburg lens antenna increases with the increase of frequency as expected. The simulated gain ranges from 17.8 dB at 8.2 GHz to 21.4 dB at 12.4 GHz, while the measured gain ranges from 17.38 dB at 8.2 GHz to 20.8 dB at 12.4
GHz. The measured and simulated H-plane HPBW results are illustrated in Figure 2-17(a). Also, the HPBW decreases with the increase of frequency as expected, ranging from 19 degrees at 8.2 GHz to 12.7 degrees at 12.4 GHz. The somewhat smaller gain (0.5 dB at 8.2 GHz and 1.4 dB at 12.4 GHz) of the measured data is probably caused by the polymer material property variation and printing tolerances.

The measured E-plane gain and HPBW value versus frequency are shown in Figure 2-16(b) and Figure 2-17(b). Similar to the H-plane results, the gain of the Luneburg lens increases with the increase of frequency, and the HPBW decreases with the increase of frequency. Measured E-plane radiation patterns of the antenna at different frequencies in X-band are plotted in Figure 2-18. One can see that the sidelobe levels are about 20 dB lower than the main peak.
at 10 GHz.

Figure 2-15. Measured H-plane radiation gain pattern of the Luneburg lens antenna from 8.2 GHz to 12.4 GHz.

Figure 2-16. Measured and simulated H-plane (a) and E-plane (b) gain versus frequencies
from 8.2 GHz to 12.4 GHz. The dotted lines are the lines for different aperture efficiency from 40% to 60%.

Figure 2-17. Measured H-plane (a) and E-plane (b) HPBW at different frequencies in X-band.
The measured E-plane radiation gain pattern of the Luneburg lens antenna from 8.2 GHz to 12.4 GHz is shown in Figure 2-18. The radiation patterns with a Ku-band waveguide feed are also measured. Due to a longer far field distance, the measurement is done outside the anechoic chamber. A standard gain horn antenna at Ku-band is used to calibrate the Luneburg lens antenna gain. The measured H-plane and E-plane radiation patterns at Ku-band are shown in Figure 2-19 and Figure 2-20, and the HPBW values are plotted in Figure 2-21. The measured gain value at 19.8 GHz is about 24 dB and the HPBW is 7 degree and 6 degree for H-plane and E-plane, respectively. Directional beams around 0 degree can be seen in both H-plane and E-plane for all frequencies up to 19.8 GHz, indicating that this Luneburg lens with 5 mm unit cell size works as a directional antenna.
in a broad frequency band at least from 8 to about 20 GHz.

Figure 2-19. Measured H-plane radiation patterns at different frequencies in Ku-band.

Figure 2-20. Measured E-plane gain radiation patterns at different frequencies in Ku-band.
2.1.6. Conclusion

In section 2.1, the design and fabrication of a spherical lens with Luneburg index distribution is proposed and demonstrated. The lens is printed using a polymer jetting rapid prototyping technique. The diameter of the lens is 12 cm (4\(\lambda_0\) at 10 GHz), with a unit cell size of 5 mm. Good agreement between experiments and simulation is achieved. Measurement results show that the gain of this lens antenna is from 17.3 dB (at 8.2 GHz) to 24 dB (at 19.8 GHz) in the X-band and Ku-band. The H-plane half-power beam width is from 19 degrees (at 8.2 GHz) to 7 degrees (at 19.8 GHz). E-plane HWPBW is from 14.3 degrees (at 8.2 GHz) to 6 degrees (at 19.8 GHz). The side lobe is measured to be about 25 dB lower than the main beam for H-plane and about 20 dB lower than the main beam for E-plane. Compared to traditional Luneburg lens fabrication techniques, this 3-D
Luneburg lens can be printed with lower cost and the printing process is more convenient and faster using the rapid prototyping technique. For the present prototype lens, although it is relative small, it may still be used in radar reflector applications [106] to give an increased radar signature. With a larger lens size, this type of 3-D Luneburg lens antennas can be easily used to realize 3-D switched beams and could be very useful for a number of communication and sensing applications. In Chapter 4, some interesting applications such as direction of arrival estimation [107] and low-cost electronic beam scanning based on this Luneburg lens [108] are reported.

2.2. 3D printed microwave patch antenna via fused deposition method and ultrasonic wire mesh embedding technique

2.2.1. Introduction

Additive manufacturing (AM), often called 3D printing creates products layer by layer additively rather than conventional manufacturing technique by removing parts from a larger piece of material. It has received much attention recently with impressive demonstrations ranging from musical instruments [109], to vehicles [110], to housing components [111] or even entire buildings [112]. Different material such as polymer [113], metal [114], ceramics [115], concrete [116] and even biological tissues [117] have been printed by various 3D printing technologies. Although it has been argued that 3D
printing could be the future of manufacturing, the potential and applicability of these methods for creating functional antennas at RF / microwave frequency have yet to be thoroughly explored.

The major advantage of using 3D printing technologies to fabricate microwave antennas include rapid realization of designs without going through conventional processes such as machining and photolithography, ease of realizing complex geometry such as 3D conformal shapes, special tailored dielectric properties such as gradient index structures, etc. In previous work, an electrically small antenna fabricated by conformal printing of conductive ink on 3D surfaces has been demonstrated at 1.7 GHz [20]. However, in order to achieve a high conductivity comparable to regular metal, the conductive ink used in [20] was heated to a high temperature of 550 ºC for annealing process. Moreover, 3D printed antennas with dielectric and metal together have also been demonstrated. For example, in [118], a meander line antenna working at 1.1 GHz was fabricated by printing conductive ink on a 3D printed polymer substrate, the conductive ink is cured at 85 ºC to improve the conductivity, but without high temperature sintering process, the conductivity of the conductive ink can achieve about only one-tenth of pure metal. This lower conductivity of the radiation part would increase the conductive loss of antenna and decrease the antenna efficiency.

In this section, a microstrip patch antenna operating around 7.5 GHz manufactured
entirely by additive manufacturing techniques is demonstrated. The conductive part of the antenna (i.e., the microstrip patch and the ground plane) is realized using an ultrasonic embedded wire mesh structure which works as good as regular metal sheet at microwave frequency and avoids the commonly required annealing process at high temperature. The dielectric part of the antenna (i.e., the antenna substrate) is printed by the fused deposition modeling (FDM) method [119]. A seamless integration procedure of these two techniques has been developed which allows robust and flexible 3D printing of passive microwave components and potential microwave systems. Compared to other 3D printed antenna using conductive ink [118, 120], this method achieves satisfactory high frequency performance while avoiding the high temperature metal sintering process which may induce deformation or damage of the dielectric substrate and prevent potential integration of active semiconductor devices. Moreover, compared to standard PCB technique that metal can be only fabricated on a planar surface, this embedded wire mesh method can be applied to any surfaces including curved ones and enable 3D conducting structures. Therefore, the presented 3D printing of both dielectric and conductor constituents may lead to advanced and high performance microwave structures and systems which are challenging to fabricate using conventional techniques.
2.2.2 Microstrip Transmission Line and Antenna Design

Our previous work has shown that the implementation of conducting traces using embedded wire works very well for DC and low frequency interconnects [121]. To evaluate the proposed 3D printing of microwave components consist of dielectric and conductor, microstrip transmission lines are studied since they are one of the most representative building blocks of microwave structures. A 50-Ω microstrip line structure on a 2.4 mm thick thermoplastic substrate (i.e., Polyethylene) with a dielectric constant $\varepsilon$ of 2.4 is designed and simulated using full-wave finite element simulations (Ansys HFSS [122]). As shown in Figure 2-22, both the top conducting trace and the ground plane of the microstrip is made of embedded metal wire mesh. The width of the microstrip is designed to be 8 mm to obtain 50-Ω characteristic impedance. The length of the microstrip line is 50 mm. The wires used in the wire mesh have a diameter of 0.5 mm.

![Figure 2-22. Full-wave finite element EM model (HFSS) of a microstrip transmission line](image)
implemented with wire mesh embedded in thermoplastic substrate.

Figure 2-23. Simulated transmission coefficient ($S_{21}$) of the microstrip line using wire mesh structures with different wire spacing compared with microstrip line made of regular conductor.

The transmission responses of this microstrip with different wire mesh spacing ranging from 1 mm to 4 mm are simulated to determine the optimum wire mesh configuration. Intuitively, smaller wire mesh spacing emulates conventional continuous conducting surface closer at the expense of longer printing time and more material consumptions. The simulated transmission results are plotted in Figure 2-23 together with that of the same microstrip made of copper, commonly used in integrated circuits and printed circuit boards. It can be observed that the 1 mm spacing wire mesh based microstrip transmission line works very well without performance degradation compared to the ideal conventional microstrip in the entire frequency range considered from 2 to 20
GHz. Even for the 2 mm wire spacing case, only slightly more transmission loss (less than 0.2 dB higher) is seen which should be acceptable for most applications. With a wire spacing of 4 mm, the wire mesh based microstrip transmission line is significantly worse than the other 3 cases. For example, after 18 GHz, the insertion loss of the microstrip line is larger than 1 dB/cm. From these simulation results, one can see that high performance microwave transmission line can be realized by the wire embedding process with the wire spacing smaller than 2 mm. This conclusion is very encouraging as there are a large number of planar microwave components such as antennas, filters, power dividers, couplers, etc., can be implemented by microstrip transmission lines. Furthermore, the wire embedding technique combined with 3D printed dielectrics can easily achieve vertical connections such as through substrate vias, hence realizing multi-stage vertically integrated circuits.

Figure 2-24. Schematic of the microwave patch antenna made of embedded wire mesh.
Based on the simulated microstrip line results, a wire mesh based microwave patch antenna is designed. Patch antenna can be viewed as a half wavelength microstrip transmission line resonator. For more performance margin, the 1 mm wire spacing is chosen to make sure the wire mesh works as close as normal conductor sheet. The schematic picture of the microwave patch antenna using the embedded wire mesh technique is shown in Figure 2-24. Both bottom ground plane and top patch radiator are made of wire mesh. The diameter of the wire chosen here is 250 \( \mu \text{m} \). The substrate of the antenna is Acrylonitrile butadiene styrene (ABS) which is one of the most common materials used for FDM method. The dielectric constant of the ABS is 2.7 and loss tangent is 0.01 [123]. The dimension of the substrate is 21 mm x 15 mm x 3.2 mm. The size of the top patch radiator is 11 mm x 9 mm. The patch antenna is probe-fed [124] through the ABS substrate using a coaxial SMA connector from the bottom of the antenna.
Figure 2-25. Simulated (a) reflection coefficient and (b) co and cross-polarization radiation patterns of the wire mesh based patch antenna in H-plane compared with an ideal conductor patch antenna.

Figure 2-25(a) plots the HFSS simulated reflection coefficient of this antenna designed with a center frequency of about 7.5 GHz. Figure 2-25(b) shows the simulated co- and cross-polarization radiation patterns of this wire mesh patch antenna at the resonance frequency. For comparison, the simulated reflection coefficient and radiation pattern of a patch antenna made of regular metal sheet with the same geometry are also plotted in the same figure. One can see that the simulated performance of the wire mesh based antenna agrees very well with the antenna made of regular metal, indicating that the wire mesh structure with 1 mm spacing has good performance just like regular metal
for the designed microwave frequency range.

Figure 2-26. Simulated antenna directivity, gain and realized gain of the wire mesh patch antenna at different frequencies from 7 GHz to 10 GHz.

The simulated antenna directivity, gain and realized gain of the wire mesh patch antenna versus frequency are shown in Figure 2-26. At the resonance peak of 7.6 GHz, the simulated directivity and gain of the antenna is 6.24 dB and 5.38 dB, respectively, corresponding to an antenna radiation efficiency of 84%. With a substrate made of thermoplastics with a lower loss tangent than 0.01 (e.g., polycarbonate with loss tangent of 0.005 or graft polymer with loss tangent of 0.001), higher antenna gain and radiation efficiency can be obtained.

2.2.3 3D Printing of the Designed Patch Antenna

The 3D printed patch antenna was created using a Multi3D Manufacturing System.
The enhanced manufacturing technology utilizes ultrasonic and thermal embedding for submerging wire and wire meshes into 3D printed thermoplastics. With wire meshes that can be submerged into the conformal, geometrically complex thermoplastic surface during fabrication, this technique can enable novel, high performance, volume-efficient RF/microwave applications.

3D printing by its nature is a non-homogeneous process. Due to the ultrasonic embedding process, as well as the small physical size of the patch antenna, the material fill percentage is predicted to be greater than 95%. Also, some typical dimensional tolerances for common 3D printers are described in more detail in [125]. This 3D printed patch antenna was designed for printing in SolidWorks® and then transferred to an appropriate STL slicing suite for printing. The thermoplastic base was then printed in a Stratasys FDM3000 with ABS thermoplastic produced by Stratasys. The Stratasys FDM3000 was outfitted with T16 printing tips calibrated to produce a printed raster of 254 μm and utilized a print temperature of 270°C and a build envelope of 70°C. The accuracy of the FDM technique is reported to be less than 0.13 mm [42]. The final design was optimized so that the thickness between the two mesh planes was 3.2 mm. Therefore, the initial ABS dielectric was constructed to be 3.556 mm thick to allow full embedding of the copper mesh. Both the top plane and ground plane copper meshes (595 μm spacing, 305 μm wire dia.) in this work were embedded ultrasonically using a custom
gantry mounted large area ultrasonic horn powered by a Cole Palmer Ultrasonic Processor.

2.2.4 Antenna Testing and Results

Figure 2-27. The photo of a 3D printed microwave patch antenna using ultrasonic wire embedding.

Figure 2-27 shows the picture of a printed antenna sample. The feeding is using a SMA connector at the back of the antenna. The SMA center conductor is inserted into the substrate and soldered to the patch antenna top radiator. The antenna S-parameter and radiation patterns are measured using a vector network analyzer (Agilent E8361A).

The measured reflection coefficient of the antenna is shown in Figure 2-28, together with the simulation results. A clear resonance peak at 7.5 GHz can be observed in the measured data which agrees very well with the simulation predicated results. The
measured reflection coefficient at the resonance frequency is -14 dB.

Figure 2-28. Comparison of measured and simulated reflection coefficient of the printed wire-mesh antenna.

Figure 2-29. Measured radiation pattern of the printed patch antenna compared to
simulation results.

The measured H-plane (YZ plane in Figure 2-24) radiation pattern of the antenna at 7.5 GHz is plotted in Figure 2-29, together with the HFSS simulated results for wire mesh and ideal conductor based patch antenna. The measured realized gain of the patch antenna is 5.5 dB and the half-power beam width (HPBW) is 114-degree. Again, the agreement between measurement data and simulation results are very good.

![Figure 2- 30. Measured and simulated broadside realized gain of the 3D printed patch antenna at different frequencies.](image)

In Figure 2-30, the measured and simulated antenna realized gains at broadside versus frequency from 7 GHz to 9 GHz are compared. The measured antenna realized gain of the printed patch has a maximum value of 5.5 dB at 7.5 GHz and ranges from
5.12 dB at 7 GHz to 1.28 dB at 9 GHz. The agreement between the simulation and measurement results are reasonable. The small discrepancy away from the resonance frequency between the simulation and measurement is probably caused by the polymer material property variation and printing tolerance of the embedded metal wire. In summary, these simulated and measured results have confirmed that the wire mesh process works well for the implementation of microwave patch antennas.

2.2.5 Conclusion

This section demonstrates 3D printing of microwave patch antenna by combining fused deposition modeling method with ultrasonic metal wire mesh embedding. No metal sintering or any other high temperature conductor printing process is needed and the printed metal wire mesh working as well as regular metal sheet is proved in both simulation and measurement. Measurement results show that the gain of this patch antenna is 5.5 dB at the resonance peak. Good agreement between experiment and simulation is achieved in both reflection coefficient and radiation pattern. Compared to traditional patch antenna fabrication method, this 3D printed antenna can be fabricated with lower cost and the fabrication process is more convenient and faster. This demonstrated 3D printing process of both dielectric and conductor can be applied to the 3D printing of more sophisticated EM structures for microwave applications including
both conventional components and new types of 3D systems such as vertically integrated phased arrays that are difficult to fabricate using conventional techniques.

2.3. 3D Printed Multilayer Microstrip Line Structure with Vertical Transition toward Integrated Systems

2.3.1 Introduction

A multilayer microstrip transmission line structure with vertical transition printed entirely by additive manufacturing techniques is demonstrated. The conductive part of the structure is realized using the ultrasonic embedded wire mesh structure which works as good as regular metal at microwave frequency to avoid the commonly required high temperature annealing process. The dielectric part of the structure is printed using the fused deposition modeling (FDM) approach [119]. A seamless integration procedure of these two techniques has been developed which allows robust and flexible additive manufacturing of microwave components and potential microwave systems. This 3D printed multilayer microstrip structure demonstrates that the additive manufacturing approach by integrating the novel wire mesh embedding technique and FDM is a good solution to achieve low cost, fast and convenient fabrication of RF components.
Compared to other 3D printed components using conductive ink, this method achieves satisfactory high frequency performance while avoiding the high temperature metal sintering process which may induce deformation or damage of the dielectric substrate and prevent potential integration of active semiconductor devices. Moreover, this multilayer microstrip transmission line structure with vertical interconnections demonstrates the possibility to realize additive manufacturing of compact RF system fully utilizing the entire 3D space. As an example, based on the similar transmission line and vertical transition structure, a 4-element 3D printable multilayer phased array is designed and simulated. The simulated results are also presented.

2.3.2 Design and Simulation

It has been demonstrated that the implementation of conducting traces using embedded wire works very well for DC and low frequency [121]. At microwave frequency, wire mesh based structures have also been successfully used to replace the solid metal sheet for antenna and electrode applications [19,126]. It can be shown that with a 1-mm wire spacing, the wire mesh based microstrip transmission line works very well without performance degradation compared to the ideal conventional microstrip made of solid metal sheet up to 20 GHz.

Based on those results, a wire mesh based multilayer microstrip line structure with
vertical transition is designed. For more performance margin, 0.5-mm wire spacing is chosen to make sure the wire mesh works as close as normal conductor sheet. The schematic picture of the designed multilayer microstrip line structure is shown in Figure 2-31. The substrate is printed using Acrylonitrile butadiene styrene (ABS) which is one of the most common thermoplastic materials used in FDM. The dielectric constant and loss tangent of the ABS are 2.7 and 0.01 at 10 GHz [123]. The printing accuracy for Makerbot FDM printer in the z-direction is about 0.4 mm, and therefore, the dielectric substrate should be thick enough so that any fabrication error will not influence the line impedance too much. Here the thickness for each layer is set to be 1.5 mm. The width of the microstrip line is designed to be 4.8 mm to achieve 50-Ω characteristic impedance. The microstrip line is fed using two N type coaxial connectors at both ends. The length of the microstrip line on the first and second layer is 50 and 40 mm, respectively. A 2-mm diameter via is used to transmit the signal from first layer to second layer.
Figure 2- 31. Designed multilayer microstrip structure with N type connectors.

Figure 2- 32. Simulated S-parameters of the multilayer microstrip line structure with N type connectors.

Figure 2-32 plots the HFSS simulated S-parameters of the designed multilayer

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microstrip line structure. Below 6 GHz, the insertion loss is smaller than 2 dB and the reflection coefficient is smaller than -10 dB. The high insertion loss and reflection coefficient close to 10 GHz is probably due to the frequency is approaching the high frequency limit of N type connector (11 GHz) and the discontinuity of microstrip impedance between two layers. Of the 1.8 dB loss at 6 GHz, 0.7 dB is due to dielectric loss and 1.1 dB is due to mismatch and radiation loss. With a substrate made of thermoplastics with a lower loss tangent than 0.01 (e.g., polycarbonate with loss tangent of 0.005 or graft polymer with loss tangent of 0.001), the insertion loss can be further decreased.

2.3.3 3D Printing of the Multilayer Microstrip Line

The designed multilayer microstrip structure is created using ultrasonic embedding to submerge wire and wire meshes into 3D printed thermoplastics. SolidWorks CAD software is used to create the design and a Makerbot Replicator material extrusion 3D printer is used to realize the design. The ABS is printed at 235 °C, using a nozzle with a diameter of 0.4 mm, and sliced using ReplicatorG at 95% fill factor. A channel is created based on mesh thickness (0.5 mm) where each mesh plane is to be embedded. The printing process is completed using a one pause build sequence located at a z-height of 2.2 mm. At the pause, a copper mesh (0.5 mm wire spacing) ground plane is thermally
embedded into the ABS substrate and fed through a 3D printed gap in order to create the second ground plane. At this same z-height, the mesh microstrip plane is thermally embedded and prepared for a vertical via to connect to the second microstrip line. The piece is then re-registered on the build plate and the build is resumed to print the next dielectric layer. After the second dielectric layer is finished, the second microstrip plane is embedded on top of the substrate. A through-via is then connected using a copper mesh formed cylinder and soldered to the top and bottom microstrip planes. Print accuracy in the Z-direction is approximately 0.4 mm and total print time is approximately 40 minutes. Additionally, there is 20 minutes time period required for registration and embedding.

![Figure 2-33. 3D printed multilayer microstrip line structure with N type coaxial connection.](a) Top view (b) Bottom view.]

### 2.3.4 Testing and Results

Figure 2-33 shows the photo of a 3D printed multilayer microstrip sample. Two N
type coaxial connectors are soldered at both ends of the microstrip structure. The $S$-parameters of the microstrip are measured using a vector network analyzer (Agilent E8361A).

![Figure 2-34. Measured $S_{21}$ of the multilayer microstrip structure compared with simulation.](image)

The measured $S_{21}$ is plotted in Figure 2-34, together with simulation results. The measured results agree with the simulation results below 7 GHz. It can be observed that the insertion loss is smaller than 2 dB below 6 GHz. The measured $S_{11}$ and $S_{22}$ are plotted in Figure 2-35. The reflection coefficients are smaller than -7 dB for all the frequencies. The main reason for the fluctuation in $S_{21}$ and higher reflection coefficient is because of the inaccurate substrate thickness during the printing process which makes the microstrip
impedance deviate from 50 Ω.

Figure 2-35. Measured $S_{11}$ and $S_{22}$ of the multilayer microstrip structure compared with simulation.

2.3.5 Multilayer Phased Array Design

Figure 2-36. 3D printable three-layer phased array antenna design.
Based on the previous multilayer microstrip structure, a 3D printable compact three-layer 4-element phased array operating at 3.5 GHz is designed. As shown in Figure 2-36, the first layer at the bottom is a 1 to 4 Wilkinson divider. The second layer includes four voltage controlled analog phase shifters and the third layer is the radiating element layer which includes four patch antennas. The simulated reflection coefficient from 3.4 to 3.6 GHz and radiation pattern at 3.5 GHz with four channels equally phased are shown in Figure 2-37. At 3.5 GHz, the reflection coefficient is lower than -10 dB and a high directive beam is shown in broadside. This multilayer phased array structure will be printed using FDM and ultrasonic wire embedding technique, the phase shifters will be integrated using a laser welding approach [127] and the beam steering capability of this
structure will be tested by controlling the phase distribution of the four phased shifters.

2.3.6 Summary

Section 2.3 demonstrates a 3D printed multilayer microstrip line structure with vertical transition by employing the ultrasonic wire embedding technique and the FDM technique. Measured results show that the insertion loss of this structure is smaller than 2 dB below 6 GHz and agrees with simulation. Compared to the conventional microstrip manufacturing approach, this 3D printed multilayer structure can be created with lower cost and the printing process is more convenient and faster. This 3D printing process of both dielectric and conductor can be applied to the additive manufacturing of more sophisticated EM structures for microwave applications. For example, based on the designed multilayer microstrip structure, a 3D printed compact three-layer phased array operating at 3.5GHz is designed. The simulated results show a reflection coefficient smaller than -10 dB at the working frequency and a high directional beam achieved at expected direction.
CHAPTER 3 3D PRINTED COMPONENTS IN THz FREQUENCY

3.1 Terahertz All-Dielectric EMXT Waveguide to Planar Microstrip Transition Structure

3.1.1 Introduction and background

Research involving the Terahertz spectrum (100 GHz - 10 THz) has experienced rapid growth in recent years. Many applications involving chemistry and biochemistry spectroscopy [94], security screening [93], medical imaging [128], radio astronomy [129], nondestructive testing and quality control [130], etc., have been proposed or demonstrated. However, at this time, the paucity of high performance and low cost components remains a major bottleneck in the realization of the promises of THz technology for many of these exciting applications. In addition, low cost and efficient integration techniques for THz micro-systems are necessary before wide range THz applications can be realized.

Electromagnetic crystal (EMXT), as a periodic arrangement of dielectric or metallic structures, provides frequency band gaps at which electromagnetic wave propagation is forbidden in the crystal [131]. It has been widely applied in waveguide / antenna applications to reduce the surface wave, improving efficiency [132,133] or enhancing
antenna gain [134,135,136]. Previously, an all dielectric circular shaped EMXT waveguide working around 105 GHz was first proposed based on a triangular lattice air-cylinder array in a dielectric background [137]. This structure exhibits electromagnetic band gap in the designed frequency bands because of the Bragg diffraction in the lattice. Wave propagation is prohibited within these band gaps and therefore a hollow core channel in the crystal structure will be able to confine and guide wave propagation along the channel. The fabrication was implemented using a polymer jetting technique [13], which enables quite convenient, fast and inexpensive fabrication of terahertz components with arbitrary complexity and shapes. A low propagation loss of 0.03 dB /mm at 105 GHz is obtained from THz time-domain spectroscopy (TDS) experimental characterization of the fabricated EMXT waveguide. Based on this EMXT waveguide structure, an EMXT horn antenna is also designed by flaring the waveguide channel out from 4.2 mm to 8 mm [25]. The return loss of this antenna is better than 30 dB over the simulated frequency range. Characterization of the antenna radiation pattern is performed using the THz-TDS set up. Highly directional radiation patterns are observed in the designed frequency bands up to 180 GHz [25].

To achieve efficient coupling from THz EMXT waveguide to planar integrated circuits and finally realize integrated THz-micro-systems with all kinds of applications, an EMXT waveguide to microstrip transition structure is theoretically and experimentally
demonstrated to convert the terahertz wave in the EMXT waveguide to a planar microstrip line. This section is organized as the following. The design of the EMXT waveguide to microstrip transition structure is first introduced. A ridge on one of the broad walls of the waveguide is used to compress the electromagnetic fields of the rectangular waveguide $\text{TE}_{10}$ mode into the microstrip fields [138]. In this section, simulation results of the transition design including insertion loss and field distribution are presented. Fabrication process of this structure employs the polymer jetting technique and gold platting process is discussed. Experimental characterization results of the fabricated structure using the TDS is reported next. 6 dB insertion loss for the designed back-to-back structure is achieved around 110 GHz. Also, insertion loss of the transition structure together with two EMXT waveguide is tested, the measured result of 8 dB in the first passband agrees well with the simulation results.

3.1.2 Waveguide to Microstrip Structure Design

The designed waveguide to microstrip line transition structure is using a ridge on the broad walls of the waveguide to convert the electromagnetic field of the waveguide $\text{TE}_{10}$ mode to the microstrip line mode. It consists of a top part and a bottom part as shown in Figure 3-1. The top part is a tapered metalized polymer ridge with a width from 5.8 mm to 0.6 mm. The wide end is inserted into the output aperture of the EMXT waveguide and
the narrow end is connected to the top conductor of the microstrip line. The bottom part is a metalized trapezoidal shaped slab which is connected to the microstrip ground plane. Both top part and bottom part are metalized using electro-plating. The transparent wings on the top and bottom parts are physical supports only, which have no electric influence to the transition structure.

![Diagram of the EMXT waveguide to microstrip transition structure](image)

**Figure 3-1.** Diagram of the EMXT waveguide to microstrip transition structure.

### 3.1.3 Simulation

The simulated transmission of the back-to-back waveguide to microstrip line transition structure is plotted in Figure 3-2. In the simulation, the feeding source are two circular waveguide port (radius 4.2 mm) placed at two ends of the transition structure to simulate the output of the EMXT waveguide. The dielectric constant of the polymer is set
to 2.75 and the loss tangent is set to 0.03, which is measured using the THz time domain spectroscopy (TDS) [93]. The microstrip line is fabricated on Duroid-RO4003C board. The permittivity of the substrate is 2.2 and loss tangent is 0.006. The thickness of the microstrip line is 0.203 mm with a length of 20 mm and a width of 0.6 mm.

Figure 3-2. Simulated transmission of the transition structure.

In Figure 3-2, the transmission of the back-to-back transition structure is -4 dB at 100 GHz and gradually decreases to -15 dB at 220 GHz. The simulated XY plane electric field distribution in the center of the structure is plotted in Figure 3-3. It shows that most of the power is concentrated between the microstrip line and its ground, indicating that the designed structure converts the waveguide mode to the microstrip mode well.
Figure 3- 3. Simulated XY plane E-field distribution at the center of the structure.

3.1.4 Fabrication

A commercial polymer jetting rapid prototyping printer Objet Eden 350 is used to print the 3-D polymer structures [13]. The photo of the printed top structure and bottom structure are shown in the left picture of Figure 3-4. The total printing time is less than an hour. After polymer structures are printed, a thin layer of gold with thickness about 100 nm is sputtered on the surface of the polymer. This thin layer of gold serves as the seed layer for the next electro-plating process. Then, the polymer structure with the thin gold layer is inserted into a gold solution and a voltage is added between the gold layer on the surface of the polymer and the ground for plating. After 6 hours plating, the thickness of the gold becomes about 6 um and the transition structure is ready to be assembled. The
photo of the top part after plating is shown in the right picture of Figure 3-4. The black region is the place for mounting the plating electrode and fixing the structure to prevent it from dropping into the gold solution. This region is out of the effective region of waveguide and transition ridge, so it has no electrical influence to the transition structure.

Figure 3- 4. The printed polymer structures before plating (left picture) and after plating (right picture).

After the bottom structure and two top structures are platted with gold, they are assembled and secured together by four screws at the end of the wings. A photo of the entire waveguide to microstrip line transition structure is shown in Figure 3-5. The blue colored portion is the polymer supports and the yellow colored portion is the gold layer plated on the surface of the polymer. The 20-mm long microstrip line fabricated on Duroid-RO4003C board is mounted on the center of the bottom structure and the two
ends of the microstrip line are connected to the ridge of the two top structures.

![Figure 3- 5. Photo of the entire waveguide to microstrip line structure.](image)

### 3.1.5 Characterization

In the work, the performance of the 3D printed transition structure is characterized using a THz time domain spectrometer (THz TDS). The printed back-to-back transition structure is mounted between the transmitter and receiver of TDS. Two parabolic mirrors and two 3D printed polymer convex lenses are used to couple the THz signal into the device under test as shown in Figure 3-6. The polarization of the THz signal is parallel to the horizontal plane. Free-space measurement with only parabolic mirrors and 3D printed polymer convex lenses is used as reference to calibrate the insertion loss of the device. The measured time domain signal and calibrated insertion loss of the device from 80~260
GHz with statistical error bars are shown in Figure 3-7. The measured insertion loss is about 6 dB around 110 GHz and 12.6 dB at 220 GHz. Figure 3-8 shows the measured insertion loss compared with simulation results. One can see that the measurement results are consistent with the simulation results.

![THz time domain spectrometer setup](image)

**Figure 3-6.** THz time domain spectrometer setup to characterize the 3D printed transition structure.
Figure 3-7. Measured (a) time domain signal and (b) calibrated insertion loss of the back to back transition structure with statistical error bars.
Figure 3-8. Measured insertion loss of the back to back transition structure compared with simulation.

Figure 3-9 shows the measurement setup to characterize the performance of the transition structure together with two 3D printed EMXT waveguides. The two ends of the transition structure are inserted into the two output apertures of the EMXT waveguides. The lengths of the two EMXT waveguides are 50 mm and 75 mm respectively. The measured insertion loss of the transition structure together with two EMXT waveguide from 80–260 GHz with statistical error bars is shown in Figure 3-10. About 8 dB insertion loss is achieved in the first pass-band of EMXT waveguide. The simulated insertion loss of the 3D printed transition structure together with two EMXT waveguides
is shown in Figure 3-11. From the simulation results, the 8 dB insertion loss in the first passband is consist of 1 dB loss from microstrip line + 2*1.65 dB loss from the back-to-back transition structure + 3.75 dB from EMXT waveguides.

Figure 3- 9. THz time domain spectrometer setup to characterize the 3D printed transition structure together with EMXT waveguides.
Figure 3-10. Measured insertion loss of the 3D printed transition structure together with two EMXT waveguides.

Figure 3-11. Simulated insertion loss of the 3D printed transition structure together with two EMXT waveguides.
To verify that the transition structure successfully changes the THz signal from waveguide mode into microstrip line mode, we also characterized the insertion loss of the system with a cross-polarization setup which the transition structure is rotated by 90 degree. The characterization setup is shown in Figure 3-12. Since the polarization of THz signal is parallel to the horizontal plane, the signal will not go through the microstrip line with this polarization. The measured insertion loss is shown in Figure 3-13. As expected, the signal cannot go through the transition and the insertion loss is larger than 20 dB for all the frequencies.

Figure 3-12. THz-TDS configuration to characterize the 3D printed transition structure with cross-polarization setup.
We also measured the insertion loss of the system with the microstrip line disconnected in the center to verify the waveguide mode changes into microstrip line mode. The experiment setup is shown in Figure 3-14. One can see that the microstrip line is not connected in the center. The measured insertion loss of the system is shown in Figure 3-15. One can see that the THz signal cannot go through the transition structure with the disconnected microstrip setup which means the original designed transition structure do successfully changes the signal from waveguide mode into microstrip line mode.
Figure 3- 14. Disconnected microstrip line configuration to verify the waveguide mode changes into microstrip line mode.

Figure 3- 15. Measured insertion loss of the 3D printed transition structure with microstrip line disconnected.

3.1.6 Summary and Future Work

Terahertz all-dielectric waveguide to microstrip line transition structures are designed,
simulated and fabricated. To characterize this waveguide to microstrip line transition structure, a THz time-domain spectrometer is employed to measure the transmission of the structure. Two parabolic mirrors and two 3D printed dielectric lenses are used to feed THz wave into the 3D printed transition structure.

The measured insertion loss of the back-to-back transition structure is 6 dB around 110 GHz. And 8 dB insertion loss is obtained in the first pass-band for the EMXT waveguide together with the transition structure. Measurement result of this EMXT waveguide to microstrip transition is consistent with design simulation. This type of waveguide to planar transition can be used to achieve efficient integration of 3-D all-dielectric passive components with active planar integrated circuits. Compared to traditional transition structure in THz frequency, this waveguide to planar transition can be fabricated with much lower cost and the fabrication process is much convenient and faster.

3.2 3D Printed Dielectric Reflectarrays: Low-Cost High-Gain Antennas towards Terahertz Applications

3.2.1 Introduction

Terahertz technology is rapidly emerging as a new frontier of electromagnetic
research, while merging the gap between microwave and optical engineering communities. Despite this increased interest in THz technology and applications, little commercial emphasis has been placed on THz systems [93]. This is perhaps due to the numerous new challenges that need to be addressed for practical implementation of THz technology at a wide scale. For antenna engineers, the special requirements of terahertz instruments demand new antenna concepts and also new ways of implementing already established designs [139]. In many applications of THz systems, such as radio astronomy, remote sensing, and radar, large reflector antennas with high surface accuracy and light weight are required. These designs however are typically high cost due to the high precision required for fabrication.

Reflectarray antennas on the other hand, combine some of the best features of reflectors and array antennas, and create a hybrid high-gain design with low-mass, low-profile, and also low-cost features [140, 141]. Most reflectarray antenna research in the recent years however has been in the microwave and sub-millimeter range [142, 143, 144, 145]. While microwave concepts can generally be extended to THz, at the short-wavelength region, several factors come into play that complicate the antenna design. Of these factors, the most important is arguably the element loss. Reflectarray antennas at THz frequencies were investigated in [146, 147, 148]. In these designs variable-size square patch elements were used for the reflectarray phasing elements.
These studies revealed that while a good performance may be attained for the elements with high-quality materials, low-cost (and typically high-loss) designs cannot achieve a satisfactory performance. The primary reason is that at THz frequencies, the conductor losses in these resonant patch elements increase to a degree that results in a significant loss of power and phase tuning range. As a possible solution to this problem, dielectric type elements were proposed in [147].

The goal of this work is to demonstrate the performance of dielectric reflectarray antennas as a solution to eliminate the cumbersome conductor losses at THz frequencies [149]. In contrast to the dielectric resonator type elements [150, 151] that use high dielectric constant materials, the focus here is on the use of conventional dielectric materials with low dielectric constant, which is compatible with the 3-D printing technique. Variable height dielectric slabs are used for the reflectarray elements design, which enables the utilization of low dielectric-constant materials. In addition, a polymer-jetting 3-D printing technology is utilized to fabricate the antenna, which can realize low-cost rapid prototyping. To demonstrate the feasibility of this approach, 3 different dielectric reflectarrays operating at 100 GHz are designed, and numerical and experimental studies are carried out for all prototypes that show a good agreement. Moreover, the proposed methodology is readily scalable and with the current material and fabrication technology, high-gain and low-cost dielectric reflectarrays operating at 1.5
THz is possible.

3.2.2 Dielectric Phasing Elements for Reflectarrays

A. Material Losses in Reflectarray Elements

For reflectarray antennas operating at THz and optical frequencies, material losses are a major concern. In general, material losses in reflectarrays include dielectric loss, conductor loss, and surface wave excitation [152,153], where the first two terms are typically dominant. While material losses are always taken into account in reflectarray designs [140, 141], in the microwave band the total material loss typically does not exceed 0.5 dB when high quality laminates are used for the design. At THz and optical frequencies however, a significant increase in material loss is observed which is primarily attributed to the losses of the conductor. High quality conductors such as gold significantly reduce these losses; however this comes at the expense of a much higher cost, particularly for high-gain arrays. Low-cost and typically high-loss conductors on the other hand pose additional problems. In addition to the loss of reflected power due to material losses, reflectarray elements that exhibit a high level of material loss may also show a different reflection phase response. In a recent study [147] it was shown that when the material loss in reflectarray elements increases beyond a certain limit, a new phase curve with a limited angular range will be observed. This reduced phase range of
the elements would ultimately result in a further reduction of the antenna efficiency.

Similar to optical fibers [154], a possible solution to the cumbersome conductor loss problems at THz frequencies is by avoiding the use of resonant conductor elements in the reflectarray unit-cell. A dielectric reflectarray element can be designed to control the reflection phase, by tuning the dimensions of the dielectric. A variety of low-loss dielectric materials are available at the THz and optical ranges that can be used for such a design, however the focus here is on both low-cost materials and fabrication techniques at a competitive cost, with the aim to reach the practical barriers of wide scale deployment.

B. Dielectric-Type Elements for Reflectarrays

In dielectric-type reflectarray elements, the resonant conductor patch is removed, and phase control is achieved by changing the geometrical parameters of the dielectric. In general two different approaches are available depending on the availability of the material and the fabrication technique. In the first approach, the phasing elements of the reflectarray are dielectric resonators [155,156], and phase tuning is typically achieved by changing the length of the dielectric cavity. These dielectric resonator reflectarray antenna (DRRA) elements exhibit high gain, however they require materials with high dielectric constant and very low loss, which makes them quite expensive.

In the second approach, one uses a dielectric slab for phase control of the reflectarray
elements where phase tuning is achieved by varying the thickness (height) of the dielectric slab in each unit-cell. These dielectric reflectarray elements put no constraint on the dielectric constant of the material, which would allow one to select low-cost materials for the design. Similar to optical mirrors [157], the ideal dielectric reflectarray would have a smooth profile; however an important consideration for low-cost fabrication is the minimum size of the pixel (unit-cell) that would yield satisfactory performance. This issue will be discussed in section 3.2.3. It is also important to note that for dielectric-type reflectarray elements, the reflective nature of the antenna still necessitates the use of a conductor ground plane; however the losses on the conductor ground are quite small compared to those of resonant conductor elements. A schematic model of a dielectric type reflectarray element unit-cell is given in Figure 3-16. Note that in the dielectric reflectarray design, the slab may cover the entire unit-cell surface, which is the case in this study, but this is generally not a necessary requirement for the elements.

Figure 3-16. A schematic model of a dielectric reflectarray phasing element.
C. Measurement of Dielectric Properties of the Material

The first step in designing the dielectric reflectarray was characterization of the electromagnetic properties of the polymer material that will be used for the design. This was done by measuring the transmission response of a uniform 3mm thick slab using a THz time-domain spectrometer (THz-TDS) [13, 158]. The system operates by propagating a picosecond-duration THz pulse through the material under test and extracting the frequency domain characteristics via Fourier transformation. The TDS spectral range is from 50 GHz to 1.2 THz, with a frequency resolution of 5 GHz. The THz-TDS system is shown in Figure 3-17.

Figure 3-17. The THz time-domain spectrometer system.
The measured dielectric properties of the sample are shown in Figure 3-18. The dielectric constant decreases slowly as the frequency increases, from 2.78 at 100 GHz to 2.70 at 600 GHz, while the loss tangent increases slowly from 0.02 to 0.05 across the measured spectrum. Despite a slightly higher loss at the high end of the spectrum, the dielectric properties of this material are stable, and therefore quite suitable for realizing dielectric reflectarrays.

![Figure 3-18. Measured dielectric properties of the polymer material.](image)

### 3.2.3 Design of THz Dielectric Reflectarray Antennas

#### A. Enabling 3D Printing Technology

In the THz spectrum (informally defined as 100 GHz to 10 THz), the corresponding wavelength is 3 mm to 30 μm, thus a fabrication technique that can provide feature sizes in the μm range is needed. The exact feature requirements however depend on the
particular design and operating frequency, and will be discussed later on in section 3.2.3.B.

A relatively new and promising fabrication technique that can offer resolutions in this range is 3-D printing, also known as additive manufacturing. Polymer-jetting rapid prototyping machines are a class of 3-D printers that can generate polymer objects with arbitrary shape and geometry from a digital model by laying down polymer materials layer by layer. For our design we used the Objet Eden 350 polymer-jetting rapid prototyping 3-D printer that has a fundamental resolution of $42\mu\text{m} \times 84\mu\text{m} \times 16\mu\text{m}$ in $x$, $y$, and $z$ dimensions, respectively. It is worthwhile to point out that in comparison with traditional machining techniques; one of the distinct features of 3-D printing is that it avoids the removal of materials by drilling or cutting methods. As such, 3-D printing allows for rapid prototyping of arbitrary shapes with low cost for mass production.

To fabricate the object, the 3-D model of the structure has to be generated and then imported into a CAD program, which is then converted to a series of layered slices, each layer representing a $16\mu\text{m}$ thick region of the model. Once the printer receives the data for each slice, a series of print heads, similar to the print head on an ink-jet printer, deposits a thin layer of ultraviolet-curable polymer on the construction stage. Then the ultraviolet lamps on the print head immediately cure the materials when they are being deposited. After one layer is completed, the construction stage is lowered by $16\mu\text{m}$, and
the next slice is printed on top. Once the entire model is complete, the construction tray rises and the part can be removed. For traditional rapid prototyping, the final step is a high-pressure water spray to remove the water-soluble support material, leaving just the model material in the desired 3D shape. A photo of the prototyping machine Object Eden 350 is shown in Figure 3-19.

![Figure 3-19. Photo of the polymer-jetting rapid prototyping machine.](image)

**B. Reflectarray Element and Aperture Designs**

With the material properties and fabrication resolutions specified, the next task was to design the dielectric reflectarray element and system. In our designs, the element is a dielectric slab covers the entire unit-cell surface. Viewing this design as an array, we set the lattice size to be half-wavelength at the center design frequency. The height of the
slab is then designed to provide the required phase shift on the reflectarray aperture. To demonstrate the feasibility of this design approach, dielectric reflectarrays are designed for the operating frequency of 100 GHz. It should be noted here that while the design methodology proposed here is readily scalable, and we are capable of fabricating the dielectric reflectarray for much higher operating frequencies, due to the availability of the feed antenna for the reflector, the antenna is designed for 100 GHz operation. Further discussions on the minimum size of the lattice and limitations in frequency scaling will be given in section 3.2.3.C.

The measured electrical properties of the dielectric material at 100 GHz are \( \varepsilon_r = 2.78 \), and \( \tan\delta = 0.039 \). The unit-cell periodicity is selected to be 1.5×1.5 mm\(^2\). For the element design, each unit-cell is treated as an infinite slab of dielectric; therefore either unit-cell analysis or analytical solutions for infinite slabs can be used to derive the reflection properties of the elements. One important consideration however is the position of reference plane for phase computation. Conventionally the reflectarray aperture has a flat surface, and the reflection phase is computed on that surface. In this design, elements have variable heights, so we define a reference plane that is placed at the top surface of the highest slab, which is schematically depicted in Figure 3-20(a). A full phase-cycle (2\(\pi\)) is then achieved by changing the slab height from 0.3 to 2.32 mm with a resolution of 16 \(\mu\)m as shown in Figure 3-20(b). It is important to note that the reflection phase
curve shown here is for an oblique incident angle of $\theta = 25^\circ$ under perpendicular polarization. Also note that while the phase range of this element is sufficient for designing a reflectarray antenna, the phase is limited to one phase cycle, so it would be necessary to zone the array. Similar to zoned dielectric lenses [159], this would reduce the overall profile and weight of the antenna, but would also result in a reduction of antenna bandwidth.
The dielectric reflectarray has a square aperture with a side length of 30 mm, and 400 variable-height dielectric elements are designed to provide the necessary phase shift on the aperture. An offset feed is selected to avoid blockage effects with a feed tilt angle of 25°. Based on the aperture efficiency analysis, the feed antenna (A-INFO LB-10-10) is placed at a distance of 22.5 mm from the aperture, and is pointing towards the geometrical center of the array. The main beam direction is also set to 25° off broadside.

Different aperture phase distributions are studied for the dielectric reflectarrays. Note that in Figure 3-20(b) a phase constant of 148° is added to the phase curve so that the tallest slab will have a quantized phase close to zero. The initial phase distribution on the reflectarray aperture is then computed as described in [140, 141]. At the center of the array, the required reflection phase is 103.7°. One concern for this offset dielectric reflectarray is the shadowing effect observed along the phase wraps on the aperture where the taller elements will intercept the incoming rays and shadow their adjacent elements. To minimize the shadowing effects, one can minimize the number of zones (phase wraps) by adding a phase constant to the aperture, which would also increase the bandwidth of the antenna as will be shown in section 3.2.4. The aim is to have elements with maximum height at the center. Since the phase moves outwards on the aperture, this ensures that all

Figure 3-20. Reflection coefficients of the dielectric reflectarray elements at 100 GHz.
element sizes are used before the first phase wrap is observed. For this design (Design 1) this corresponds to a phase constant of -100°, and a reflection phase of 3.7° at the center of the array. This phase distribution is shown in Figure 3-21(a). It is worthwhile to point out that for the extreme case (when the tallest and shortest slabs are placed next to each other); the angular limit arising from this shadowing effect is about 20.5°. As such in this design where the main beam is scanned to 25° off broadside, very few elements will observe this shadowing effect. Nonetheless this design approach (Design 1) will minimize the number of these shadowed elements.

Another practical consideration is regarding the material losses of the elements. As shown in Figure 3-20(b), the element loss increases with the element thickness. In this case, the target is to minimize losses by using an element distribution that achieves the lowest antenna loss. Similarly, this is obtained by adding a phase constant. The function to be maximized is the sum of the weighted element reflection coefficients for the array. The weighted element loss [160], which takes into account both the element loss and the aperture illumination, is computed as:

\[
WEL = \frac{\sum_{m} \sum_{n} \text{Illumination}(m, n) \cdot |\Gamma(m, n)|}{\sum_{m} \sum_{n} \text{Illumination}(m, n)}. \tag{3-1}
\]

For this design (Design 2), the minimum loss will be realized with a phase constant of +82° with the aperture phase distribution shown in Figure 3-21(b). This corresponds to
a reflection phase of 185.7° at the center of the array.

A third design is also studied which uses only two slab thicknesses. This design is basically a Fresnel zone plate reflector antenna, which in array terminology can be referred to as a 1-bit design [161]. In general this design will suffer from the classical phase quantization errors of phased array antennas. Nonetheless, it is a necessary reference study to quantify the fabrication limits (resolution in slice thicknesses for 3-D printing) for higher frequency operations. The phase distribution for this design (Design 3) is given in Figure 3-21 (c).

Figure 3- 21. Aperture phase distributions for the dielectric reflectarrays: (a) design for minimum phase wraps (Design 1), (b) design for minimum element loss (Design 2), (c) 1-bit design (Design 3).

C. 3-D Models and Radiation Performance of THz Dielectric Reflectarrays

For these reflectarray designs, all elements have a unit-cell lattice of 1.5×1.5 mm²,
but with a different height which is determined by the required phase shift. To automate
the process of 3-D model generation, geometry files based on standard 3-D formats need
to be created. A suitable file format for this design is STL, which is used by our Objet
Eden printer CAD program, and is also available in many commercial electromagnetic
solvers such as ANSYS HFSS and CST Microwave Studio. For this design, each cube is
defined by specifying the vertex positions of 12 triangles. A MATLAB code is developed
to generate the STL files that contain the location and dimensions of the elements. The
3-D models of these dielectric reflectarrays are shown in Figure 3-22.

![Figure 3-22. 3-D models of dielectric reflectarrays in ANSYS HFSS: (a) Design 1, (b) Design 2, (c) Design 3.](image)

The radiation performance of the dielectric reflectarrays is obtained using the
full-wave simulation software CST Microwave Studio. The radiation patterns of the 3
dielectric reflectarrays are given in Figure 3-23 where it can be seen that in all designs the
main beam is correctly scanned to 25°. A summary of the antenna performances is given in Table 3-1. Note that the radiation patterns of Design 1 and 2 are almost similar, with a slightly better side-lobe performance for Design 1. On the other hand Design 2 achieves a lower element loss and higher gain and radiation efficiency as expected. Comparison of the performance of the 1-bit design (Design 3) with the other two dielectric reflectarrays indicates a directivity loss of about 3 dB, which is due to the phase quantization errors.

Figure 3-23. Simulated gain patterns of the dielectric reflectarrays at 100 GHz.
To observe the effect of profile smoothness on the performance of the array, Design 1 is also studied with a finer lattice resolution. As discussed earlier, the dielectric reflectarrays were designed with a lattice size of $\lambda/2$ at the center design frequency. Here we study the performance of this design with a lattice size of $\lambda/10$. The 3-D model of the dielectric reflectarray is given in Figure 3-24(a). The simulated gain patterns for these two different lattice sizes are compared in Figure 3-24(b), where it can be seen that despite a slightly higher gain (about 0.3 dB), the radiation performance of these two designs is almost identical. This study reveals that while there is some advantage in increasing the accuracy of the model, a resolution of half-wavelength is quite sufficient to achieve a good performance with these designs. As such, this minimum lattice size can be used directly to determine the upper frequency limit based on the available fabrication capability. With the Objet Eden 350 printer, a lattice size of 100$\mu$m×100$\mu$m can be reliably realized. For half-wavelength cells, this would correspond to an operating frequency of 1.5 THz, and with a slicing resolution of 16$\mu$m for the slab heights, the

<table>
<thead>
<tr>
<th>Design</th>
<th>DIRECTIVITY</th>
<th>GAIN</th>
<th>RADIATION EFFICIENCY</th>
<th>SLL</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>26.48 dB</td>
<td>24.69 dB</td>
<td>66.22%</td>
<td>-20.7 dB</td>
</tr>
<tr>
<td>2</td>
<td>26.34 dB</td>
<td>24.96 dB</td>
<td>72.78%</td>
<td>-18.2 dB</td>
</tr>
<tr>
<td>3</td>
<td>23.80 dB</td>
<td>23.09 dB</td>
<td>84.92%</td>
<td>-11.8 dB</td>
</tr>
</tbody>
</table>

Table 3-1 Summary of Dielectric Reflectarray Antenna Radiation Performances at 100 GHz
maximum quantization errors for this design will be less than \( \lambda/8 \), which is quite acceptable for high-gain arrays.

![Diagram](image1)

(a)

![Diagram](image2)

(b)

Figure 3-24. Effect of lattice size on the performance of dielectric reflectarrays: (a) aperture phase and 3-D model of Design 1 with a lattice size of \( \lambda/10 \), (b) radiation patterns of Design 1 at 100 GHz with two different lattice sizes.
3.2.4 Prototype Fabrication and Measurements

The three dielectric reflectarrays designed in the previous section are all fabricated using our Objet Eden 350 polymer-jetting rapid prototyping 3-D printer. After the three polymer structures are printed, a thin layer of gold with thickness about 100 nm is sputtered on the back surface of each polymer structure as a seed layer. Then, the polymer structure together with the thin gold layer is inserted into a gold solution and a voltage is applied between the gold layer on the surface of the polymer and the ground for plating. After 6 hours plating, the thickness of the gold becomes about 6 μm and the reflectarray antenna is ready for test. Photos of fabricated prototypes are shown in Figure 3-25.

Figure 3-25. Top view of the fabricated dielectric reflectarray prototypes: (a) Design 1, (b) Design 2, (c) Design 3. The back side is gold plated.
The radiation patterns of the dielectric reflectarrays are measured using a vector network analyzer (Agilent E8361A) with W-band extension heads. A pyramidal horn antenna with a measured gain of 12 dB at 100 GHz is used to feed the dielectric reflectarray. The feeding horn is placed at a distance of 22.5 mm away from the aperture and pointing toward the center of the reflectarray with a tilted angle of 25° as described in the previous section. Another W-band horn antenna is located at the far field distance to measure the radiation pattern. Another standard gain horn is also used as the transmitter to calibrate the gain value of the dielectric reflectarray. Comparison between the simulated and measured radiation patterns in the $xz$-plane at 100 GHz are shown in Figure 3-26. Note that for the measurements the array is rotated 25° in the aperture plane, so the main beam is pointing to 0°.

![Graph showing simulated and measured radiation patterns.](image)
Figure 3- 26. Comparison of measured and simulated radiation patterns of the dielectric reflectarrays at 100 GHz: (a) Design 1, (b) Design 2, (c) Design 3.

It can be seen that for all three designs, a close agreement between the measured and
simulated radiation patterns is observed. The measured gain of the three prototypes at 100 GHz is 22.5, 22.9, and 18.9 dB, respectively. The measured half-power beam-widths are 6.95, 6.65, and 6.80 degrees, respectively. The discrepancy between measured and simulated gain is attributed to material property uncertainty, alignment errors, and fabrication and measurement errors, which becomes large at this high frequency. Furthermore, Design 3 suffers more from the fabrication error, since the middle part of the aperture (0.3 mm) is too thin to maintain the flatness.

The gain of these dielectric reflectarray prototypes was also measured across the frequency range of 70 to 110 GHz. These results are given in Figure 3-27.

![Figure 3-27. Measured gain versus frequency for the dielectric reflectarray prototypes.](image)
As expected, Design 1 demonstrates the widest bandwidth which is attributed to the smaller number of phase wraps on the aperture. Design 2 on the other hand demonstrates the highest gain due to the lower loss of its elements. Despite a much lower gain in comparison with the other 2 designs, Design 3 shows a slightly broader bandwidth than Design 2. A summary of the gain performance of these antennas is given in Table 3-2.

<table>
<thead>
<tr>
<th>Design</th>
<th>Gain at 100 GHz</th>
<th>HPBW at 100 GHz</th>
<th>1-DB Gain Bandwidth</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>22.5 dB</td>
<td>6.95°</td>
<td>89.4-110.0 GHz (20.66%)</td>
</tr>
<tr>
<td>2</td>
<td>22.9 dB</td>
<td>6.65°</td>
<td>92.0-105.3 GHz (13.48%)</td>
</tr>
<tr>
<td>3</td>
<td>18.9 dB</td>
<td>6.80°</td>
<td>90.9-104.6 GHz (14.02%)</td>
</tr>
</tbody>
</table>

3.2.5 Conclusions

In this section, dielectric reflectarray antennas are proposed as a low-cost solution for high-gain terahertz antennas. Variable height dielectric slabs (low-cost polymers) are used for the reflectarray elements, and a polymer-jetting 3-D printing technology is utilized to fabricate the antenna which has the capability to realize rapid prototyping at a low-cost. Three different prototypes operating at 100 GHz have been designed and experimental results demonstrate good performance. The proposed methodology is readily scalable and with the current material and fabrication technology, low-cost, high-gain antennas up to
1.5 THz can be realized. This study shows that the proposed method is a promising approach of realizing high gain THz antennas.
CHAPTER 4. DIRECTION OF ARRIVAL (DOA) ESTIMATION SYSTEM USING 3D PRINTED LUNEBURG LENS

4.1 Introduction

Microwave direction of arrival (DOA) estimation is important in many sensor systems and has attracted a lot of attention due to its wide applications in the commercial and military areas, such as wireless communications [162] and electronic warfare [163]. To achieve high resolution in the incident angle, a typical microwave direction finding system is based on antenna arrays with a large number of elements and sophisticated algorithms. However, the cost, speed and power consumption associated with the large number of hardware components and complicated signal processing algorithm could be impractical, especially for portable and commercial applications. Accurate and efficient direction finding will be very useful in next generation wireless communication system for location based services and applications.

Gradient index device is a kind of device with gradually changing refractive index inside the material. A Luneburg Lens [47] is an attractive gradient index device for wide angle radiation scanning because of its broadband behavior, high gain and the ability to form multiple beams. It has a superior performance compared with conventional uniform
material lenses that every point on the surface of an ideal Luneburg Lens is the focal point of a plane wave incident from the opposite side (the optical path of waves coming into the lens is shown in Figure 4-1). This special property allows precise direction finding based on amplitude only information, as proposed in [164].

However, the conventional methods for building 3-D spherical Luneburg Lens are quite expensive and time consuming. Recently developed polymer-jetting rapid prototyping technique has allowed 3-D Luneburg lens be fabricated easily and with much lower cost than traditional fabrication methods [22]. Figure 4-2(a) shows the designed structure of a Luneburg lens. Desired gradient index \( \frac{n^2}{\varepsilon_r} = 2 - \left(\frac{r}{R}\right)^2 \) is realized by controlling the filling ratio of air voids and polymer.

![Figure 4-1. Optical path of Luneburg lens. Every point on the surface of an ideal Luneburg Lens is the focal point of a plane wave incident from the opposite side.](image)
Figure 4-2(b) is the simulated and measured gain of the Luneburg lens at 10 GHz. The feed is an X-band waveguide mounted on the surface of the lens. It can be seen that when there is only one detector mounted on the lens, the antenna is mainly receiving incoming signals from the opposite direction. Then it is natural to imagine that with a number of detectors mounted on the Luneburg lens, direction of arrival can be estimated by considering the received power on all the detectors. The advantageous of this Luneburg lens based DOA estimation is that it is very wideband and it does not require any expensive phase shifter component. In addition, the high gain property of a Luneburg lens leads to small correlation between received power distributions for different incident angles, thus high accuracy for the incident angle estimation can be achieved.

Figure 4-2. (a) Discrete polymer cubes with different size used to control the dielectric constant distribution of the lens. (b) Simulated and measured radiation pattern of the lens.
In this work, a broadband Luneburg lens based DOA estimation system is studied. The lens is fabricated using the polymer jetting 3D printing technology and 36 detectors equally spaced with 10° separation on the equator of the spherical lens are used to receive the signal from all 360 degrees in the azimuth plane. A simple correlation algorithm is applied to estimate the DOA. The direction finding results show that the averaged estimation error is smaller than 1° for signals incident from all 360 degree angle, demonstrating that this Luneburg lens based direction finding system is a good candidate for portable and low cost applications.

4.2. Direction Finding Algorithm

Luneburg lens has the property that every point on the surface of an ideal Luneburg lens is the focal point of a plane wave incident from the opposite side. Therefore, if a number of detectors are distributed on the lens surface, different detectors will receive different power which depends on the direction of the incident signal. For example, the detector directly facing the incident wave will receive the highest power and the other detectors will receive smaller power. By strategically distribute a number of detectors and analyze their receive responses, the direction of the incident wave can be easily estimated. In this work, 36 zero biased diodes are used as detectors and equally mounted on the surface of a Luneburg lens with 10 degree separation to cover all 360 degree angle.
Here a correlation algorithm is used to calculate the DOA estimation. First, the output voltages of all the 36 detectors are recorded with different incident angles from 0º to 360º (step 1º). These voltage values at different incident angles are stored as the calibration file $V_{cal}$ (a vector of all the detector outputs). Next, direction finding performance of the Luneburg lens system is tested with an incident wave coming from all 360º angles. Again, the output voltages ($V_{signal}$) of all the 36 detectors are measured. For each incident angle, the measured output voltages of the 36 detectors are correlated with the calibration file.

$$Corr = \sum V_{cal} \cdot V_{signal}$$  \hspace{1cm} (4-1)

Then, the direction with the largest correlation will be the estimated direction of the incident wave.

$$\hat{\theta} = \arg \max_0 \{Corr(\theta_1), Corr(\theta_2) ... Corr(\theta_M)\}$$  \hspace{1cm} (4-2)

### 4.3 Experiment setup and measurement results

Figure 4-3 shows the schematic configuration of the Luneburg lens based DOA estimation system. The lens is designed to work at a broadband working frequency from 4 to 20 GHz. The diameter of the lens is 24 cm and fabricated using polymer jetting rapid prototyping technique [22]. Compared with conventional process, the fabrication process using this 3D printing technique is quite fast, convenient and the cost is much lower. The
receiver part of the system is consisted of 36 zero biased diodes (Model No. SMS7630-061) equally spaced with 10° separation on the equator plane of the spherical lens. 36 monopole antennas are used as the receiving antennas for each detector to receive the signal of the incident wave. The antenna and detector circuit is fabricated on an 8-mil Rogers-4003 substrate and mounted around the lens as shown in Figure 4-4. The outputs of the 36 detectors are measured using a voltage meter and a 1 to 36 multiplexer is used to select the reading of different detectors.

In the experiment, a signal generator (Agilent E8257C) connected to a double ridged horn antenna is used as the source. The photo of the transmitting antenna is shown in Figure 4-5(a) and the configuration of the 36 detectors connected to the multiplexer is shown in Figure 4-5(b). When we do calibration, the Luneburg lens is 3 meters away from the source. When we do DOA estimation, the Luneburg lens with detectors is moved further away from the source to a different distance (4 meter) to make sure the received signal is different from the calibration data and therefore more close to the real application case. Although the Luneburg lens used (as shown in Figure 4-4) is broadband (4 – 20 GHz), for proof of concept without losing generality, an operating frequency of 5.6 GHz is selected at which the detectors have its peak sensitivity.
Figure 4- 3. Schematic configuration of the Luneburg lens based DOA estimation system in the experiment.

Figure 4- 4. Experiment setup of the Luneburg lens for direction finding. 36 detectors are
mounted on the surface of the lens to receive the signal from different directions.

Figure 4-5. (a) Double ridged horn antenna used as the radiating source (b) 36 detectors connected to a multiplexer.

Since the sensitivity for different detectors are not exactly the same, therefore, equalization is needed for different detectors in the direction finding algorithm. In this case, all the measured voltages from the 36 detectors are normalized to its measured peak values (when they are directly facing the source) in the calibration data. Figures 4-6(a) and (b) plot the normalized voltages of each detector for the calibration and DF test measurements, respectively. The reason for the peak values of different detectors being different in Figure 4-6(b) is that the sensitivity of each detector is different at different
power levels (the output voltages of some detectors decrease faster than others with decreasing power).
The correlation algorithm discussed in the previous section is applied to estimate the source direction. For each testing incident angle, the normalized signal data in Figure 4-6 was correlated to the normalized calibration data from all the directions. The angle which has the largest correlation is the estimated direction of the incident wave.
Figure 4-7 plots the estimated direction using this Luneburg lens system versus the actual incident angle. Figure 4-8 plots the error of the estimated angle for different incident angles using the correlation algorithm. The averaged error over all 360 degree incident angles is 1.05 degree. This accuracy of incident angle will be enough to satisfy the requirement of many applications such as forward collision warning (require angle resolution within 1.5 degree) [165].

Other than the simple correlation algorithm, we also tried a compressive sensing (CS) algorithm to estimate the coming direction of the incident. Before doing DOA estimation, the calibration data of all the 36 detectors with different incident angles from 0º to 360º (step 1º) are used as prior knowledge. By applying the calibration data from all different directions as the projection bases \([h]\) and the measured data from all detectors as the output matrix \([g]\), a TWIST compressive sensing algorithm is employed to estimate the probability \([f]\) of signal coming from different directions.

\[
\begin{bmatrix}
g_1 \\
g_2 \\
\vdots \\
g_N \\
\end{bmatrix} =
\begin{bmatrix}
h_{11} & h_{12} & \cdots & h_{1M} \\
h_{21} & h_{22} & \cdots & h_{2M} \\
\vdots & \vdots & \ddots & \vdots \\
h_{N1} & h_{N2} & \cdots & h_{NM} \\
\end{bmatrix}
\begin{bmatrix}
f_1 \\
f_2 \\
\vdots \\
f_M \\
\end{bmatrix}
\]

(4-3)

The advantage of DOA estimation using CS algorithm is that it can provide the probability of incident wave for different directions. The disadvantage is that it will take more computational time compared to the simple correlation algorithm. Figure 4-9 shows
the estimated direction using the CS algorithm and Figure 4-10 plots the error of the estimated angle. One can see the maximum error using the CS algorithm is 3 degrees which is smaller than the maximum error using the correlation algorithm. The averaged error for all 360 degrees is 0.978 degree which is also smaller than the results using the simple correlation algorithm. Figure 8-11 shows an example of the calculated probability results when the incident wave is coming from -70 degree using the CS algorithm. One can see the probability result shows a clear peak at -70 degree direction.

In this work, the detectors are only mounted on the equator plane of the lens. If detectors are populated in a 3-D fashion on the lens surface, accurate 3-D direction finding can also be achieved.
Figure 4-7. Estimated direction results at different incident angles from all 360 degrees using the correlation algorithm.
Figure 4-8. Error of the estimated angle for different incident angles using the correlation algorithm.
Figure 4-9. Estimated direction results at different incident angles from all 360 degrees using the compressive sensing algorithm.
Figure 4-10. Error of the estimated angle for different incident angles using the compressive sensing algorithm.
Figure 4-11. Calculated probability results of an incident wave from -70 degree using the CS algorithm.

4.4 Conclusion

A direction finding system employing a 3D printed Luneburg lens is presented in this chapter. A system consisting of 36 detectors equally spaced with 10° separation on the equator of the spherical lens is demonstrated at 5.6 GHz. Using a simple correlation algorithm and a compressive sensing algorithm, the direction finding result shows the averaged error is 1.05 degree and 0.978 degree for incoming waves of incident angle
covering all 360 degrees. This kind of Luneburg lens based direction finding system may be a good candidate to achieve a portable, low cost and accurate direction finding system that will be useful for many applications.
CHAPTER 5. A NOVEL ELECTRONICALLY SCANNED ARRAY BASED ON LUNEBURG LENS

5.1 Introduction

Phased array technology is commonly used to obtain high antenna gain and control antenna radiation pattern. Because agile beams provide significant system advantages, phased arrays play an important role in high performance radar and communication systems and have attracted considerable amount of attention in broadcasting, weather, radio astronomy, and other space and ground based applications [166, 48]. The advantages of phased array include high gain and low side lobes, ability to scan the beam from one target to the next in a few microseconds, ability to provide an agile beam under computer control and multifunction operation by emitting several beams simultaneously.

Despite the advantages of phased arrays, there are still several challenges to be solved before a wide range of applications can become reality. First, the beam scanning coverage is limited to a 90-120 degree sector in the azimuth and elevation planes. Second, deformation of beam appears when scanning to different angles. Third, the bandwidth of a phased array is limited by the phase shifters used unless true time delay lines are incorporated. Most importantly, the high complexity of the phased array system, for example, large number of phase shifters, power splitters, interconnects, control units, etc.,
remains a barrier for a robust, low weight and low cost system.

To address these issues, a novel broadband electronic scanning array based on Luneburg lens is proposed [108] and a simple fix tuned demonstration has been accomplished [167]. The radiation elements of this array are mounted on the surface of a 3D Luneburg lens. By varying the phases and amplitudes of these elements, electronic beam controlling can be realized. With this Luneburg lens based phased array, the scan angle can cover the whole 360 degrees continuously. Due to the spherical symmetry of the Luneburg lens, the beam shape is almost identical for all scanning directions. Since to the first order, the refractive index of the designed Luneburg lens is independent of frequency, the array can operate in a very large frequency range. For example, a previously reported Luneburg lens works at least from 4 GHz to 20 GHz [22]. Meanwhile, unlike the conventional phased array that all the elements need to work at the same time, this Luneburg lens based electronically scanning architecture only needs very few number of feeds working at the same time to achieve high directional beam scanning. Therefore, it requires much less system complexity to achieve a high gain directional beam than the conventional phased array system. This reduction in system complexity allows the electronic scanning system to be built at much lower cost. In this chapter, a detailed design procedure of the Luneburg lens based continuously scanning array is developed. In addition, thorough beam synthesis process is described. Different beam
synthesis examples such as fan beam and null beam forming for different applications are discussed. Moreover, 2D beam scanning in both the azimuth and elevation planes is realized by distributing feeding elements over a 2D area on the lens surface.

This chapter is organized as following. Section 5.2 introduces the principles of the Luneburg lens based phased array. Then, the developed beam synthesize procedure is discussed in section 5.3. Next, some beam scanning pattern examples using the proposed phased array are shown in section 5.4. Finally, conclusion is given in section 5.5.

5.2 Proposed Luneburg Lens Phased Array Principle

5.2.1. Luneburg lens

Luneburg lens is a kind of gradient index component used for wide angle radiation scanning because of its broadband behavior, high gain and the ability to form multiple beams. It has an outstanding performance compared with traditional lenses made of uniform materials. Every point on the surface of an ideal Luneburg lens is the focal point of a plane wave incident from the opposite side as shown in Figure 5-1. This property of Luneburg lens makes it a very good candidate to form multiple beams with high gain and broadband behavior.

Usually, for a lens made of non-magnetic ($\mu_r = 1$) material, the index of refraction n
distribution of a spherical Luneburg lens is given by Equation (5-1) [22]:

\[ n(r)^2 = \varepsilon_r(r) = 2 - \left(\frac{r}{R}\right)^2 \] (5-1)

in which \( \varepsilon_r \) is the relative permittivity, \( R \) is the radius of the lens and \( r \) is the distance from the point to the center of the sphere.

Figure 5- 1. The focusing property of a standard non-magnetic Luneburg lens.

5.2.2. Lens Fabrication

A polymer-jetting rapid prototyping technique [13] is employed here to enable efficient and accurate fabrication of 3-D Luneburg lens. The desired gradient index is realized by controlling the filling ratio of polymer / air based unit cells [22]. The polymer jetting rapid prototyping technique is a technique that allows fast fabrication of polymer
components with arbitrary shapes and complexity [25,137,168,169]. A commercial polymer jetting 3D printer Objet Eden 350 is employed to fabricate the Luneburg lens [22]. The printer has a droplet size of 42 μm x 42 μm x 16 μm, which is more than sufficient for fabricating Luneburg lens below 100 GHz. Moreover, large structures with a size of up to 30 cm x 30 cm x 30 cm can be printed. With this polymer jetting technique, the fabrication process is relatively fast, convenient and inexpensive. The total printing time for a 24 cm diameter lens is less than 8 hours.

One example of a printed Luneburg lenses is shown in Figure 5-2. The polymer / air unit cell size is 5 mm and the lens has a diameter of 24 cm. The left photo shows the cross-section cut through the center of the lens and the right photo shows the entire lens. It can be seen that the unit cell filling ratio is larger in the center and decreases to zero at the surface of the lens.
Figure 5-2. Photographs of the fabricated Luneburg lens: (a) the cross-section cut through the center of the lens; (b) the entire lens (24 cm diameter).

5.2.3. Measured radiation pattern with a single feed

The radiation pattern of the fabricated Luneburg lens with a single feed (a J-band WR-137 coaxial to waveguide adapter on the surface of the lens) is shown in Figure 5-3. A standard gain horn antenna is used to calibrate the gain of Luneburg lens. It can be seen that the Luneburg lens works as a narrow beam antenna in a broad frequency band as predicted. The measurement results show that the half-power beam width of the $8\lambda_0$ (24 cm) diameter lens is $8^\circ$ at 10 GHz and the gain of the antenna is 23.7 dB. Although the J-band waveguide has a frequency limit from 5.85 to 8.2 GHz, this Luneburg lens has a much broader frequency range from 4 GHz to 20 GHz which has been tested using a X-band and Ku-band waveguide feed [22].
5.2.4. Luneburg lens phased array

Based on the Luneburg lens’s ability to form multiple beams with high gain and broadband behavior, a novel electronically scanning array structure was designed by mounting several sources / detectors around the lens as shown in Figure 5-4. Instead of having only fixed beams, it is proposed here to control the phase and amplitude of several adjacent feeding elements, similar to the conventional phased array, to obtain finer beam scanning as well as other desired radiation patterns. However, a key distinctive advantage
is that, unlike a conventional phased array that needs all the elements working together, this electronically scanning system only needs very few number of feeds working at the same time to achieve high directional beam scanning due to the high gain nature of the Luneburg lens itself. For example, if we need a high directional beam scanning between two adjacent sources / detectors, only several nearby feeding elements will be activated to achieve the desired pattern. Based on our simulation result, for a 12-degree HPBW Luneburg lens, when the feeding elements are placed 10 degrees apart (i.e., 36 elements in the horizontal plane), 3 - 5 adjacent elements working at the same time is sufficient to achieve adequate beam scanning with a 1-degree accuracy. Therefore, a much smaller number of phase shifters and control units are needed compared to a conventional array, leading to much reduced system complexity and cost. Other attractive advantages of this array architecture include ultra-wide frequency range, no scan angle coverage limit and no beam shape variation during scanning.
Figure 5-4. Schematics of the Luneburg lens based phased array structure. (a) A number of sources / detectors mounted around the lens. (b) Switching network to select required feeds and common digital beam formers (DBF) to control the amplitude and phase.

5.2.5. Mutual coupling

Mutual coupling is the coupling effect between different elements in a phased array which is very important since it would alter the matching characteristic of the antenna elements and array radiation pattern. To estimate the mutual coupling effect of this Luneburg lens based phased array structure, the S-parameters of 36 elements mounted on
the surface of a 12-cm diameter Luneburg lens as shown in Figure 5-5 were simulated. The radiating elements are dipole antennas with 10 mm length. The simulated $S$-parameters of these 36 elements are shown in Figure 5-6.

From Figure 5-6, one can see that the $S$-parameter from $S_{1,1}$ to $S_{1,36}$ are all smaller than -20 dB, which means the mutual coupling between these elements are very weak, most of the power are radiated out. Also, from the small value of $S_{1,19}$, one can see that the opposite side blocking of this architecture is not a big issue. This is because the monopole effective aperture size is much smaller than the lens aperture size. Therefore, majority of the power will be radiated out without any interference from the opposite side elements blockage. This small mutual coupling effect between different feeding elements leads to a much simpler and convenient procedure in beam synthesis and antenna impedance matching.
Figure 5. Simulation setup for 36 feed elements on the surface of a 12-cm diameter Luneburg lens.
Figure 5-6. Simulated S-parameters of the 36 dipoles mounted on the surface of a Luneburg lens as shown in Figure 5-5.

5.3 Beam Synthesis

Unlike the traditional phased array that all the radiation elements have the same radiation pattern, different elements around the lens has different radiation patterns since they are facing different directions. To implement beam synthesis for this Luneburg lens based phased array, a pseudoinverse matrix method is used to find the minimum squared error solution for the desired pattern. The procedure is as following:
1) Determine Lens size from gain requirement

As a typical lens antenna, the gain of Luneburg lens has a relationship with its size as shown in Eq. (5-2) [48]. The maximum gain of the proposed Luneburg lens based electronically scanning array has the same value as the Luneburg lens antenna itself. The pattern of the feeding element will influence the array side lobes, but has limited impact on the maximum gain value. Therefore, using Eq. (5-2), the lens aperture size A can be determined from the gain requirement of the system.

\[ G = \eta \frac{4\pi}{\lambda^2} A \]  \hspace{1cm} (5-2)

Here \( \eta \) is the aperture efficiency of the antenna. From our previous measured results at X-band [22] (a lens with a diameter of 12 cm, 19 dB gain at 10 GHz), the aperture efficiency \( \eta \) for the Luneburg lens is about 50% with a waveguide feed.

After the Luneburg lens size is determined, one can roughly estimate its half power beam width (HPBW) using Eq. (5-3) [170]:

\[ \theta_{\text{HPBW}} = 1.1 \times \frac{\lambda}{D} \]  \hspace{1cm} (3)

Here \( \lambda \) is the free space wavelength and D is the diameter of the lens. The value 1.1 used here is also from the previous measured results [22] (a lens with a diameter of 12 cm, 15-degree HPBW at 10 GHz).
2) Determine the number of elements, their placement and the number of controllers

To achieve arbitrary direction beam scanning using the Luneburg lens array, the angular separation $\Delta \theta$ between two adjacent elements needs to be approximately HPBW or smaller. If the distance between two adjacent elements is too large, it will be difficult to scan the beam in between the opposite directions of these two elements. After the angular distance $\Delta \theta$ is selected, the total number of elements is determined by the requirement of scanning range $\theta_{\text{scanning}}$ using Eq. (5-4):

$$\text{Number of elements} = \frac{\theta_{\text{scanning}}}{\Delta \theta} \quad (5-4)$$

From Eqs. (5-2) and (5-3), one can see that a larger lens size leads to a higher gain and a narrower HPBW. Narrower beam width means a smaller angular distance is required and therefore more elements are needed to scan the same range.

As mentioned before, although this Luneburg lens array may need a large number of feeding elements mounted around the lens to satisfy the scanning range requirement, only very few (i.e., 3 to 5) elements are required to be excited at the same time. The exact number of elements working at the same time can be determined by the allowed system
cost and complexity and the scanning accuracy desired. The more number of elements one can use, the higher scanning accuracy one can achieve.

3) Obtain feed elements excitation distribution

For conventional phased array, the radiation pattern for an identical \( N \)-element array can be written as [171]:

\[
\text{Pattern}_{\text{total}} = \text{Pattern}_{\text{single}}(\theta) \ast \text{ArrayFactor}
\]

\[
\text{ArrayFactor} = a_0 + a_1 e^{j\psi_1} + a_2 e^{j\psi_2} + a_3 e^{j\psi_3} \ldots + a_{N-1} e^{j\psi_{N-1}}
\]

in which \( a_i \) is the amplitude of the \( i \)th element, \( \psi_i \) is the phase difference between that element and the reference element (\( \psi_i = ikd\cos(\theta) + \beta_i \) for linear 1D array [171]). However, for this Luneburg lens based phased array, the radiation patterns of different elements are no longer identical since they are placed on different positions of the lens. Therefore, the Luneburg lens array can be viewed as a phased array with heterogeneous elements. In this case, the total radiation pattern can be written as:

\[
\text{Pattern}_{\text{total}} = a_0 \text{Pattern}_0(\theta) e^{j\psi_0(\theta)} + a_1 \text{Pattern}_1(\theta) e^{j\psi_1(\theta)} + a_2 \text{Pattern}_2(\theta) e^{j\psi_2(\theta)} \ldots + a_{N-1} \text{Pattern}_{N-1}(\theta) e^{j\psi_{N-1}(\theta)}
\]

\[
\psi_i = \Delta\phi_i + \beta_i
\]

The phase difference \( \psi_i \) here has two parts, \( \Delta\phi_i \) is the phase difference due to spatial distribution (similar like \( kd\cos(\theta) \)), and \( \beta_i \) is the source current phase difference from
different elements. The pattern of different elements and $\Delta \phi_i$ are related to the property of the Luneburg lens which can be determined either from simulation or from measurement.

The amplitude $a_i$ and phase difference $\beta_i$ are controllable parameters of the source. By controlling these values, different radiation patterns can be synthesized for different applications.

By separating the controllable and fixed parameters, Eq. (5-7) can be rewritten as:

$$\text{Pattern}_{\text{total}}(\theta) = \text{Pattern}_0(\theta)e^{j\lambda \phi_0(\theta)}a_0 + (\text{Pattern}_1(\theta)e^{j\lambda \phi_1(\theta)}) \ast (a_1 e^{j\beta_1})$$

$$+ (\text{Pattern}_2(\theta)e^{j\lambda \phi_2(\theta)}) \ast (a_2 e^{j\beta_2})$$

$$+ \ldots$$

$$+ (\text{Pattern}_{N-1}(\theta)e^{j\lambda \phi_{N-1}(\theta)}) \ast (a_{N-1} e^{j\beta_{N-1}})$$

$$= \begin{bmatrix}
\text{Pattern}_0(\theta_1)e^{j\lambda \phi_0(\theta_1)}, \text{Pattern}_1(\theta_1)e^{j\lambda \phi_1(\theta_1)}, \text{Pattern}_2(\theta_1)e^{j\lambda \phi_2(\theta_1)}, \ldots, \text{Pattern}_{N-1}(\theta_1)e^{j\lambda \phi_{N-1}(\theta_1)}

\text{Pattern}_0(\theta_2)e^{j\lambda \phi_0(\theta_2)}, \text{Pattern}_1(\theta_2)e^{j\lambda \phi_1(\theta_2)}, \text{Pattern}_2(\theta_2)e^{j\lambda \phi_2(\theta_2)}, \ldots, \text{Pattern}_{N-1}(\theta_2)e^{j\lambda \phi_{N-1}(\theta_2)}

\ldots

\ldots

\text{Pattern}_0(\theta_y)e^{j\lambda \phi_0(\theta_y)}, \text{Pattern}_1(\theta_y)e^{j\lambda \phi_1(\theta_y)}, \text{Pattern}_2(\theta_y)e^{j\lambda \phi_2(\theta_y)}, \ldots, \text{Pattern}_{N-1}(\theta_y)e^{j\lambda \phi_{N-1}(\theta_y)}
\end{bmatrix} \\
\times \\
\begin{bmatrix}
a_0 \\
a_1 e^{j\beta_1} \\
a_2 e^{j\beta_2} \\
\ldots \\
a_{N-1} e^{j\beta_{N-1}}
\end{bmatrix}$$

$$= [P] \times [A]$$

(5-8)

Here, matrix $[\text{Pattern}_{\text{total}}]$ is the desired pattern, matrix $[P]$ is obtained from the property of the Luneburg lens with certain feeding elements and matrix $[A]$ is the excitations of all the elements to achieve the designed pattern. Since $P$ is usually not an $N \times N$ matrix, Eq. (5-8) may not have an exact solution for excitation $A$. However, the pseudoinverse matrix...
of $[P]$ can be used to find the minimum squared error solution for the desired pattern:

$$[A] = \text{Pseudoinverse}([P]) \times \text{Pattern}_{\text{total}}$$

(5-9)

In the following section, some beam synthesize examples using the simulated pattern of a 12-cm diameter Luneburg lens are shown.

### 5.4 Pattern Synthesis Examples

The pattern amplitude and phase of a 12-cm diameter Luneburg lens is simulated in HFSS by mounting 36 small dipoles with 10-degree angular spacing around the lens as shown in Figure 5-5. The dipole length is 10 mm and a lumped port with impedance matched to the dipole is employed as the feed. With an incident plane wave, different elements on the lens will receive different voltages. From the simulated magnitude and phase of these voltages and using the symmetry of Luneburg lens, the single element pattern and phase difference information can be obtained. Figure 5-7 plots the normalized radiation pattern of dipole 0 (located at angle 0-degree) at 10 GHz with all other feeding elements matched. Figure 5-8 is the received phase for different incident angles. With the simulated results in Figure 5-7 and 5-8, and using the symmetry of lens, matrix $[P]$ can be obtained. For example, $\text{Pattern}_0(\theta)$ are those values in Figure 5-7, $\Delta\phi_0(\theta)$ are the values in Figure 5-8. $\text{Pattern}_1(\theta)$ and $\Delta\phi_1(\theta)$ will be $\text{Pattern}_0(\theta+10^\circ)$ and $\Delta\phi_0(\theta+10^\circ)$ due to the symmetry of the lens, and so on. Next, the excitation $[A]$ of the
feeding elements can be solved using Eq. (5-9) to obtain desired patterns.

Figure 5- 7. Simulated single element pattern of a 12-cm diameter Luneburg lens surrounded by 36 small dipoles (10 mm length) at 10 GHz.

Figure 5- 8. Simulated phase difference information for different incident angles.
5.4.1 Horizontal plane beam scanning

For regular beam scanning application in the horizontal plane, the desired pattern is set to be the single element pattern horizontally shifted by an angle $\Delta \theta$. Then after solving Eq. (5-9), the minimum squared error solution of the necessary magnitude and phase distribution can be obtained. The magnitude and phase distribution of 5 adjacent feeding elements (located at 340°, 350°, 0°, 10°, 20°) for scanning angles $\Delta \theta = 2°$, 5° and 8° are shown in Figure 5-9. From this minimum squared error solution $[A]$, the actually achieved pattern is calculated using $[P] \times [A]$ and plotted in Figure 5-10 for $\Delta \theta$ from 0 to 10 degrees with 1-degree scanning step. It can be clearly seen that with these 5 adjacent elements, radiation patterns very similar to the single element pattern can be achieved from 0 to 10 degree. To investigate the number of elements needed for the fine beam scanning, the achieved patterns with only 3 adjacent elements are shown in Figure 5-11. It can be observed that although the obtained maximum gain values have slightly more deviation than the results with 5 adjacent elements, it is still satisfactory. To implement beam scanning more than 10 degrees, different set of feeding elements can be selected (e.g., elements located at 350°, 0°, 10°, 20°, 30° will cover 10 to 20 degrees) and the magnitude and phase of the elements would be the same because of the symmetry of the
lens.

\[ \Delta \theta = 2 \text{ degree} \]

\[ \Delta \theta = 5 \text{ degree} \]

\[ \Delta \theta = 8 \text{ degree} \]

Figure 5-9. Magnitude and phase distribution of 5 adjacent feeding elements (located at 340\(^\circ\), 350\(^\circ\), 0\(^\circ\), 10\(^\circ\), 20\(^\circ\)) to realize beam scanning to 2\(^\circ\), 5\(^\circ\), and 8\(^\circ\).
The radiation patterns in Figure 5-10 and 5-11 indicate that the gain of this electronic scanning Luneburg lens is almost constant which is determined by the lens itself (within 0.5 dB) for the entire scanning range. This scan angle independent performance represents a key advantage compared to the traditional phased array which suffers the beam deformation effect.

Figure 5-10. Achieved scanning pattern from 0 to 10 degrees with 5 adjacent elements.
Figure 5-11. Achieved scanning pattern from 0 to 10 degrees with 3 adjacent elements.

In addition, compared to the traditional phased array, the size of this Luneburg lens based system is almost the same because the gain is limited in both cases by the effective aperture and wavelength as shown in Eq (5-2). However, unlike the traditional phased array that a large number of elements are needed, the number of elements needed for the Luneburg lens array is much smaller. If only a narrow scanning range is required, very limited number of elements is sufficient. Even if a large scanning range is desired, only a few elements need to be excited simultaneously and a switch matrix can be utilized to
select the appropriate set of elements. Therefore the number of phase shifters and control units is much smaller which will lead to greatly reduced system complexity and cost compared to the traditional phased array.

5.4.2 Fan beam & null beam forming

Other than the regular beam scanning application, this Luneburg lens electronic scanning array can also be used to realize other patterns such as fan beam or differential beam by applying different element magnitude and phase excitation \([A]\).

To achieve a fan beam, the desired pattern in Eq. (5-9) can be set to a constant value in the desired beam width range and to 0 outside the beam width range. Using the same procedure, the element magnitude and phase excitation \([A]\) can be solved and the achieved pattern can be calculated with \([P]\) x \([A]\). The results of fan beam examples with beam width 60, 90 and 150 degrees are shown in Figure 5-12. It can be seen that almost constant gain is obtained within the beam width and the gain outside the beam is smaller than -20 dB.
Figure 5-12. Achieved fan beam patterns with 60, 90 and 150 degree beam width and the excitation magnitude and phase distributions of the 36 feeding elements.

With this Luneburg lens based beam scanning system, a null beam forming can also be achieved to block the potential interference signal in certain directions. One example of the null beam forming is shown in Figure 5-13. The main beam direction is set to 180° with a null from 30° to 70°. The result shows a smaller than -30 dB null is achieved in the desired angular range.
5.4.3 3D pattern synthesis

Using the proposed Luneburg lens array, not only 1D beam scanning, but also 2D beam scanning can be realized by positioning feeding elements over a 2D area on the lens surface instead of only in the horizontal plane as discussed previously. For the 2D case, beam synthesis can be done by replacing Pattern(\(\theta\)) in Eq. (5-8) by Pattern(\(\theta, \varphi\)) as the following.
Pattern_{total}(\theta, \phi) = \text{Pattern}_0(\theta, \phi)e^{j\Delta\phi_0(\theta, \phi)}a_0 + (\text{Pattern}_1(\theta, \phi)e^{j\Delta\phi_1(\theta, \phi)}) * (a_1e^{j\beta_1}) \\
+ (\text{Pattern}_2(\theta, \phi)e^{j\Delta\phi_2(\theta, \phi)}) * (a_2e^{j\beta_2}) \\
+ \cdots \\
+ (\text{Pattern}_{N-1}(\theta, \phi)e^{j\Delta\phi_{N-1}(\theta, \phi)}) * (a_{N-1}e^{j\beta_{N-1}})

\begin{bmatrix}
\text{Pattern}_0(\theta_1, \varphi_1)e^{j\Delta\phi_0(\theta_1, \varphi_1)}, \text{Pattern}_1(\theta_1, \varphi_1)e^{j\Delta\phi_1(\theta_1, \varphi_1)}, \ldots, \text{Pattern}_{N-1}(\theta_1, \varphi_1)e^{j\Delta\phi_{N-1}(\theta_1, \varphi_1)} \\
\text{Pattern}_0(\theta_1, \varphi_M)e^{j\Delta\phi_0(\theta_1, \varphi_M)}, \text{Pattern}_1(\theta_1, \varphi_M)e^{j\Delta\phi_1(\theta_1, \varphi_M)}, \ldots, \text{Pattern}_{N-1}(\theta_1, \varphi_M)e^{j\Delta\phi_{N-1}(\theta_1, \varphi_M)} \\
\text{Pattern}_0(\theta_2, \varphi_1)e^{j\Delta\phi_0(\theta_2, \varphi_1)}, \text{Pattern}_1(\theta_2, \varphi_1)e^{j\Delta\phi_1(\theta_2, \varphi_1)}, \ldots, \text{Pattern}_{N-1}(\theta_2, \varphi_1)e^{j\Delta\phi_{N-1}(\theta_2, \varphi_1)} \\
\text{Pattern}_0(\theta_2, \varphi_M)e^{j\Delta\phi_0(\theta_2, \varphi_M)}, \text{Pattern}_1(\theta_2, \varphi_M)e^{j\Delta\phi_1(\theta_2, \varphi_M)}, \ldots, \text{Pattern}_{N-1}(\theta_2, \varphi_M)e^{j\Delta\phi_{N-1}(\theta_2, \varphi_M)} \\
\text{Pattern}_0(\theta_L, \varphi_1)e^{j\Delta\phi_0(\theta_L, \varphi_1)}, \text{Pattern}_1(\theta_L, \varphi_1)e^{j\Delta\phi_1(\theta_L, \varphi_1)}, \ldots, \text{Pattern}_{N-1}(\theta_L, \varphi_1)e^{j\Delta\phi_{N-1}(\theta_L, \varphi_1)} \\
\text{Pattern}_0(\theta_L, \varphi_M)e^{j\Delta\phi_0(\theta_L, \varphi_M)}, \text{Pattern}_1(\theta_L, \varphi_M)e^{j\Delta\phi_1(\theta_L, \varphi_M)}, \ldots, \text{Pattern}_{N-1}(\theta_L, \varphi_M)e^{j\Delta\phi_{N-1}(\theta_L, \varphi_M)}
\end{bmatrix}

= \begin{bmatrix} a_0 \\ a_1e^{j\beta_1} \\ a_2e^{j\beta_2} \\ \vdots \\ a_{N-1}e^{j\beta_{N-1}} \end{bmatrix} = [P] \times [A]

(5-10)

Then using the same procedure as the 1D case, the element excitation magnitude and phase distribution [A] can be solved. Figure 5-14 shows several examples scanning to different directions using 25 elements (5 x 5) covering an area from \( \theta = 70^\circ \) to \( 110^\circ \) and \( \phi = -20^\circ \) to \( 20^\circ \) with 10° angular spacing in \( \theta \) and \( \phi \). It can be observed that good radiation patterns are achieved when this system is scanning in the opposite direction of the region covered by the elements. All the advantages discussed for the 1D scanning case can be
applied to the 2D scanning.

Figure 5-14. Achieved 2D normalized radiation patterns scanning to different directions with 25 elements located within an area from $\theta = 70^\circ$ to $110^\circ$ and $\varphi = -20^\circ$ to $20^\circ$ with $10^\circ$ angular spacing in $\theta$ and $\varphi$.

5.5 Conclusion

A novel broadband electronic scanning array based on Luneburg lens is proposed and studied in this chapter. By controlling the excitation amplitude and phase of different
elements located on the surface of the lens, 1D and 2D versatile electronic beam scanning can be realized. Compared to traditional phased array systems, this new electronic scanning approach has several advantages such as broadband, no scan angle limit, no beam deformation effect when scanning to different directions, and much reduced system complexity and cost to achieve a highly directional pattern.
CHAPTER 6. THZ CHARACTERIZATION OF CARBON BASED NANO-MATERIALS

6.1 Introduction

Carbon nanotubes (CNT) are rolled-up graphene sheets with hollow cylindrical geometry, which are classified as single-walled (one graphene layer) and multi-walled (multiple graphene layers) nanotubes. A single-walled carbon nanotube (SWNT) can be either metallic or semiconducting, depending on its chirality [172]. Because of their superb mechanical and electrical properties, there have been extensive interest and research efforts on CNTs as nano-scale circuit building blocks, as well as on numerous potential applications in the areas of field emission displays, microscopy and scanning / tunneling microscope tips, fuel cells and batteries [173]. Many microwave and Terahertz (THz) applications have also been suggested, such as antennas and interconnect [174,175,176].

Recently, isolated graphene, two-dimensional flat monolayer honeycomb lattice composed of carbon atoms, has been discovered and appears to be a more promising material than CNT because of its comparable electrical and mechanical properties with CNT and its potential to be fabricated macroscopically (~ cm wide graphene sheets of only several nanometers thick) with newly developed techniques [177]. Graphene is also
believed to have interesting nonlinear frequency multiplication effects in the Terahertz frequency range [178,179,180,181].

In order to verify and enable many proposed applications of these carbon-based nano-structures, it is essential to understand their electrical and optical properties. Despite extensive studies performed at DC, low frequencies and optical frequencies [173], the electrical properties of CNTs over the microwave and Terahertz regimes have not yet been well studied [182, 183]. Direct characterization of individual nanotubes in this frequency region is impeded by the practical difficulty of test fixture fabrication and the large impedance mismatch between single nanotube and testing ports. An alternative approach is to characterize a large ensemble of nanotubes. This way the collective response of carbon nanotubes is much stronger and easier to measure. With the Terahertz Time-domain Spectroscopy (TDS), either free-standing carbon nanotube paper or thin CNT film deposited onto substrates can be characterized by transmission and / or reflection measurements. Transmission characterization is easier to implement because it does not require precise alignment of the sample surface position with the reference [183].

In [183], the multi-wall carbon nanotubes (MWNT) sample studied has a thickness of around 90 µm, which was treated as a bulk material. Given the THz test equipment dynamic range, thin-film sample measurement has better signal-to-noise ratio (SNR)
because of its smaller CNT layer thickness and lower tube density than tubes in a free-standing CNT paper. Other benefits of thin-film samples include more uniform tube distribution, as well as the ease of gating electrodes fabrication onto the substrate if nonlinear measurements are pursued. Moreover, carbon based thin films are much more promising nano-material candidates than bulk MWNT for a variety of applications such as thin film transistors [184], so that THz characterization of them would be significant and useful.

In this chapter, SWNT thin films on glass substrates with thicknesses ranging from 70 nm to 300 nm and an on-substrate graphene sample (2 to 3 layers) are characterized via Terahertz Time-domain Spectroscopy. These thin films are treated as a surface boundary at the substrate-air interface rather than a bulk material. Characterization of these thin films is quite more challenging compared to the bulk MWNT samples studied in [182] due to the potential influences of substrate (thickness and dielectric property). By applying an iteration process, a more precise way to determine the substrate thickness is applied to improve the accuracy [185,186,187,188]. Numerical technique attempted to account for substrate Fabry-Pérot effects are also applied, although it is proven to be imperfect in the results of the first batch of SWNT samples. The substrate thicknesses of the second batch of SWNT samples are then chosen so that multiple reflections within the substrates are circumvented by truncation of the time domain signals in data analysis. The
SWNT results show consistent surface conductivities for samples on different substrates and with different film thicknesses. The resulting graphene conductivity in Terahertz frequency is quite comparable to the values reported in the literature for graphene at DC and optical frequency. Also, this surface conductivity characterization technique is successfully applied as a means to evaluate metallic content of CNTs samples [189].

This chapter is outlined as follows. The sample fabrication process is first presented. The Terahertz time-domain spectroscopy and the associated thin film property extraction procedures are discussed. The measured SWNT and graphene thin film properties and uncertainty analysis are then presented. This Terahertz thin film characterization technique is finally used to evaluate metallic SWNT content in a material purification process.

6.2. Sample Fabrication

The single-walled nanotubes studied are commercially available SWNT synthesized by high-pressure CO conversion (HiPCO) process [190, 191]. SWNT powder is first dispersed in 1 wt% of sodium dodecyl sulfate (SDS) solution via ultrasonication treatment, and centrifuged at 25000 G for 2 hours to remove catalyst particles. Then, the SWNT suspension is filtrated through 200 nm Millipore polycarbonate membrane. A layer of SWNT thin film is formed on the membrane and the SDS is washed away by
excessive de-ionized water. The filtration membrane is then transferred onto a glass or quartz substrate and immersed in chloroform bath for 6 hours to remove the membrane. Resulting SWNT thin film samples on substrates (Figure 6-1(a)) are dried at 75°C for 3 hours. The photograph, atomic force micrograph (AFM) and scanning electron micrograph (SEM) of a CNT thin film sample on a glass substrate are shown in Figure 6-1.

![Figure 6-1](image)

Figure 6-1. (a) Photograph, (b) AFM image, and (c) SEM image of a thin film SWNT sample on glass (from [189]).

The graphene sample was directly synthesized on copper foil using liquid precursor hexane in chemical vapor deposition system [192]. As supporting layer, a thin PMMA film was deposited on the graphene/Cu substrate for transferring. After that, the underlying Cu substrate was dissolved in dilute HNO₃, and the film was transferred onto a target glass substrate and acetone was used to remove PMMA from the sample, only leaving graphene on substrate for further characterization.
6.3 Thin Film Terahertz Characterization

Terahertz Time-domain Spectroscopy is well suited for the characterization of SWNT and graphene films because of its high signal-to-noise ratio in its frequency range [105]. A photoconductive Terahertz-TDS system from Picometrix Inc. is employed to characterize the SWNT and graphene films. As shown in Figure 6-2, a Terahertz pulse is generated by biased coplanar lines on a low-temperature grown GaAs substrate, under the excitation of a femto-second laser. The same femto-second laser pulse is also guided to the detector through an optical delay line as the gating signal for recording the received Terahertz waveform. The detector is a 5-μm gap dipole antenna, which is also fabricated on a low-temperature grown GaAs substrate. The sample under test is placed in the Terahertz pulse beam path between the emitter and the detector. The measured time-domain response is then transformed into frequency domain. Because the measurement is coherent, both the magnitude and phase of the sample responses are obtained at the same time. The complex transmission coefficient of the sample is obtained by dividing the sample transmission spectrum by the reference spectrum taken without the sample in the beam path. The sample material properties can then be extracted from the measured complex transmission coefficient. In this work, bare substrates and films on substrate are characterized to obtain intrinsic film properties.

The measured time domain pulses of a SWNT thin film on glass sample together
with a reference pulse are shown in Figure 6-3(a). The dashed and solid lines are the reference and the transmitted pulses, respectively. Figure 6-3(b) plots the magnitude of the Fourier transformed signals in the frequency domain. The normalized sample transmission coefficient is obtained by taking the ratio between these two corresponding spectra.

Figure 6- 2. Terahertz Time-domain Spectroscopy (TDS) characterization setup.
Figure 6-3. Terahertz Time-domain Spectroscopy (TDS) measurement results: (a) transmission waveforms of the reference (dashed line) and a SWNT thin film sample (solid line); (b) The reference signal and thin film sample signal in frequency domain.

6.4. Thin Film Property Extraction and Results

6.4.1. Algorithm

Given that the film thickness of SWNT (on the order of 100 nm) and graphene (2 - 3 layers, \( \leq 1 \text{ nm} \)) is much less than the Terahertz wavelength (on the order of 1 mm), the incident Terahertz wave can be assumed to be uniform within the film. Therefore, instead
of treating the film as a block of material [183, 193, 194], it can be regarded as a boundary condition with a surface conductivity [105]. As illustrated in Figure 6-4, at the air-film interface, the reflection coefficient $R_i$ and transmission coefficient $T_i$ can be written as [105]:

$$R_i = \frac{Y_- - \sigma_s}{Y_+ + \sigma_s} \quad (6-1)$$

$$T_i = \frac{2Y_1}{Y_+ + \sigma_s} \quad (6-2)$$

where $\sigma_s$ is the surface conductivity of the thin film; $Y_+$ and $Y_-$ are functions of $Z_1$, $Z_2$ and $Z_0$ as given in Eqs. (6-3) and (6-4), with $Z_1$, $Z_2$ and $Z_0$ being the wave impedances in medium 1 (air), medium 2 (substrate) and free space; $n_1$ and $n_2$ are the indices of refraction of air and the substrate, respectively.

$$Y_\pm = Y_1 \pm Y_2 \quad (6-3)$$

$$Y_{1,2} = \frac{1}{Z_{1,2}} = \frac{n_{1,2}}{Z_0} \quad (6-4)$$

Due to the impedance mismatch between air and the substrate, multiple reflections of the Terahertz pulse occur within the substrate underneath the thin film. These reflections would interfere with the main pulse and appear as ripples in the resulting sample transmission spectrum, which is called the Fabry-Pérot effect. Taking multiple reflections
into consideration, the frequency ($\omega$) dependent transmission coefficient $T$ of the whole sample can be written as:

$$
T(\omega) = \frac{4X \cdot n_{\text{sub}}}{n_{\text{sub}} + 1} \exp(-j(n_{\text{sub}} - 1)k_0 \ d_{\text{sub}}) \\
\cdot \sum_{FP=0}^{N} (\exp(-2jn_{\text{sub}}k_0 \ d_{\text{sub}}) \frac{n_{\text{sub}} - 1}{n_{\text{sub}} + 1} (2n_{\text{sub}} \ X - 1))^{FP}
$$

(6-5)

$$
X^{-1} = 1 + n_{\text{sub}} + \sigma_s Z_0
$$

(6-6)

![Figure 6-4](image)

Figure 6-4. Schematic of the thin film sample: the thin film under study is treated as a surface-conductivity boundary between air and the substrate.

Herein $n_{\text{sub}}$ and $d_{\text{sub}}$ are the substrate complex refractive index and thickness, respectively; $k_0$ is the free-space wave number; $Z_0 = 377 \ \Omega$ is the free-space impedance; $\sigma_s$ is the effective complex surface conductivity of the thin film, with a unit of Siemens/square; and $FP$ denotes the number of multiple reflections of the signal within the substrate. If multiple reflections are included in the calculation, $FP$ would be an integer $N$ and
Equation (6-5) becomes a \((N+1)\) order complex polynomial of \(X\). Numerical fitting techniques are then necessary to solve for \(X\). However, within the time window applied in the calculation, the number of multiple reflections taken into account could very well be a non-integer. Therefore this numerical process could not account for the Fabry–Pérot effect exactly. Nevertheless, if the substrate is thick enough, the multiple reflections could be excluded by proper waveform truncation, thus \(FP\) would be equal to zero. In that case, with known substrate properties, the \(X\) and thus \(\sigma_s\) could be calculated analytically from the transmittance \(T\) using the following equation:

\[
\sigma_s = \left[ \frac{4n_{sub}}{T \cdot (n_{sub} + 1)} \exp \left( -j(n_{sub} - 1)k_0 \cdot d_{sub} \right) - 1 - n_{sub} \right] / 377 \quad (6-7)
\]

### 6.4.2. Substrate Characterization

From (6-5), it can be seen that the refractive index of the substrate \(n_{sub}\) is an important parameter that influences the transmission coefficient. Therefore, the refractive index of the bare substrate should be measured first before the surface conductivity of the thin film could be extracted.

The bare glass substrate transmission spectrum is measured using the Terahertz-TDS to extract its dielectric properties. The permittivity of the air is assumed to be 1. Then using the transmission spectrum equation in Ref [187]
\[ T(\omega) = \frac{4 n_{\text{sub}}}{(n_{\text{sub}} + 1)^2} \exp(-j(n_{\text{sub}} - 1)k_0 d_{\text{sub}}) \]
\[ \cdot \sum_{FP=0}^{N} \left( \exp(-2j n_{\text{sub}} k_0 d_{\text{sub}}) \left( \frac{n_{\text{sub}} - 1}{n_{\text{sub}} + 1} \right)^{FP} \right) \]

(6-8),

the complex index of refraction \( n_{\text{sub}} \) could be extracted numerically. In (6-8), \( d_{\text{sub}} \) is the substrate thickness; \( k_0 \) is the free-space wave number; and \( FP \) denotes the multiple reflections number within the substrate that can be calculated by the truncation window employed in the Fourier transformation. In the Fourier transformation, the time window is set to be from 6 picoseconds before the main peak to 24 picoseconds after the main peak. According to the substrate refractive index of 2.4 (roughly estimated using \( FP = 0 \)) and thickness of 170 \( \mu \)m, the number of multiple reflections within the truncated waveform is estimated to be 8. The numerical process starts with an initial value of \( N_0 = 3 \). The solution of \( n_{\text{sub}} \) for \( FP = N_0 + 1 \) case is then obtained. After several iterations, the solution converges, meaning that any higher order reflected signals have little effect on the solution. The refractive index is then calculated using the \( FP \) in the last iteration.

In the extraction, the substrate thickness is initially set to 170 \( \mu \)m, which is measured by a vernier caliper. However, the accuracy of the vernier caliper (10 \( \mu \)m) is not enough in our extraction. To determine the sample thickness more accurately, we employ the total variation technique [185]. In general, an error in \( d_{\text{sub}} \) is noticeable in the extracted result of \( n_{\text{sub}} \) as an oscillation in frequency [186] [188]. So, the substrate refractive index versus
frequency curves are extracted using a range of values of substrate thickness. By applying a smoothness characterization for these different curves, the smoothest frequency response for both \( n' \) and \( n'' \) (the real and imaginary parts of \( n_{\text{sub}} \)) can be identified. Then, the corresponding thickness is the closest thickness to the reality among the values in the considered range. The smoothness of the curves defined here is based on the absolute difference between the nearby frequencies in the interested frequency range. For a given substrate thickness \( d_{\text{sub}} \), the total variation is given by Eq. (6-9):

\[
D(m) = |n'(m - 1) - n'(m)| + |n''(m - 1) - n''(m)| \\
Total\ Vari\ a\ i\ t\ i\ o\ n(d_{\text{sub}}) = \sum D(m)
\]

Determining the substrate thickness is equivalent to finding the \( d_{\text{sub}} \) that the total variation is minimal.
Figure 6-5 plots the total variation of the substrate index of refraction with different thicknesses for the SWNT sample. It can be seen that the lowest total variation is at the thickness of 174 μm. Using this more precise thickness, the complex refractive index of the glass substrate could be extracted using (6-8). The result of the SWNT sample glass substrate is shown in Figure 6-6.
Figure 6. The measured complex refractive index $n$ of the 170-μm bare glass substrate of the SWNT sample with statistical error bars: (a) Real part. (b) Imaginary part.

It can be seen that the glass substrate used here has a flat real part of refractive index around 2.42 from 150 to 750 GHz, and an imaginary part of refractive index increasing with frequency, from 0.02 at 150 GHz to 0.14 at 750 GHz.

Applying the same method used in characterizing the SWNT sample glass substrate, the complex refractive index of the bare glass substrate used for the graphene sample is also extracted. For this sample, the thickness of the substrate is about 1 mm. Therefore, according to the complex refractive index and thickness of the substrate, only the first-order multiple reflection is included in the time window, so $FP$ equals to 1 is assumed in the numerical fitting process to extract $n_{sub}$. 

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The measured refractive index of substrate of the graphene sample is shown in Figure 6-7. The real refractive index of the bare substrate is around 2.62 from 160 to 720 GHz, while the imaginary part of the refractive index increases with frequency, from 0.02 at 160 GHz to 0.13 at 720 GHz.

![Graph showing real and imaginary parts of refractive index](image)

Figure 6-7. The measured complex refractive index of the 1-mm bare glass substrate of the graphene sample with statistical error bars: (a) Real part. (b) Imaginary part.

### 6.4.3. SWNT Film Surface Conductivity

After the complex refractive index of the glass substrate is obtained, the transmitted pulse waveforms of two SWNT samples residing on the 170-μm thick substrate are measured. One film is approximately 70 nm thick, while the other one is about 3 to 4
times thicker. The solid line in Figure 6-3 is the untruncated transmission pulse for the 70-nm SWNT film on the 170 μm substrate. Except the main pulse, no apparent higher-order reflection pulse are observed in the transmitted waveform through the thin SWNT film sample, which indicates the main pulse and the higher-order reflection pulses overlap because of the small substrate thickness. Also, according to the averaged glass substrate refractive index in Figure 6-6 and the thickness of the sample, the number of multiple reflections within the truncated waveform is estimated to be 8. Therefore $FP = 8$ is applied in (6-5), and the surface conductivity $\sigma_s$ is then extracted using the numerical fitting process with the measured $n_{\text{sub}}$ data in Figure 6-6. Again, the similar minimum total variation method is used in determining the precise substrate thickness (in this case, $\sigma_s$ are calculated at different thicknesses). After that, the surface conductivities $\sigma_s$ of the SWNT films are finally extracted. It is worth to note that both the phase and magnitude terms of the measured transmission coefficient contribute into the extracted surface conductivity so that it is a complex quantity here. And the imaginary part of the surface conductivity represents the dielectric term of the material property. Because the films are mainly metallic, the real part of the surface conductivity dominates (the imaginary part of the surface conductivity is an order of magnitude smaller than the real part).

The results of surface conductivity (real part) are plotted in Figure 6-8, in which the Orientation 1 and Orientation 2 are the same sample tested at different angles with
respect to the polarization of the Terahertz pulse. One is rotated by 90-degree to the other.

It can be seen that the measured surface conductivities are almost the same for these two orientations for both the thick and thin film samples. This indicates that the nanotubes are randomly oriented in the film samples, which is expected from the film fabrication process and consistent with the sample micrographs shown in Figure 6-1. Moreover, it can be seen that the surface conductivity $\sigma_s$ of the 70-nm thin film can almost be scaled to that of the thicker film by a factor of ~ 3.7, which is fairly consistent with the expected thickness ratio between the films. Therefore, the two films have about the same volume conductivities, on the order of 500 Siemens/cm over the 100 GHz to 700 GHz frequency range. These results indicate that the solution-deposition method could fabricate uniform carbon nanotube thin films with controllable thickness on the order of hundreds of nanometers.
Figure 6-8. Extracted real surface conductivities of SWNT films on 170-μm glass substrate with error bars. The circles and crosses are the thick SWNT film with different orientation (rotation by 90-degree) respect to the incident Terahertz wave. The triangles and squares are the thin SWNT film with different orientation respect to the incident Terahertz wave.

It is worth noting that the extracted surface conductivities in Figure 6-8 show local minima around 180 GHz and 520 GHz (more obvious for the thicker sample). This is likely due to the residual effects of substrate multiple reflections, which cannot be totally eliminated by the iterative numerical process applied here because of the non-integer $FP$. 
problem stated previously.

However, if the substrate is thick enough so that the main pulse could be well separated in time domain from the first multiple-reflection pulse, then proper truncation of the transmitted waveform can be applied to eliminate the multiple reflected pulses while the main pulse integrity is left intact. In this case, the multiple reflections number $FP$ would be equal to zero and $\sigma_s$ could be calculated analytically from the transmittance $T$ using (6-7). Meanwhile, the substrate thickness should not be too large either to prevent substantial loss in the substrate which would degrade the measurement dynamic range. Under these considerations, another batch of SWNT samples are fabricated with specially selected substrate thickness.

For this second batch of samples, three much thicker substrates are used, including: a 3.2 mm Pyrex brand glass, a 3.2 mm window glass and the same window glass with a 5 mm thickness. The dielectric constant of the Pyrex glass is measured to be around 4.38, and its loss tangent increases from 0.007 at 100 GHz to 0.046 at 800 GHz. The window glass has similar refractive index properties with the glass substrates in the first batch. All three films are deposited with similar thicknesses around 70 nm. With proper time-domain truncation, multiple reflections are excluded, and the film surface conductivity $\sigma_s$ could be analytically calculated by (6-7). Measured results of all three samples are presented in Figure 6-9. The samples yield comparable surface conductivity,
which is expected since the CNT solution is identical and the film thicknesses are comparable. It shows an almost monotonic increase of surface conductivity with frequency from 50 GHz to 400 GHz. These results of the SWNT films are consistent with the surface conductivity of the SWNT film deposited on the thin substrate shown in Figure 6-8.

![Surface Conductivity vs Frequency](image)

Figure 6-9. Real surface conductivities of three SWNT films on thick substrates as a function of frequency.

In Figure 6-9, no periodic oscillation is observed over the band, indicating that the multiple reflection effect is excluded. However, the measured surface conductivities are only valid up to 400 GHz because of the increased attenuation in substrates. Substrate
materials with lower loss will be helpful to improve the measurement SNR at higher frequencies.

### 6.4.4. Graphene Surface Conductivity

The Terahertz surface conductivity $\sigma_s$ of a direct-synthesized and transferred graphene thin film on glass substrate is also characterized using Terahertz-TDS. The sample is a thin film consisting of 2~3 graphene layers on a 1 mm thick glass cover slip (the measured substrate refractive index data is shown in Figure 6-7). Using exactly the same waveform truncation of the bare substrate (6 picoseconds before and 24 picoseconds after the main peak), the complex transmittance $T$ of the graphene sample with respect to free space is obtained. Then, the surface conductivities $\sigma_s$ of the graphene sample is extracted by numerical fitting using Equation (6-5) with the measured complex refractive index $n_{\text{sub}}$ of the bare substrate. The obtained surface conductivity of the graphene sample is shown in Figure 6-10. It can be seen that the graphene surface conductivity ranges from 0.5 to 1.55 mSiemens/sq over the 160 - 720 GHz frequency range. Because of its extremely small thickness, the extra power attenuation caused by the graphene sample is indeed quite small when compared to the bare substrate, leading to large uncertainties in the extracted $\sigma_s$. Nevertheless, taking the averaged value over
frequency of 1 mSiemens/sq in Figure 6-10, it translates to a sheet resistivity $R_s$ ($R_s = 1/\sigma_s$) of 1000 $\Omega$/sq, which is quite consistent with measured 700 $\Omega$/sq value at optical frequency [195] as well as 1000 Ohm/sq value at DC [196] for few-layer-graphene (1 to 3 layers) samples.

Figure 6-10. Measured real surface conductivity of graphene thin film on glass substrate with error bars.

**6.4.5. Uncertainty Analysis**

From (6-5), it can be seen that the extracted surface conductivity $\sigma_s$ of a thin film depends on the measured transmission coefficient $T$ of the sample and the refractive
index \( n \) of the substrate. Therefore, to evaluate the uncertainties of this characterization method, the transmission coefficients of the samples and the bare substrates are measured multiple times. In Figures 6-6 and 6-7, the averaged refractive index results are plotted with the statistical error bars (the standard deviation).

As the glass substrate is much thicker than the thin film for our samples, the transmitted Terahertz waveforms can be quite sensitive to the refractive index of the substrate. Uncertainties in the substrate refractive index and the measured sample transmission coefficient are both included in error analysis of the extracted thin film surface conductivity. Therefore, the error bars in Figures 6-8 and 6-10, for the SWNT and graphene surface conductivities respectively, include contribution from both effects. For the SWNT samples, most of the uncertainty is from the substrate refractive index. While for the graphene sample that has larger uncertainties as discussed previously, some of the uncertainty is due to the substrate refractive index uncertainty, but the uncertainty due to the transmission coefficient noise is also obvious here because of the relatively thinner film and thicker substrate.

6.5. An Application Example – Evaluation of Metallic SWNT Content

These measured electric properties of SWNT thin films at Terahertz frequency
provide valuable data for potential applications. As an example, the surface conductivities of SWNT films are characterized using this technique to provide a direct indication of the metallic content in the films. As there is not yet an effective way to control the species of SWNTs during their growth, it is important to find certain means to separate the semiconducting tubes from the metallic ones, or vice versa, in order to realize mass production of CNT circuits. One attractive method is using microwave irradiation induced current to selectively breakdown metallic nanotubes [189]. The scheme of this de-metalization method is shown in Figure 6-11(a): after being exposed to high power microwave irradiation for a period of time, metallic nanotubes in a mixed SWNT film may be broken down and evaporated, while leaving semiconducting nanotubes intact.

SWNT thin films under high power microwave irradiation have been studied. Raman spectroscopy performed before and after microwave irradiation indicated indirectly that metallic content of the SWNT films was decreased after microwave irradiation. However, attempt to evaluate the metallic content of these films via DC conductivity measurement was hindered by contact issues. Contactless Terahertz characterization of those thin films provides an effective method that is much more reliable than DC conductivity characterization to evaluate the metallic content change under various irradiation conditions. For example, Figure 6-11(b) plots the measured SWNT surface conductivity versus microwave irradiation time at 200, 400 and 600 GHz. The observed drastic
decrease of Terahertz conductivity clearly demonstrates significant metallic content reduction due to breakdown of metallic SWNTs in the film after the microwave irradiation [189].

Figure 6-11. (a) Microwave-induced selective breakdown scheme. (b) Surface conductivity (at 200 GHz, 400 GHz and 600 GHz) decreases as a function of irradiation time (from [189]).

6.6. Conclusion
Carbon based nano-material thin films including SWNT and graphene (2-3 layers) on transparent substrates are characterized via the Terahertz time-domain spectroscopy. With uniform field approximation, the films are treated as surface boundary condition between the substrate and air, and their surface conductivities are obtained. To improve accuracy, the precise thickness of sample substrate is calculated through a minimum variation process. The SWNT results show consistent Terahertz surface conductivity for samples with the same film thickness on different substrates and reasonable scaling between films of two different thicknesses. The measured graphene surface conductivity at Terahertz frequency is quite comparable to the reported values in literature for few-layer-graphene at DC and optical frequencies. Taking Terahertz surface conductivity as an indication of metallic nanotube content within a SWNT film, this characterization method has been utilized as a convenient and effective verification method for a potential metallic nanotube removing approach.
CHAPTER 7. THz PHOTOCONDUCTIVE ANTENNA ARRAY BASED NEAR FIELD IMAGING

7.1 Introduction

Terahertz time-domain spectroscopy (THz-TDS) is a very useful tool in various applications such as material characterization and identification, biomedical imaging and nondestructive detection. In a typical far-field imaging setup, the sample is placed in the far-field region of the THz antenna, either in transmission or reflection configuration. However, the resolution of a far-field system is restricted by the diffraction limit of half wavelength. Near field imaging can be applied to improve the resolution which is independent of wavelength but mainly determined by the scanning probe configuration [197].

Most of the previously reported THz-TDS near-field imaging uses the detection mode [197], where the sample is placed very close to the THz detector. Emission mode is also reported in [198] where a single emitting antenna is used for near field scanning. In this work, a 2 × 2 photoconductive antenna (PCA) array is used in a THz near field imaging setup as THz emitters while the sample is placed close to the antenna array as shown in Figure 7-1 (the antenna-sample distance is about 10 μm). A micro-lens array is used to couple and focus femto-second laser pulse onto each antenna. The response of a
sample of gold pattern on quartz is measured. A FDTD model combined with HFSS simulation is used to predict the time domain current and near field scanning result. Good agreement between simulation and experiment is obtained.

With this configuration, a number of useful techniques can be applied including use Hadamard multiplexing method to improve the SNR [199] and apply compressive sensing approach to decrease the number of measurements [200].

Figure 7-1. Microlens array and photoconductive antenna configuration for near field imaging.

7.2 Near field scanning

The PCA array is fabricated on an 800 nm laser-transparent sapphire substrate with 1μm-thick MBE-growth GaAs layer on top. The microscope image of the array is shown in Figure 7-2. The antenna chip is later wire-bonded to a printed circuit board where the
DC biases are connected and individually controlled by switches. Backside laser illumination method is used and the sample is mounted within the sub-wavelength regime of the emitting antennas, about 10 μm in our experiment. The pump laser is split into four beams and each beam is focused onto one element of the PCA array using a micro-lens array with 500 μm pitch size. Another PCA is used as detector which is placed in the far-field region of the transmitter. A sample providing a dielectric-metal edge is fabricated with thin gold film partially covering a quartz substrate is used as the near field imaging example. The sample is mounted on a sample holder which is controlled by a 3D stage for scanning.

Figure 7-2. Microscope image of the 2×2 PCA array (The stripline has 50 μm gap and 20 μm linewidth, the dark circles are SiO₂ passivation layers).
Figure 7-3 shows the configuration of the near field scanning system. The laser is using an 800 nm pulsed laser and the laser power illuminated to the emitter is 10 mW. A chopper is used to modulate the signal and a lock-in amplifier is used to read out the signal.

Initial investigation of this near-field imaging system is carried out by scanning the gold film on quartz sample in the x-direction. Moreover, a time-domain simulation is performed by combing an in-house FDTD model with HFSS simulation. The received time domain signal at a fixed time delay (the first peak position of the signal when antenna is facing the quartz region) versus the scanning positions (with 20 μm step) using just a single antenna are plotted in Figure 7-4 for both the simulation and experimental
data. The values are both normalized to their peak values correspondingly. One can see that the measured result agrees with simulation. The discrepancy between simulation and measurement is mainly due to the inaccurate uniform port assumption in the HFSS simulation setup. Also it can be observed that the spatial resolution defined by 10% to 90% edge response is about 360 μm corresponding to 0.142 λ for the TDS peak frequency of 118 GHz. This result shows the near field scanning system achieved a better resolution than the diffraction limit as expected.

Figure 7-4. Comparison of simulated and measured time domain signal at a fixed time delay versus different scanning positions.
7.3 Hadamard Coded aperture to improve SNR

Figure 7- 5. (a) Photo of the fabricated PCA array. (b) 4 elements Hadamard matrix applied to the antennas to improve SNR.

Hadamard matrix is a square matrix whose entries are either +1 or −1 and whose rows are mutually orthogonal. A Hadamard matrix based multiplexing technique is a common technique in optical frequency to improve the SNR of system [201]. In optical systems, encoded masks are often used to control the transmitting, absorption or reflection of the light.

In this work, the DC bias of each antenna in the antenna array can be independently applied at different voltages. Therefore, by modulating the DC bias of each antenna, a Hadamard code matrix can be applied to the antenna array and improve the SNR of the system. Here we used a 2x2 antenna array as an example. The four antennas in the
antenna array are labeled in Figure 7-5(a) and Figure 7-5(b) shows the 4 elements Hadamard matrix applied to the antennas.

There are two states in this Hadamard matrix: 1 and -1. We used an electronically modulated bias signal applied to the antenna to separate this two cases and used a locked-in amplifier to read the output. The modulating frequency is at 300 Hz. For case 1 and -1, the phase of the modulated antenna bias has a 180 degree shift as shown in Figure 7-6. In the Hadamard matrix, the value in one line indicates the weighting factor of four different antennas in single measurement and the value in one column represents the weighting factor for single antenna at different measurement time. 4 times of measurement in total are needed to obtain the information for 4 pixels. Therefore the total measurement time will be the same as using conventional method which the antennas are turned on one by one each time.

\[
\begin{array}{c}
+1 \\
-1
\end{array}
\]

Figure 7-6. Modulated bias signal applied to the antenna array for two different cases.

If we define the real signal values at 4 antenna locations are \(\psi_1 \sim \psi_4\), the noise for each
measurement are $e_1$-$e_4$. If we use conventional method which turns on one single antenna each time, the output signal $\eta$ at four antenna locations will be:

\[
\eta_1 = \psi_1 + e_1 \\
\eta_2 = \psi_2 + e_2 \\
\eta_3 = \psi_3 + e_3 \\
\eta_4 = \psi_4 + e_4 
\]

For coded aperture case using Hadamard matrix, the output signals after coding at different measurements time will be:

\[
\eta_1 = \psi_1 + \psi_2 + \psi_3 + \psi_4 + e_1 \\
\eta_2 = \psi_1 - \psi_2 + \psi_3 - \psi_4 + e_2 \\
\eta_3 = \psi_1 + \psi_2 - \psi_3 - \psi_4 + e_3 \\
\eta_4 = \psi_1 - \psi_2 - \psi_3 + \psi_4 + e_4 
\]

Therefore, the signals at 4 antennas can be calculated using:

\[
\psi_{1H} = (\eta_1 + \eta_2 + \eta_3 + \eta_4)/4 \\
\psi_{2H} = (\eta_1 - \eta_2 + \eta_3 - \eta_4)/4 \\
\psi_{3H} = (\eta_1 + \eta_2 - \eta_3 - \eta_4)/4 \\
\psi_{4H} = (\eta_1 - \eta_2 - \eta_3 + \eta_4)/4
\]

Assume the measured noise has a Gaussian distribution with standard deviation $\sigma$. Therefore, the standard deviation of signal for the single antenna case is:
\[ E((\hat{\varphi}_i - \varphi_i)^2) = \sigma^2 \quad (7-4) \]

And the standard deviation of signal for Hadamard matrix scanning case is:
\[ E((\hat{\varphi}_{iH} - \varphi_{iH})^2) = \frac{\sigma^2}{4} \quad (7-5) \]

which indicates a fact of 2 SNR improvement compared to the single antenna scanning case.

Figure 7-7 shows the four output time domain signals for single antenna independent measurement case. Figure 7-8 plots the output time domain signals for Hadamard matrix based coded aperture scanning case. Figure 7-9 shows the decoded time domain signals of the 4 antennas from the Hadamard matrix based coded aperture scanning results. It can be clearly seen that the decoded signals from the coded aperture data successfully reproduced the antenna signals compared to the independent measurements. Moreover, Figure 7-10 shows the measured standard deviation with 10 times of measurements for individual measurement case and Hadamard matrix based coded aperture measurement case. One can clearly see that the coded aperture case has a smaller standard deviation than individual measurement case, which means the SNR for the coded aperture scanning case has been improved as expected. The averaged time domain SNR improvement for antenna A, B, C and D are 2.07, 1.96, 2.03 and 2.03 respectively.
Figure 7-7. Output time domain signals for independent measurement using four single antennas.
Figure 7-8. Output time domain signals for Hadamard matrix based coded aperture scanning.

![Image of Figure 7-8](image)

Figure 7-9. Decoded time domain signals of the 4 antennas from the Hadamard matrix based coded aperture scanning results.

![Image of Figure 7-9](image)
Summary

In this chapter, the near field imaging system incorporating a PCA array as THz emitters is designed and realized. A spatial resolution of 360 μm corresponding to 14% of wavelength is achieved with the near field scanning setup. In addition, a FDTD algorithm combined with HFSS time domain simulation is used to modeling the near field performance of the TDS photoconductive antenna. Good agreement between simulation and experiment is obtained. Moreover, a Hadamard matrix based coded aperture scanning method is applied using 2x2 antenna array to improve the SNR. Measured results clearly indicate improved SNR compared to independent antenna measurement.
CHAPTER 8. COMPRESSIVE SENSING BASED MICROWAVE IMAGING SYSTEM

8.1 Introduction

Compressive sensing is a novel sampling / sensing paradigm that enables significant reduction in sampling and computation cost for signals with sparse or compressible representation. It has been experiencing rapid growth in recent years and attracted much attention in electrical engineering, optics, signal processing, statistics and computer science. Using compressive sensing technique, the number of measurements needed can be greatly reduced compared to traditional methods when the signal is sparse in a known basis. The fundamental idea behind compressive sensing is that rather than sampling at high rate first and then compressing the sampled data, it would be much better to directly sample the data in a compressed format [202]. For example, efficient sampling protocols can be designed to capture small amount useful information of the signal in a sparse domain. After sampling, the full length signal from the small amount of sampled data is reconstructed using numerical optimization algorithm.

In ref [203], compressive sensing technique was applied to a microwave imaging system in which a guided wave metamaterial aperture is used to generate different
radiation patterns for compressive sensing. The reconstruction of compressive images at 10 frames per second was achieved at K-band. However, the radiation patterns generated by the metamaterial aperture are basically random and the sampling protocol for this system is not optimized to capture the signal information. In this chapter, a microwave imaging system for human body scanning is investigated. Principal component analysis (PCA) method is used to optimize the measurement radiation patterns for compressive sensing and a reconfigurable array is employed to realize the obtained patterns. Compared to random patterns based compressive sensing system, fewer numbers of measurements is required for this PCA based system to achieve the same performance.

8.2 Principal Component Analysis (PCA) of human images

Principal component analysis (PCA) is one of the most commonly used tools in statistics and data-mining areas for compression and classification of data. The purpose of PCA is to reduce the dimensionality of a data set consisting of a large number of interrelated variables by transforming it to a new set of smaller number of variables, while retaining the sample information as much as possible [204]. These new variables, which are called principal components (PCs), are uncorrelated and are ordered by the fraction of the total information each retains. Therefore, keeping only the values of the
first few principal components would still retain most of the information in all of the original variables. In practice, this PCA is achieved by calculating the covariance matrix of the full data set. The eigenvectors and eigenvalues of the covariance matrix are then computed and sorted according to decreasing eigenvalues [204].

Figure 8-1. Some image examples in the statistical library. The image size is 1.5 m x 2 m.

In this chapter, PCA is applied to achieve a library based compressive sensing system. Before doing compressive sensing, a statistical library which includes a wide range of image examples is used as prior knowledge to obtain the PCA bases. Here we use a
human body scanning system as example, to investigate the compressive sensing performance using PCA generated radiation patterns. 11880 different gray scale images (75x100 pixels) of different people with different height, at different locations, carrying and without carrying a threat weapon are applied as a statistical library. The image resolution / pixel size is 2cm x 2cm. PCA is used to obtain the best projection bases to represent this library. Several example images in the statistical library are shown in Figure 8-1. In practice, an actual implementation would use RF images to train the PCA dictionary. The optical images we used here were surrogates for the more desirable RF data that would eventually be used. Figure 8-2 shows the first few principal components obtained using PCA. It is well-known that the image energy is strongly biased toward low order PCA generated components. This then is a form of sparsity which allows us to obtain very good reconstructions from only a few measurements of the lowest order PCA projections.
These PCA generated bases are applied as the measurement bases in the compressive sensing algorithm. The compressive sensing optimization algorithm applied here is TwIST [205]. Figure 8-3 illustrates the original object images without (top) and with threat (bottom) compared to the compressive sensing reconstructed images using ideal PCA generated bases and randomly generated bases. Both images are reconstructed using 200 bases, each base represents a measurement. It can be seen clearly that the reconstructed image using PCA generated bases has much better performance than that
using randomly generated bases. Basically, with the small number of measurements, it is hard to obtain much information in the random base reconstructed image.

![Figure 8-3](image_url)

(a) Original image (b) reconstructed image using ideal PCA generated bases (c) reconstructed image using randomly generated bases.

Figure 8-3. Compressive sensing reconstructed images using 200 numbers of measurements: (a) Original image (b) reconstructed image using ideal PCA generated bases (c) reconstructed image using randomly generated bases.
Figure 8-4. Root mean square (RMS) error of the compressive sensing reconstructed images using randomly generated bases, wavelet bases and PCA generated bases.

Figure 8-4 plots the root mean square (RMS) error of the compressive sensing reconstructed image with different number of measurements from 50 to 1000 using randomly generated bases, Harr wavelet bases and PCA generated bases. The RMS error of the PCA based compressive sensing system is several times smaller than the RMS error of the random and the wavelet based compressive sensing system for all cases (50 to 1000 measurements).
8.3 Realizing PCA generated radiation pattern using reconfigurable array

8.3.1 Reconfigurable array to control the field distribution

To implement the optimum bases generated by PCA, a reconfigurable array aperture is employed to realize the resulting radiation patterns. By varying the phase and amplitude distribution of the reconfigurable array aperture, the radiation pattern of the aperture and the projected field on the object scene can be controlled. Each projected field distribution thus represents a measurement of the scene.

If we define the object image as $O_i(r)$, the radiated field on the scene as $U(r)$, the measured reflection coefficient $m_i$ of the array will be proportional to $\frac{j}{\beta_0} \sum_{r} [U(r)]^2 O_i(r)$ [203]. By setting appropriate amplitude and phase to achieve $[U(r)]^2$ equal to the PCA generated bases, a discrete set of measurements can be performed and compressive sensing algorithm can be used to estimate information of the scene.

A schematic picture of a reconfigurable array is shown in Figure 8-5. In this example, a 40x40-element reconfigurable array with a unit cell size 4 mm x 4 mm is employed to generate the desired radiation patterns. The operating frequency is at 30 GHz. The scene is selected to be a surface with equal distance to the origin (center of the array) to minimize the distance induced phase difference of the projected field on the scene. The
distance from the origin to the scene is 1.6 m.

8.3.2 Beam synthesis algorithm to control the projected field

If we consider the object scene is in the far-field region of a single element on the array aperture, the field distribution $U(\vec{r}_S)$ on the scene can be approximately calculated using:

$$U(\vec{r}_S) \propto \sum_{\vec{r}_A} A(\vec{r}_A) e^{jP(\vec{r}_A)} \cdot e^{-jkR(\vec{r}_A)} / R(\vec{r}_A)$$

(8-1)

in which $A(\vec{r}_A)$, $P(\vec{r}_A)$ are the amplitude and phase distribution of each element on the reconfigurable array. $R(\vec{r}_A)$ is the distance from the array element to the object which is $\vec{r}_S - \vec{r}_A$.

To synthesize the beam and control the projected field on the object, we applied an
iteration method [206] to optimize the radiated field. First, the required E-field distributions on the object scene generated by PCA are converted into far-field distributions. After that, a far field beam synthesis method [206] is applied to find out the required amplitude and phase distribution of the array elements to achieve this far field distribution for the first iteration. Then, these calculated amplitude and phase distributions of the array elements are inserted into Equation (8-1) to evaluate the achieved field distribution on the object. Of course, the first iteration result may not be able to generate the perfect required E-field distribution on the object scene because the object is not in the far field region of the whole aperture and the number of elements on the array aperture is not infinite. However, using an intersection approach in [206], a new field distribution which is between the perfect pattern and the achieved pattern can be calculated and applied back to the second iteration process. After several iterations, the optimized amplitude and phase distributions of array elements can be obtained.

During the iteration process, a mandatory requirement on the amplitude distribution, such as a uniform amplitude distribution, of the array element can be applied [206]. Therefore, the beam synthesis of a reconfigurable array with phase only control can also be realized since the implementation of a phase-only array is much easier compared to an array that needs both amplitude and phase controls. In the following section, the beam synthesis results using both amplitude and phase controls and phase only control are
compared.

8.3.3 Reconfigurable array to realize PCA generated bases

Figure 8-6. Beam synthesis results to realize the first three bases generated from PCA using both amplitude and phase controls.

From the PCA generated principle components using the previously mentioned statistical image library, there are both positive and negative values in the generated bases.
(i.e., 180 degree phase difference in the E-field distribution). Since it is not easy to implement both positive and negative values using a single pattern, a dual-rail approach [207] is employed in which all the PCA generated bases are separated into positive and negative parts. Each part is treated as an independence base to be realized using the beam synthesis method.

The results using the beam synthesis method to realize the first three bases in Figure 8-2 with both amplitude and phase controls are shown in Figure 8-6, while Figure 8-7 plots the required amplitude and phase distribution of the array elements to achieve these patterns.

Figure 8-7. Amplitude and phase distribution of the array elements to achieve the patterns
in Figure 8-6.

Figure 8-8. Beam synthesis results to realize the first three bases generated from PCA using phase only control.
Figure 8-9. Array elements phase distribution to achieve the pattern in Figure 8-8.

Figure 8-8 illustrates the beam synthesis results with phase only control and Figure 8-9 is the required phase distribution of the array elements. It can be seen that the achieved pattern is worse than the results using both amplitude and phase controls. However, with a uniform amplitude distribution, the reconfigurable array will be simpler and lower cost. For example, a reflect array architecture can be employed [208].
8.3.4 Compressive sensing results using reconfigurable array generated PCA patterns

After the achieved radiation patterns using the reconfigurable array are obtained, these non-ideal bases are applied in the compressive sensing algorithm to evaluate how much the pattern inaccuracies would influence the reconstructed image. To keep generality, the testing objects used here are not selected from the statistical library. Also, noises are added in the measured data, assuming a 10 dB SNR. Figure 8-10 shows the compressive sensing reconstructed image using 200 reconfigurable array synthesized patterns with both amplitude and phase controls (representing only 100 PCA bases because of the dual-rail approach), and the obtained image using 200 random bases. Figure 8-11 plots the RMS error of the compressive sensing reconstructed images with different number of measurements using reconfigurable array generated bases, random bases and the RMS error using full data imaging method. In order to make fair comparison, the time-per-sample for the full data imaging method was reduced to keep the same total measurement time for all techniques. Therefore, the full data measurement operates at a lower corresponding SNR than compressive sensing method. Compared to images obtained using ideal PCA bases as shown in Figure 8-4, the system using reconfigurable array generated PCA bases needs a greater number of measurements to achieve the same RMS error level. However, the reconfigurable array system still yields
much better performance than that of random bases.

Figure 8-10. (a) Original image (b) reconstructed image using 200 reconfigurable array generated patterns with both amplitude and phase controls (c) reconstructed image using 200 random bases.

Figure 8-11. RMS error of the reconstructed image using full data imaging method and the
compressive sensing method with random bases and reconfigurable array generated PCA bases using both amplitude and phase controls.

Figures 8-12 and 8-13 plot the compressive sensing reconstructed image and its RMS errors using reconfigurable array with phase only control. It can be observed that the system performance degrades compared to the reconfigurable array with both amplitude and phase controls. Nevertheless, it still has much better performance than the compressive sensing system using random patterns.

Figure 8-12. (a) Original image (b) reconstructed image using 200 reconfigurable array generated patterns with phase only control (c) reconstructed image using 200 random bases.
Figure 8-13. RMS error of the reconstructed image using full data imaging method and the compressive sensing method with random bases and reconfigurable array generated PCA bases using phase only control.

8.4 Conclusion

In this chapter, a PCA based microwave compressive sensing imaging system is designed. The required radiation patterns from PCA are generated by employing reconfigurable array technique. An iterative beam synthesis method is used to obtain the amplitude and phase distribution of the array elements. The compressive sensing results using both amplitude and phase controlled array and phase only controlled array are reported. Compared to compressive sensing system using random bases, this kind of PCA
based system needs much smaller number of measurements to achieve the same imaging performance.
CHAPTER 9. CONCLUSIONS

This dissertation discusses the design, fabrication and applications of 3D printed components and AM components based system. Also, advanced materials characterization procedure for carbon based samples is studied. Moreover, near field imaging system in THz frequency and compressive sensing imaging system in microwave frequency are discussed.

First, the design, fabrication and characterization for several 3D printed components which includes 3D printed broadband Luneburg lens, 3D printed patch antenna, 3D printed multilayer microstrip line structure with vertical transition, THz all-dielectric EMXT waveguide to planar microstrip transition structure and 3D printed dielectric reflectarrays in microwave and THz frequency are reported. Simulation results are compared with measurements. Good agreement has been achieved for all these components.

Second, using the special property of a Luneburg lens that every point on the surface of the Lens is the focal point of a plane wave incident from the opposite side, the additive manufactured 3D Luneburg Lens is employed for DOA estimation application. 36 detectors are mounted around the surface of the lens to estimate the direction of arrival (DOA) of a microwave signal. The direction finding results using a correlation algorithm
and compressive sensing algorithm are reported. The results show that the averaged error is smaller than 1° for all 360 degree incident angles.

Third, a Luneburg lens based novel broadband electronic scanning system is studied. The radiation elements of the scanning array are located on the surface of a Luneburg lens. By controlling the phase and amplitude of different elements, electronic beam scanning with various radiation patterns can be easily achieved. Compared to conventional phased array systems, this Luneburg lens based phased array structure has a broadband working frequency and has no scan angle coverage limit. Because of the symmetry of Luneburg lens, no beam shape variation would occur for angle scanning. Moreover, this structure requires much less system complexity to achieve a highly directional beam. This reduction in system complexity allows the electronic scanning system to be built at much lower cost than traditional phased arrays.

Fourth, characterization for carbon based (graphene and carbon nanotube) thin films on different substrates via TDS are reported in this dissertation. The film under test is treated as a surface boundary condition between the substrate and air. Using the uniform electrical field approximation, the electromagnetic properties of the film can be precisely extracted. To improve accuracy, precise thickness of sample substrate is calculated through an iteration process for both dielectric constant extraction and surface conductivity extraction. Uncertainty analysis of the measured thin film properties is also
performed.

Fifth, a transmitter based coded aperture TDS near field imaging system by employing photoconductive antenna (PCA) array is studied. Silicon lens array is used to couple and focus the femto-second laser into each PCA. By controlling the DC bias of each PCA element, the ON/OFF state or power level for different PCA elements can be independently controlled. In the experiment, the sample object is placed 10 μm away from the PCA array to measure the THz near field image. A Hadamard matrix is applied to code the 2x2 antenna array to improve the SNR. Measured results clearly indicate an improved SNR compared to independent single antenna measurement.

Finally, a design of a Principal Component Analysis (PCA) algorithm based microwave compressive sensing system using reconfigurable array is proposed. An iterative beam synthesis process is employed to realize the required radiation patterns obtained from PCA algorithm. A human body scanning system is studied as an example to investigate the compressive sensing performance using PCA generated radiation patterns. Optical images are used as surrogates for the RF images in the implementation for training the PCA dictionary. Compared to random patterns based compressive sensing system, this PCA based compressive sensing system requires fewer numbers of measurements to achieve the same performance.
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