3D Printed Frequency Scanning Slotted Waveguide Array with Wide Band Power

Divider

Thesis

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Abstract

The design, fabrication and measurement results of a Ku-band 3D printed frequency scanning SWA array with an accommodating wide band power divider are presented. The antenna system consists of 3 components: an 18×4 non-resonant SWA, a wide band power divider, and 4 waveguide matched load. A compact wide band power divider is proposed as the feed network. To improve the broadband impedance matching, inductive walls are added to the power divider.

The antenna system is fabricated using PolyJet 3D printing technique. Metallization process including nickel electroless plating and copper electroplating are developed. To aid the metallization process, non-radiating slots are added to the waveguide. Measurement result shows *H*-plane beam scanning range from $+15^\circ$ at 12 GHz to -17° at 18 GHz. The lowest and highest realized gains are 4.6 dBi and 21.4 dBi at 12.2 GHz and 15.8GHz, respectively. The measurement result shows good agreement with the simulation result.

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- Antennas
- RF/Microwave Circuit

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Chapter 1. Introduction

1.1 Background

Antenna arrays with beam scanning ability have been widely studied since they provide significant advantages over mechanically scanned antenna. Compared to mechanically scanned antenna, electronically scanned antenna arrays (ESAs) have faster response time, higher accuracy and lower weight. Due to these advantages, ESAs are particularly preferred in military applications such as ground-based or aircraft radar, where high-speed multi-target tracking is required.

ESAs, also known as phased arrays, can be achieved in three different forms: a. passive ESAs, b. active ESAs and c. frequency scanned arrays. A passive ESA is composed of one transmitting/receiving module and several antenna units working as radiating elements with individual phase shifters in front of each unit to control the phase. The transmitting/receiving module is connected to antenna elements through a feed network (usually a power divider). With this configuration, there is no control over amplitude of each element, thus the beam angle can be tuned based on the phase difference between different antenna elements. The disadvantage of this configuration is that the amplitude of 500each element is the same. Thus, the side lobe level and gain cannot be precisely controlled. In addition, the phase shifter is found to be very lossy as

the frequency increases, especially above V-band (40-75 GHz). Thus, the application of the passive ESA is further limited since future high data-rate communication systems require a larger bandwidth, which is achieved by moving the frequency to the millimeter-wave band or even the terahertz band. A block diagram is shown in Fig.1 to illustrate the configuration of a conceptual passive electrical scanned antenna array [1].

Figure 1. Diagram of a passive ESA [1]

An active ESA consists of multiple antenna elements and each antenna unit is connected with an individual transmitting/receiving module. Fig. 2 illustrates the configuration of a conceptual active ESA.

Figure 2. Diagram of an active ESA [1]

With this configuration, each antenna element is not only phase controlled, but also amplitude controlled. In addition, the use of individual transmitting/receiving modules compensate the loss induced by the phase shifter, which is the limiting factor of passive ESAs. Furthermore, as the loss is not the limiting factor, a low loss feed network is no longer a significant concern, which reduces the design complexity of the feed network. Those advantages of agile beam control and low loss come with the cost of elevated fabrication cost and complicated, bulky structure. The cost and structure size can be reduced by the so-called tile structure, where the feed network, the transmitting\receiving module and the antenna aperture are built in individual layers and connected in parallel with each other. Compared with the traditional brick structure where the transmitter modules are rectangular like a brick, the tile structure has significantly reduced the size and cost of the active array. Fig. 3 illustrates a typical tile structure active phased array [2].

Figure 3. An example of tiled active ESA [2]

Fig.4 illustrates a configuration of a conceptual hybrid ESA. A hybrid ESA is the combination of active and passive arrays. In the hybrid array, a main transmitting\receiving module is used to feed the entire array just like the passive array. In addition, a LNA and a power limiter are added to each antenna element together with the phase shifter to improve the noise figure and counteract the loss induced by the phase shifter. This hybrid type phased array compensates the lossy disadvantage of the passive array, with a moderate cost in fabrication. However, the amplitude of each antenna element is still uniform, thus the side lobe problem remains unfixed.

Figure 4. Diagram of hybrid ESA [1]

Frequency scanning array is a special case of the phased array antenna where the main beam steering occurs by the frequency shift of the exciter. The beam steering is a function of the transmitted frequency. This type of antenna is called a frequency-scanning array. The normal arrangement is to feed the different radiating elements from one waveguide. The frequency-scanning array is a special case of serial feeding type of a phased array antenna and is based on a particular property of wave propagation in waveguides. In order to achieve frequency scan, the feed network is typically a travelling wave structure, where the transmitted signal is either absorbed in the load or radiated through the antenna element. Fig. 5 illustrates the conceptual configuration of a frequency-scanned array.

Figure 5. Illustration of frequency scanning array [3]

During the past several decades, passive and active ESAs have been massively studied and widely applied in military applications. However, frequency scanned arrays have not received as much attention due to the limited frequency range and the design complexity of the feed network. A typical frequency scanned antenna array is the non-resonant SWA. Non-resonant SWAs feature structural simplicity, high gain, and low profile. However, the feed power divider accommodating waveguide array structure remains under developed. In this work, we present a non-resonant SWA with a novel power divider as the feed network.

1.2 Slotted Waveguide Array Antenna

SWA is a high gain, low loss, low profile and mechanically robust array. Depending on the termination property, the slotted waveguide antenna can be categorized into two types: the resonant SWA and the non-resonant SWA. The resonant SWA is terminated with a short metal cap in the non-feed end of the waveguide. The non-

resonant slotted waveguide antenna is terminated with a matched load thus preventing all reflection of the transmitted signal. For the resonant slotted waveguide antenna, due to the resonant nature of the shorted waveguide, the bandwidth is limited to a single frequency. Thus, the antenna only has one fixed beam. The advantages of the frequency singularity is that the broad side gain can be massively increased by increasing the slot number and put multiple waveguides in parallel to form a high gain, highly directive and high efficiency antenna array. In addition, since the radiating element (slots) is part of the feed network (waveguide), the slotted waveguide antenna inherently avoids the loss induced by the feed network, making itself a high efficiency and low loss structure. For the reasons above, the resonant slotted waveguide has been widely used in military and space applications. Fig. 6 shows the landing radar of the Apollo aircraft using slotted waveguide antenna array [4].

Figure 6. Landing radar in spacecraft [4]

Traditionally, the resonant slotted waveguide antenna array is fed by a so-called π -junction. Fig. 7 illustrates the structure of an 8 by 10 resonant slotted waveguide fed by a 1 by 8 π -junction. A π -junction is a feed waveguide that is placed orthogonal to the radiating waveguide. The coupling window is opened to transfer the energy to the radiating waveguides. Every two adjacent radiating waveguides will share one coupling window of the feed waveguide. The center-to-center distance of the adjacent coupling windows is set to be one guided wavelength λ_g of the central frequency so that all radiating waveguides are fed in phase. Due to the fixed distance of the adjacent coupling window, this feed method is inherently a narrow band structure. If the frequency changes, the electrical length between each coupling window will no longer be one λ_g , thus the radiating waveguides are no longer fed in phase and the main beam of the antenna pattern will split, creating large grating lobes.

Figure 7. Planar SWA with end feed [5]

Another typical feed structure for resonant SWAs is called bottom feed. Fig. 8 illustrates a 3D printed resonant SWA with a bottom feed structure[5]. Similar to the π -junction, the bottom feed structure utilizes a feed waveguide to couple energy to the radiating waveguide. However, the feed waveguide is placed under the radiating waveguide. A tilted slot is opened to couple the energy to the radiating between the feed waveguide and each radiating waveguide,. The tilting angle of each coupling slot is employed to control the amount of energy coupled to each radiating waveguide. This feed network structure gives the designer the flexibility to lower the side lobe levels in the E -plane. A tapered feed amplitude distribution (i.e. Taylor distribution, Chebyshev distribution) is usually applied to the feed waveguide to minimize the side lobe level [6]. However, since this feed structure will feed the transmitted signal through the middle of the radiating waveguide, its band width is limited to a single frequency point where the slot spacing is exactly $\frac{1}{2}\lambda_g$. In this case, all slots are excited in phase, providing a broad side beam pattern. If the frequency is different from this central frequency, the main beam will tilt due to the phase shift between adjacent radiating slots. However, since the wave is fed through the middle of the waveguide, the beam in the different side will tilt to the opposite direction, which will cause main lobe split and large grating lobes. Thus, this type of feed network is also a narrow band structure.

Figure 8. 3D printed SWA with bottom feed structure [7]

A single column micro-fabricated non-resonant SWA is shown in Fig. 9 [8]. A non-resonant SWA consists of three parts: the feed port, the slotted waveguide antenna and the matched load. The beam steering is achieved by changing the frequency of the feed port. As the frequency changes, the electrical length between adjacent slots is changed. The changed electrical length will provide phase shift between radiating slots. Each radiating slot can be seen as an individual radiating antenna. From array theory, the different phase shift will provide different beam angle, which cause the beam steering. The key point to achieve frequency scan in the non-resonant slotted waveguide is that the matched load must absorb all the electromagnetic waves that reach the end of the waveguide. The reflected wave will generate another beam at the opposite direction of the main beam, which will then create a large grating lobe in the final radiation pattern.

Figure 9. A micro-fabricated non-resonant SWA and radiation pattern from 130 to 180 GHz [8]

One disadvantage of the non-resonant slotted waveguide is that as the frequency differs from the central frequency, the radiating slot length is no longer half wavelength. However, the slot can radiate most of the energy only when the slot length is equal to half wavelength. The major problem induced by the wavelength mismatch is that the antenna gain is small when the frequency is moved away from the center frequency. One way to solve this reduced gain effect is to put multiple identical non-resonant slotted waveguide in parallel and feed them simultaneously. The part missing in this approach is the feed network. Due to its frequency scanning nature of the non-resonant SWA, the feed network has to be a wide band structure with low return loss through the entire band. In addition, due to its travelling wave nature of the waveguide, the signal can only be fed through one end of the waveguide. Thus, the aforementioned feed structures (π -junction, bottom feed) are not applicable to the non-resonant slotted waveguide.

Waveguide power divider is a potential candidate for the end feed network of a non-resonant SWA. Research on wide band waveguide power divider has shown the feasibility this approach. Fig. 10 shows a wide band power divider with a return loss greater than 10 dB of 75 GHz to 110 GHz [9].

Figure 10. Wideband power divider and its S -parameter [9]

There is one limiting factor for the current wide band power divider. The adjacent output ports are separated away to minimize the reflection within the T-junction. However, for the waveguide array structure, the adjacent ports are joined together to

avoid grating lobe. Thus, a novel wide-band power divider that accommodates waveguide array structure must be developed.

1.3 Thesis Statement

This thesis reports the design, fabrication and measurement of a Ku-band 18 by 4 non-resonant slotted waveguide antenna array. The main contributions are: design of a novel wide band waveguide power divider, fabrication using both fused deposition modeling and PolyJet 3D printing technique, copper electroplating followed with nickel electroless plating of all the 3D printed part, the characterization of the metallization performance using optical profilometer, array impedance and pattern measurement of the antenna.

Thesis Statement: *a 3D printed, Ku-band frequency scanning, non-resonant slotted waveguide antenna array is successfully designed, fabricated, and characterized with a novel wide band waveguide power divider accommodating for waveguide array structure and non-radiating slots that enable the metallization of the 3D printed parts.*

1.4 Thesis Outline

The remaining chapters will be organized as follows: Chapter 2 presents a thorough literature review about 3D printing antennas, non-radiating slots and wide band power dividers; Chapter 3 presents the design and simulation of the power divider and the non-resonant slotted waveguide array; Chapter 4 presents the 3D fabrication of the slotted waveguide array, the power divider and the fixture of the termination, and the electroless plating of the 3D printed parts; Chapter 5 presents the measurement of the electroless

plated copper thickness, the one-port and two-port measurement of the divider, and the antenna pattern measurement; finally Chapter 6 presents the summary of this work and the contributions.

Chapter 2. Literature Review of 3D Printing on RF Applications

2.1 3D Printing: Overview

3D printing technique, also known as additive manufacturing, was first invented in the 1980s [10, 11]. The first technology that was used for 3D printing was stereolithography (SLA). With this technique, an arbitrary three dimensional object can be built in a layer by layer fashion. Traditionally, the plastic parts are built with a process called inject molding, where a module of the fabricated part was built elsewhere before the mass production of the object. Design and the fabrication of the prototype are both expensive and time consuming. Therefore, only mass production of the plastic object is possible to compensate the aforementioned cost. SLA effectively reduced the cost and time of fabrication of a single prototype. Lithographical-curable material is used in this process [12-14]. A thin layer of this curable material is formed, and then a movable spot of focused UV light is programmed to cure the fluid material. The motion of the light source is computer programmed in the way that each layer of the material is patterned as the cross section of the desired 3 dimensional object. After building the first layer, the successive layers are built with the same fashion until the entire structure is built.

The stereolithography effectively combined the advantages of computer aided designing (CAD) or computer aided manufacturing (CAF) and 2-dimensional lithographic technology. The use of CAD effectively increase the design simplicity, allowing engineers to create complex, multi-layer, and multi-surface structure in a short time period. After the three dimensional CAD model is designed and optimized, the file

is exported to a special computer program particularly designed for the 3D printing. The CAD model is rendered and sliced into multiple layers according to the technical specification of the 3D printer (temperature, trace width, and layer thickness).

One advantage of the 3D printing is that the increased structural complexity does not increase the fabrication time linearly. This method allows the fabrication of three-dimensional objects with complex, multi-layer structure without increasing the cost significantly. One other advantage of the 3D printing technique is that the fabrication process is highly time and cost efficient. This method allows the designer to easily produce visible prototype, based on which the successive modification and optimization can be performed with a solid perception of the functionality of the prototype. Compared to traditional inject molding fabrication process where a model has to be built before mass production of the prototype, the cost of manufacturing a single prototype is reduced, which enables more iterations of the design optimization, which eventually results in an optimized and robust product.

2.2 3D Printing in RF Application

Traditionally, three-dimensional stereolithography was used to build medical and biomedical devices, which usually utilizes the ceramic material of zirconia and alumina. The first reported 3D printing for RF application is from [15, 16] where a high Q three-dimensional millimeter wave filter is built by ceramic stereolithography. In this work, $\text{Ba}_3\text{ZnTa}_2\text{O}_9$ (BZT), a high permittivity and low loss ceramic material is used for 3D stereolithography. A dielectric cavity resonator is built with center frequency at 33 GHz and a fractional bandwidth of 1%. The high permittivity ceramic is patterned in a

periodical cube to generate the resonant frequency as well as to prevent the leakage of electromagnetic energy. The entire fabrication process is a combination of 3D stereolithography and traditional metal fabrication. The ceramic structure inside the cavity is fabricated by stereolithography while the outer cavity is made from metal, as shown in fig. 11.

Figure 11. Dielectric cavity resonator with 3D stereolithography [16]

The increase of the application of 3D printing in RF devices took place in the early 2010s. A 3D printed 60 GHz dielectric lens is reported using stereolithography is integrated with an rectangular waveguide for gain enhancement and impedance matching [17]. The linear corrugation is covered on top of the lens to reduce the contrast of the permittivity between the air and the ceramic lens. The feeding waveguide is a commercial metal waveguide. A dielectric terahertz EM-crystal waveguide fabricated with polymer jetting is reported in [18]. Polymer jetting is a similar fast-prototyping process with stereolithography. A support structure is built in this process with water-soluble material. The support structure helps to build complicated hollow structure. After completing the print, the fabricated part is rinsed by high-pressure water to remove the support structure.

In this project, a hollow-core electromagnetic crystal (EMXT) type waveguide is used to build the waveguide. Since the entire waveguide is built by dielectric, no metal plating is required in the process.

A conformal electrically small antenna fabricated with omnidirectional printing of metallic nanoparticle inks is reported in [19]. In this process, the conductive metal ink is directive printed on top of a three dimensional surfaces. The silver meander line is designed to miniaturize the antenna. The meander line is patterned onto a convex or concave glass conformal semi-sphere. However, this structure is not by definition a 3D printed structure as the metallic ink can only be deposited onto a planner surface instead of keep stacking layer by layer. The picture of the fabrication in process is shown in Fig. 12.

Figure 12. Printed conformal electrically small antenna [19]

A 3D printed electrically small antenna using selective laser sintering (SLS) is reported in [20]. The antenna utilize the property of a folded helix, which is essentially an electrical dipole. During the printing process, the support structure is printed to help the wire structure stand and stay fixed. The SLS utilizes the plastic material such as ABS or

PLA. Thus, a copper coating is performed onto the printed wire to make it conductive. The initial trial is done by copper paint. However, the measured radiation efficiency is 20% less than simulation result, which is caused by the low conductivity of the copper paint. Thus, polycrystalline copper is used as an alternative method for metallization. Measurement result shows that the plating with polycrystalline copper can provide equivalent performance with completely metal structure.

3D printed horn has been studied due to its high gain and wide bandwidth [21-28]., Horn antennas operating in Ku-band fabricated using electron beam melting (EBM) is reported in [21]. EBM is a 3D printing technique special for metal printing. EBM and selective laser melting (SLM) are the major techniques for metallic 3D printing. In this work , it is shown that 40 um of surface roughness (rms value) caused by the 3D printing process will reduced the gain by around 1 dB compared with traditional metal antenna. Serval horn antennas operating in millimeter and submillimeter wave band fabricated with SLM 3D printing technique are reported in [23-25]. During the SLM process, the metal power (Cu-S15N in this case) was melted by a focused laser source in a layer-by-layer manner to form a 3D structure. Compared to traditional micro-machine process, the SLM process is low-cost and time-efficient. It was demonstrated that SLM printed horn antenna could provide equivalent performance for millimeter wave horn antenna compared to traditional technique.

Serval horn antennas with supplemental feeding network 3D printed using stereolithography or fused deposition modelling (FDM) process are reported in [22, 26-29]. Non-conductive UV-photosensitive resin and ABS\PLA plastic are used as the

material. Thus, the conductive layer must be deposited onto the printed surface. In [29] and [22], the plating is performed by copper spray-coating. The disadvantage of spray coating is that the antenna must be printed in two parts to coat the inside of the antenna. Then the two parts is assembled with nails. The gap between the two parts might degrade the RF performance of the antenna. In [26-28], electroless copper plating is performed. With electroless plating, the entire structure is pretreated and then immersed into a copper salt solution. This way, the entire antenna system can be printed in a single part without the need of disassembly. A picture of 3D printed horn antenna is shown in Fig. 13

Figure 13. 3D printed horn antenna [23]

Other than horn antenna, different RF devices have been demonstrated using 3D printing technique including slotted waveguide antennas [7, 30-33], waveguide filters [15, 16, 34-37], waveguides [38-42], Luneburg Lens [43], dielectric reflectarrays [44], frequency selective surfaces (FSS) [45], and dual reflector antennas [46]. Among those devices, slotted waveguide antennas has attract most attentions since it features high gain, low loss, and low profile.

2.1.2 3D Printed Slotted Waveguide Antenna

Several works have been reported on 3D printed SWA. Yet all the reported work are focused on resonant SWA. In [7], several resonant SWAs operating at X-band or K-band are 3D printed using SLA technique. The metallization process utilized electroless nickel plating followed by copper electroplating. In order to form a uniform copper plating layer inside the structure, non-radiating slots are added to the waveguide where the transverse surface current are minimum for TE_{10} mode. The use of non-radiating slot is critical for electroplating because it provides space for the solution to flow inside the waveguide. A conformal resonant SWA operating at 15 GHz fabricated using direct metal laser sintering (DMLS) 3D printing is reported in [31]. The material used here is alumina alloy. Due to the conductive material used in the process, no metal plating is necessary. The radiation patterns shows good consistency with simulation. However, due to the fabrication defects, the poor return loss is observed. A resonant SWA antenna operating at 78.7 GHz is designed and 3D printed using the SLA technique followed by silver coating [32]. Non-radiating slots are designed to enable metal coating. The metallization process utilizes silver spray coating. All the critical parameters are optimized using machine learning. The gain and S -parameter shows good agreement to the simulation result.

It was observed from the literature review that there is no reported work on 3D printed non-resonant SWA. A recent work from [8] shows a micro-fabricated 130-180 GHz non-resonant SWA. The waveguide is fabricated with PolyStrata sequential copper deposition process, which is similar to 3D additive manufacturing process. The array is designed to have 20 slots. The measured gain ranges from 12dBi to 19 dBi with a beam

scan range of -38° to -2° . The antenna is designed with standard WR-05 waveguide width, providing identical cutoff frequency of TE_{10} mode. The height of the waveguide is slightly different due to the constraint of PolyStrata process. A custom designed PolyStrata *E*-plane bend is designed to connect the waveguide to standard WR-05 flange.

Table 1. Current work on 3D printed SWA antenna

TYPE	Resonant [7]	Resonant [31]	Resonant [32]	Non-Resonant [8]
FREQ	12.5 GHz	15 GHz	78.7 GHz	130-180 GHz
FEED	Bottom Feed	Bottom Feed	Bottom Feed	End Feed
GAIN	25dB	20 dB	22.5 dB	6-13 dB

2.3 Waveguide Power Divider

As concluded in chapter 2.2, currently there is no reported work on two-dimensional frequency SWA. The major challenge is identified as the lack of wide band feed structure. The non-resonant SWA is a travelling structure, thus requiring an end feed structure. Due to the frequency scanning nature, a large bandwidth is required for the feed network. Traditionally, end feed structure with a “ π -junction” is employed in the resonant SWA [5, 47, 48] as shown in Fig. 7. A π -junction is composed of a feeding waveguide perpendicular to the radiating antenna. In the conjunction of each radiating and the feeding waveguide, a coupling window is open to transfer the EM energy to the

radiating array. The distance of adjacent coupling window is exact half-guided wavelength at the center frequency, which assure all the radiating waveguides are feed in phase. However, since the physical distance of the feed port to each radiating waveguide is not the same, the phase will not be the same if the frequency of the transmitting signal differs from the center frequency. Therefore, the π -junction is inherently a narrow band structure. There have been research on adding inductive post to the π -junction to improve the bandwidth [47, 49] as shown in Fig 14. However, the fractional bandwidth is still limited to 10%.

Figure 14. Pi-junction with inductive post [49]

Recently, research on wide band power dividers have achieved wider bandwidth [50, 51] as shown in Fig. 15. A triangular wedge is used in the T-junction to improve the impedance matching, the depth and the width is turned to minimize the return loss [50]. A -10 dB return loss from 93 GHz to 107 GHz is achieved by this topology., Tetrahedral protrusion placed in the T-junction has been shown to improve the impedance matching for power dividers [9]. The tetrahedral protrusion also helps to counteract the reflection from the wedge. A -10 dB return loss from 75 GHz to 110 GHz is achieved by this topology.

However, the configuration of current wideband power divider can not apply to SWA structure because the adjacent output ports in the power divider are separated to avoid increased capacity in the waveguide T-junction. The capacity in the T-junction will degrade the impedance matching of the power divider. As the output ports are joined together, the return loss of the power divider is significantly increased.

Figure 15. Examples of wide band power divider development [9, 50]

Other topology of feed network has also been explored. In [52], a 2-D scanning non-resonant slotted waveguide antenna is designed with an alternating phase wide band power divider. As shown in Fig. 16, the power divider is designed with different physical length for each output path, which feeds the waveguide in different phase. The phase difference between different waveguide provide beam scanning at one direction. The non-resonant SWA provide beam scan in another direction. In this topology, the power divider avoids the problem of the adjacent output ports.

Figure 16. 2D scanning slotted waveguide array

In [53], an oblique feed power divider is employed in a 2D resonant SWA. As shown in Fig. 17, the feed waveguide is tilted, which prevents the reflected wave from adding in phase. However, due to the same problem with the π -junction, the band width of this topology is limited.

Figure 17. Oblique feed SWA

2.4 Conclusion

In this chapter, the author reviewed the development of the 3D printing technique, with a focus in its application in RF engineering. Different printing techniques such as stereolithography (SLA) and fused deposition modelling (FDM) are discussed. It is found that 3D printing has become a popular tool for antenna fabrication due to its advantage of low-cost and fast-prototyping. However, it is found that 3D printed non-resonant SWA is still underdeveloped. The challenge is identified to be the lack of wide band end-feed structure for waveguide array. Then, a literature review of feed topology of the SWA is conducted. It is found that current wide band power divider configuration cannot be directly applied for the waveguide array structure. Thus, a novel wide band power divide must be developed to feed the non-resonant slotted waveguide, which will be further discussed in Chapter 3.

Chapter 3. Design of the SWA Antenna System

3.1 Waveguide Array Design

The array design follows the process outlined in [54-57]. The slots are treated as admittance periodically loaded in a transmission line as shown in Fig. 18. The end of the transmission line is loaded with a matched load G_0 .

Figure 18. Equivalent circuit of a non-resonant SWA

From circuit analysis, it can be drawn that

$$V_n = \frac{K_1 f_n V_n^S}{Y_n^A / G_0} \quad (1)$$

$$\frac{Y_n^A}{G_0} = \frac{K_2 f_n^2}{Z_n^A} \quad (2)$$

Where V_n^S is the voltage of the n^{th} slot and V_n mode voltage referenced at the center of n^{th} slot. $\frac{Y_n^A}{G_0}$ is the slot admittance of the n^{th} slot. K_1 is a constant whose value is unimportant

to the analysis, K_2 is a constant defined by $K_2 = 152.37 \frac{a/b}{\beta/k}$ and f_n is

$$f_n = \left\{ \cos \left[\frac{\beta}{k} v(x_n) y_n \right] - \cos v(x_n) y_n \right\} \sin \frac{\pi x_n}{a} \quad (3)$$

In which a and b is the length of the broad and narrow wall of the waveguide, respectively. $\beta = 2\pi/\lambda_g$ is the guided wave number and $k = 2\pi/\lambda$ is the free space

wave number, x_n is the slot offset from the centerline of the waveguide. $v(x) = kl_{res}(x)$ where $l_{res}(x)$ is the slots length of the slot.

After studying the equivalent circuit in Fig. 18, we can derive the radiated power of each slot as

$$P_n = \frac{1}{2} Re \frac{V_n V_n^* Y_n^{A*}}{G_0} \quad (4)$$

G_0 is the equivalent admittance of the load. Multiple iterations are required to get a precise power distribution value. With the use of array theory, power distribution can be calculated once the antenna pattern is set. Then the slot parameters (length and spacing) can be calculated from the power distribution.

3.1.1 Preliminary Array Design

We started the design with a 1-D 20 slots SWA. The slot spacing is chosen as half guided wavelength at 15 GHz. Here we choose uniform distribution for simplicity and uniformity for all frequencies. Although Taylor distribution or Chebyshev distribution will provide lower side lobe level (SLL) at the designed frequency, the uniform distribution will make the SLL and *HPBW* consistent at all frequencies. The slot offset value is calculated using the equation from [6].

$$d_u = \frac{a}{\pi} \sqrt{\arcsin\left[\frac{1}{N \times G}\right]} \quad (5)$$

Where N is the number of slots and

$$G = 2.09 \times \frac{a}{b} \times \frac{\lambda_g}{\lambda_0} \times \left[\cos\left(0.464\pi \times \frac{\lambda_g}{\lambda_0}\right) - \cos(0.464\pi) \right]^2 \quad (6)$$

λ_g is the wavelength in waveguide, which is

$$\lambda_g = \frac{\lambda_0}{\sqrt{1 - \left(\frac{\lambda_0}{\lambda_{cutoff}}\right)^2}} = \frac{c}{f} \times \frac{1}{\sqrt{1 - \left(\frac{c}{2a \times f}\right)^2}} \quad (7)$$

The length of the slots is optimized to maximize the realized gain at center frequency. The CAD model and the parameters are shown in Fig. 19

Figure 19. Preliminary SWA design (unit: mm)

The entire structure is simulated using both Ansys HFSS and CST Microwave Studio full wave simulator. It is shown from the simulation results that the array has a continuous *H*-plane scanning range from -12° to $+20^\circ$ at the 12 – 18 GHz frequency band. The minimum and maximum realized gain is 4.6 dBi and 17.5 dBi at 12 and 16.2 GHz, respectively. The broadside radiation pattern occurs at 15 GHz, which is resulted from the half guided wavelength slot spacing at 15 GHz. The side lobe level (SLL) is 13 dB, which is expected from the uniform slot offset distribution. Fig. 20 shows the *H*-plane realized gain pattern at 12, 13, 15, 18 GHz.

Figure 20. H -plane realized gain

The Simulated and measured S_{11} is shown in Fig. 21. The measured S_{11} is lower than -12 dB for the entire Ku-band.

Figure 21. Simulated and measured S_{11}

The problem with this design is that the scanning range is not symmetrical off the broadside. In addition, the gain is not consistent (range from 4.6 to 17.5 dBi) across the frequency band. The peak gain is 12 dB larger than the smallest gain. Thus, the design needs to be improved for better performance.

3.1.2 Improved SWA Design

The aim of the modification is to design a non-resonant SWA with symmetrical beam scan range and maximum gain flatness. The beam scan range is determined by the spacing of adjacent slots. As shown in Fig. 22, the spacing of the slots are reduced to 13.8 mm, which corresponds to half guided wavelength at 14 GHz. The slot length is 8.5 mm after optimization. The goal of the optimization is to equalize the realized gain at 12 GHz and 18 GHz.

Figure 22. SWA parameters of the improved design

Figure 23. Simulated 3D radiation pattern of 1D array with improved scanning range

The simulated 3D beam pattern is shown in Fig. 23. As shown in Fig. 23, the scanning range of the 1-D array is from -16.8° to $+14.4^\circ$, which is nearly symmetrical off the broadside. The broad side gain is 8 dB at 14 GHz, the gain and the end of frequency band is 5.1 dB and 5.8 dB at 12 GHz and 18 GHz, respectively. After optimization of the single 1D array, the 2D array is constructed by putting four of the identical 1D array in parallel as shown in Fig. 24.

Figure 24. CST model of 2-D SWA antenna

3.1.3 Non-radiating Slots to Facilitate Metallization Process

Non-radiating slots are added to the SWA for the purpose of better metallization performance. The array is fabricated with PolyJet 3D printing technique. The material for the fabrication process is chosen to be Verowhite photopolymer. Since the material is non-conductive, a RF conductive layer must be deposited onto the plastic surface. Electroless plating process can provide a uniformly plated layer for both outside and inside the structure. However, the thickness of the conductive layer deposited with electroless process is limited to approximately $0.5 \mu m$. Electroplating process can be used to build desired layer thickness. However, the deposition rate is related to the electric field density of the structure. For a hollow structure e.g. waveguide, it is difficult to plate the inner surface. Therefore, the non-radiating slots are designed to increase the flow rate of the electroplating solution.

Non-radiating slots are placed at locations where the transverse surface current is zero to minimize the impact to the antenna performance. The surface current distribution in the rectangular waveguide for TE_{10} mode is derived from [7]. At broad wall, the surface current distribution is,

$$K = \pm \left(\hat{x} j \cos \frac{\pi x}{a} + \hat{z} \frac{\beta_{10}}{\frac{\pi}{a}} \sin \frac{\pi x}{a} \right) e^{j\omega t - j\beta_{10}z} \quad (8)$$

The surface current on narrow wall is

$$K = \hat{y} j e^{j\omega t - j\beta_{10}z} \quad (9)$$

Where

$$\beta_{10} = \frac{2\pi}{\lambda_g} \quad (10)$$

$\hat{z}$ is the direction of the wave propagation along the waveguide. $\hat{y}$ is perpendicular to the narrow wall and $\hat{x}$ is perpendicular to the broad wall. As shown in the equation, the vertical slots that are perpendicular to the broad wall can be added to the narrow wall without impacting EM performance since the surface current distribution on the narrow wall does not have component on the transverse direction. On the broad wall, the transverse current is zero when $x=0$, which is the center line of the broad wall. This indicates that a longitudinal slots located at the centerline of the broad wall can be added without interfering with the antenna performance.

As shown in Fig. 25, the surface electric field of the radiating slots and the non-radiating slots at 15 GHz has an over 40 dB difference, which indicates that the electromagnetic wave does not resonate around the non-radiating slots. Fig. 26 shows the E -field of the bottom wall at 15 GHz where the non-radiating slots are located in the center. As is shown from Fig. 26, there is a little resonance at the bottom electrical field. This little resonance indicates that there is radiation leakage from the bottom wall, which will create a back lobe in the radiation pattern. This is solved by attaching a metal cap or a metal tape on the back of the waveguide after the metallization process. Fig. 27 compared the electric field inside the waveguide with or without the non-radiating slots.

Figure 25. Surface E -field of the radiating slots and the non-radiating slots

Figure 26. Non-radiating slots on the bottom wall

Figure 27. Top: E -field without non-radiating slots; Bottom: E -field with non-radiating slots

As is shown in Fig. 27, the electric field inside the waveguide remains identical before and after adding the non-radiating slots, indicating that the non-radiating slots do not induce mutual coupling between adjacent waveguides or interfere with the antenna performance.

3.1.4 Final 2D Array Simulation Result of the Waveguide Array

The normalized H -plane realized gain at Ku-band is shown in Fig. 28. The beam scan angle across the Ku-band is from -14.4° to $+16.7^\circ$. The simulated 3D realized gain pattern is shown in Fig. 29. The H -plane gain pattern is shown in Fig. 30. The SLL remains below -13 dB from 12 GHz to 17 GHz. The SLL is increased to -10.8 dB at 18 GHz due to the grating lobe effect at higher frequency. The minimal realized gain and the maximum realized gain are 9.5 dBi and 23.7 dBi at 18 GHz and 15 GHz, respectively. The broad side gain is 18.1 dBi at 14 GHz.

Figure 28. Normalized H -plane gain pattern

Figure 29. 3D realized gain pattern at various frequency

Figure 30. *H*-plane realized gain pattern

The return loss is calculated individually at four different waveguide port, which is shown in Fig. 31.

In conclusion, an 18 by 4 SWA antenna is designed and optimized. Simulation results showed the beam scan angle across the Ku-band is from -14.4° to $+16.7^\circ$. Next, the design of the feed power divider and the termination will be discussed.

Figure 31. Simulated S_{11} of the two-dimensional SWA at each waveguide port

3.2 Design of Wide Band Power Divider

The proposed non-resonant SWA antenna is a traveling wave structure. Thus, the feed structure must be implemented at the end of the waveguide. Traditional SWA end-feed structure utilizes a so called π -junction [5, 47, 48], which is composed of a feeding waveguide perpendicular to the radiating antenna. In the conjunction of each radiating and feeding waveguide, a coupling window is open to transfer the EM energy to the radiating array. The distance of adjacent coupling window is exactly half-guided wavelength at the center frequency, which assures all the radiating waveguides are feed in phase. However, since the physical distance of the feed port to each radiating waveguide is not the same, the phase of different radiating waveguide will not be the

same if the frequency of the transmitting signal differs from the center frequency.

Therefore, the π -junction is inherently a narrow band structure. There have been research on adding inductive post to the π -junction to improve the bandwidth [47, 49]. However, the fractional bandwidth is still limited to 10%.

Recently, researches have been done on wide band power dividers [50, 51]. However, the configuration of current wideband power divider can not apply to SWA structure because the adjacent output ports are separated to a certain distance to avoid increased capacity in the waveguide T-junction. The capacitance in the T-junction will degrade the impedance matching of the power divider. As the output ports are joint together, the return loss of the power divider is significantly increased. A preliminary model of a power divider with output port joint together is designed as shown in Fig. 32. Simulation was performed in Ansys HFSS. As shown in Fig. 33, a strong resonance occurs at 15.5 GHz which results in a large return loss up to -0.4 dB. This large return is due to the resonance that occurs in the T junction at 15.5 GHz. In order to eliminate the resonance, the length of the septum is increased to interrupt the resonance in the T-junction as shown in Fig. 34. The return loss of the power divider with increased septum is shown in Fig. 35.

Figure 32. 3D schematic of the power divider with output port joint together

Figure 33. (a) E -field at 15.5 GHz showing resonance, (b) S_{11} of the power divider showing resonance at 15.5 GHz

Figure 34. 3D schematic of the power divider with longer septum

Figure 35. S_{11} of the power divider with longer septum

As is shown in Fig. 35, although the resonance is eliminated compared to the previous case, the return loss in the frequency band is significantly increased. Therefore, a new configuration of wide band power divider must be designed.

3.2.1 Power Divider Design with Arc Wedges

As shown in Fig. 36, the power divider is designed with arc wedges. In this thesis, first divider will refer to the 1 by 2 divider which connects directly to the SWA antenna, second divider will refer to the 1 by 2 divider that connects the input port to the first divider. The output ports are joined together to connect to the SWA antenna. The tapered wedges helps to reduce the reflection from the discontinuity at the corner of the T-junction. As shown in Fig. 37, the return loss of the power divider is less than -10 dB for the entire Ku-band.

The second power divider is designed in the same manner as shown in Fig. 38. This divider is used to split power from the feed port to the aforementioned model. Therefore, the output ports are separated by a distance of $2a$, where a is the length of the broad wall of the WR-62 waveguide. As is shown in Fig. 39, the return loss of the power divider is lower than -10 dB for the entire Ku-band showing good performance. Moreover, the return loss is lower than the first power divider because that the output ports are separated by a certain distance.

Figure 36. 3D schematic of the power divider with arc wedges.

Figure 37. S_{11} in Smith chart and S -parameter of the first power divider

Figure 38. 3D schematic of the modified power divider

Figure 39. S_{11} in Smith chart and S -parameter of the second power divider

With the aforementioned two-way power divider design, we can now cascade them to achieve a preliminary design of the 1 by 4 power as shown in Fig. 40. The return loss of the 1 by 4 power divider is shown in Fig. 41. One should notice that the return loss the return loss is increased significantly (up to -6 dB) after cascading the two divider

in series, despite that the return loss of each individual stage is smaller than -10 dB. The root cause of the increased return loss is that the reflection of each stage adds up in phase at certain frequency, while cancels each other at other frequencies. In order to suppress the adding up behavior of the reflection, the return loss of each individual stage of the power divider must be further suppressed.

Figure 40. 3D schematic of the 1 by 4 power divider

Figure 41. S_{11} of the 1 by 4 power divider

3.2.2 Improved Power Divider Design with Inductive Wall

The reflection of the power divider is mainly induced by the capacitance at the T-junction in the divider. Research has shown that the introduction of inductive walls can counteract the capacitance in the T-junction, which effectively reduces the return loss [58].

The inductive walls are implemented in the beginning of the T-junction as is shown in Fig. 42. The return loss of the first power divider comparing before and after adding the inductive walls is shown Fig. 43. As it is shown in the plot, the $|S_{11}|$ is reduced significantly after adding the inductive walls, especially at low frequencies. Parametric study has been done to determine the optimal length of the induct walls. The width of the inductive walls is 0.4 mm, which is limited by the resolution of the 3D printer. The inductive walls is symmetrical by the centerline of the power divider, which then ensured the equal phase of all output ports.

Figure 42. 3D schematic of the first divider with inductive walls

Figure 43. S_{11} in Smith chart and S-parameter of the first divider after adding the inductive walls

The second divider is modified in the same manner as shown in Fig. 44. The S -parameter of the first power divider comparing before and after adding the inductive walls is shown Fig. 45. As it is shown in the plot, the $|S_{11}|$ is reduced significantly after adding the inductive walls, especially at low frequencies.

Figure 44. 3D schematic of the second with inductive walls

Figure 45. S_{11} in Smith chart and S -parameters of the second divider

With the aforementioned 1 by 2 power divider design, we can now cascade them to achieve a preliminary design of the 1 by 4 power as shown in Fig. 46.

Figure 46. 3D schematic of the 1 by 4 power divider with inductive walls

Figure 47. S-parameters of the 1 by 4 power divider

As shown in Fig. 47, the return loss of the 1 by 4 power divider is significantly reduced. The return loss is lower than -10 dB from 12 GHz to 18 GHz. However, the return loss is still larger than -10 dB from 17 to 18 GHz. In order to suppress the return loss at high end of the band, 2 groups of extra inductive walls are added to the first divider as shown in Fig. 48.

The return loss of the power divider comparing the three cases: without inductive walls, with inductive walls and with extra inductive walls are presented in Fig. 49. With the extra inductive walls, we can achieve -10 dB return loss for the entire Ku-band.

Figure 48. *E*-field inside the 1 by 4 power divider with extra inductive walls

Figure 49. S_{11} of the 1 by 4 power divider

3.3 Modelling of the Terminations

As noted earlier, the non-resonant SWA is a travelling wave structure, thus requiring a matched load in the non-feed end. Research has shown that certain type of 3D printing material has the property of high dielectric constant and high tangent loss [59]. Based on the study of material property, 3D printed waveguide terminations have been developed [60, 61], where an extremely lossy ABS plastic with a relative permittivity of 10 and a loss tangent of 0.03 at X-band is used. Thus, consideration has been given to 3D print the waveguide termination in our lab. The material used here is PLA (TW-PLA175BK, Torwell. Inc.), whose electrical property is measured using Keysight material analyzer. The average measured dielectric constant $\epsilon = 3.0$, the measured

tangent loss $\tan \delta = 0.0145$ at Ku-band. A 10 cm long conical waveguide termination is fabricated in our lab using a FDM 3D desktop printer Ultimaker 2 extended plus. The fabricated prototype and the measured return loss when connected to a single waveguide is shown in Fig. 50.

Figure 50. 3D printed waveguide termination and measured S_{11}

As it is shown in the plot, the average return loss of the termination is -20 dB, whereas commercial waveguide termination normally requires a -40 dB return loss. The root cause of the high return loss is identified as the insufficient permittivity and tangent loss of the 3D printed material used for the prototype. In order to study the impact of the size and the material property on the termination performance, parametric study is performed for different length of the termination and different electrical property of the material of the termination. Fig. 51 shows the CST model of the termination and the simulated return loss. The material used here has a dielectric constant of 10 and a tangent loss of 0.03 (material property extracted from [60, 61]). The length of the termination is 5 cm. Return loss above 40 dB is achieved with this specification.

Figure 51. (a) CST model of the termination, (b) simulated S_{11} with a high loss material

Then, a parametric study of different length with a fixed material property is carried out using the same model. The material used here utilizes the measured result with a dielectric constant of $\epsilon = 3.0$ and a tangent loss of $\tan \delta = 0.0145$. The simulated return loss with the length of the termination varying from 50 mm to 200 mm (notified as lint in the plot) is shown in Fig. 52. It is observed from the result that -30 dB return loss can be achieved with 20 cm long termination with the current material property. It should be noticed that the 100 mm simulation result matches well with the aforementioned measurement result shown in Fig. 50, which indicates that the simulation is accurate. Above all, the parametric study shows that lossy material and longer termination will help reduce the return loss of the termination. In order to make a compact waveguide termination, the material must be highly lossy. Due to the lack of the lossy 3D printed material in the author's lab, four commercial Ku-band waveguide terminations are used for this project. The commercial terminations CTL-0733 are purchased from Resin Systems Corporation with a maximum VSWR of 1.01, which is also characterized in the

simulation (shown in Fig. 53). A fixture of the conical termination is also 3D printed, which will be further discussed in chapter 4.

Figure 52. Simulated S_{11} of the termination with low loss material for different length and 3D schematic model of the termination

Figure 53. (a) Commercial Ku-band waveguide matched load used as a termination, (b) CST model of the termination

3.4 Final Simulation Result of the SWA Antenna System

The final simulation is performed with CST Microwave Studio. The simulation tool is switched from HFSS to CST time domain solver to reduce the computation time.

Fig. 54 shows the CST model of the entire array.

Figure 54. CST model of the entire array

Figure 55. *H*-plane pattern and 3D pattern of the entire array

The *H*-plane radiation pattern and the 3D pattern at 12 GHz, 14 GHz and 18 GHz is shown in Fig. 55. As is shown in the plot, the entire array has a scan range of -16.6° to $+14.3^{\circ}$. The broad side gain is 14 dB. The lowest gain is 8 dB at 12 GHz. The maximum

side lobe level is 11.4 dB at 18 GHz. This increased side lobe level is because that the grating lobe starts to appear at high frequency as the slot spacing approaches to one wavelength.

The S_{11} of the entire array is shown in Fig. 56. As shown in the plot, the return loss of the entire array is less than -10 dB for the entire Ku-band. The beam scan angle across the Ku-band is from -14.4° to $+16.7^\circ$.

Figure 56. S_{11} of the entire array

3.5 Conclusion

An 18 by 4 non-resonant SWA antenna operating at Ku-band is designed and simulated. The entire array has a scan range of -16.6° to $+14.3^\circ$. The broad side gain is 14 dBi. The lowest gain is 8 dB at 12 GHz. A novel wide band power divider is designed to feed the array. Inductive walls are added to the power divider to improve the impedance matching. Simulation performed with HFSS and CST shows good return loss performance.

Chapter 4. Fabrication and Metallization of the SWA Antenna

The fabrication of the proposed antenna was initially performed by fused deposition modelling (FDM) 3D printing technique. ABS plastic is used for all the printed parts due to its high heat deflection temperature and mechanical robustness. However, due to the high surface roughness of FDM printing, the metallization on the fabricated antenna failed. Therefore, PolyJet printing technique is used for lower surface roughness. After the printing process, the copper plating is performed to provide a conductive layer on the waveguide surface. The plating process is composed of the electroless Ni plating to provide a seed layer of conductive Ni. The Ni electroless plating requires etching with HF and surface sensitizing of the plastic. After sensitizing of the plastic surface, the Ni is deposited on it without the need of DC power supply. Due to the nature of electroless plating, both outside surface and inner surface is covered uniformly with nickel. The deposition rate is about 0.1 mil per 20 minutes under the reaction temperature of 45°C. After the Ni electroless plating, the copper electroplating is performed to deposit a thin layer of copper, which provides the required thickness and RF conductivity at Ku-band. The copper electroplating requires a preelectrolyzed copper acid ion solution. The plated part and a copper chunk is placed in the solution and connected to a DC power supply. The plated part is connected to the cathode and the copper metal chunk is connected to the anode. The DC power is controlled to provide the desired surface current density for the electroplating process. After copper electroplating, the waveguide is covered with a uniform copper layer. All three parts are fabricated and

metallized with the same process. Fig. 57 shows the diagram of the entire fabrication process

Figure 57. Diagram of entire fabrication process

4.1 FDM 3D Printing of the SWA Antenna

The entire SWA antenna consists of three parts: the power divider, the SWA, and the termination housing. Initially, all parts were fabricated using the desktop printer Ultimaker 2 plus. Fig. 58 shows the zoomed view of the printed in operation. The attempt to print the first two parts in our lab failed due to wall thickness and the difficulty on removing supporting structure. Fig. 59 shows an example of a failed waveguide print. The wall thickness of the waveguide is 0.8 mm. As a result, during the printing process,

the printer head could not be precisely located on the top of the previous layer of the waveguide wall. The printing of the waveguide power divider was also challenging. The power divider is a tapered structure. Thus, it is inevitable to build support structure inside the waveguide to support the bended divider arm during the plating process. It was found that it is very changing to remove the supporting structure inside the power divider. Thus, another option considered was to build the power divider in a tilted direction as proposed in [31].

Figure 58. FDM printer (Ultimaker 2 extended) in operation

Figure 59. Failed printed parts

Figure 60. Tilted printing direction to avoid supporting structure inside the waveguide [31]

As shown in Fig. 60, the supporting structure was printed in the bottom of the waveguide without interrupting the inner waveguide. Therefore, a trial printing of the power divider is performed by placing the divider in 45° in respect to the build plate of the 3D printer. The result is shown in Fig. 10. As shown in the plot, although the tilted placement of the power divider avoids the supporting structure to be printed inside the waveguide, there are still redundant supporting structures attached to the opening of the waveguide which were difficult to remove.

Due to the difficulty in 3D printing mentioned above, the capability of the FDM 3D printer was not enough for the waveguide structure we need. Therefore, a commercial source was sought as the solution of the 3D printing issue. The 3D printing was done by the Mechanical and Aerospace Department of the Ohio State University. The 3D printer model is the Stratasys Dimension 1200es which utilizes FDM 3D printing technology. The most important feature of the printer is that it uses Soluble Support Technology which dissolves the supporting structure in a water based solution. Therefore, the aforementioned challenge of the supporting structure is prevented. The material for the printing is Stratasys *ABSplus*-P430 with a heat deflection temperature of 96 °C. The final printed part is shown in Fig. 62.

Figure 61. Example of a failed waveguide power divider built in 45° to the build plate

Figure 62. FDM fabricated parts

The dimensions of the fabricated SWA is measured with a digital caliper. The measured slot length is 8.36 mm (8.50 mm in simulation), slot width is 1.65 mm (1.60 mm in simulation), and slot length is 13.63 mm (13.80 mm in simulation). The relative error in longitude direction (alone the wave propagation direction) is less than 1.6%, the relative error in azimuth direction is less than 3%.

4.2 Metallization on 3D Printed Antenna

The 3D printing material is a non-conductive resin. Thus, a conductive layer must be deposited on to the surface of the waveguide to make it electrically conductive. The metallization process of plastic or plating on plastic (POP) has been widely studied since 1960s due to the advantages of low-cost and light-weight compared pure metal structure [62]. ABS is the most plated plastic. The traditional method for ABS plating is usually processed by electroless plating followed by electroplating. The electroless plating process can be preformed by either nickel or copper. In this work, electroless copper plating and electroless nickel plating are both conducted. It was found that nickel plating has better adhesion to the plastic substrate. Electroplating with copper was preformed after the electroless plating to build up desired thickness of copper.

4.2.1 Copper Electroless Plating

The copper electroless plating process is composed by three steps: surface sensitizing, surface activation and electroless copper deposition. The first step, surface sensitizing, issued for the etching of the plastic. The sulfuric acid was used to create micro pores on the plastic. After etching, the surface of the plastic was roughened to provide good adhesion to the future plating layer. The material was converted from hydrophobic to hydrophilic. The second step, surface activating, was used to deposit a thin layer of a precious metal such as palladium. The palladium acts as a bound between the substrate plastic layer and the copper plating. After the surface activating, the treated part is dipped into a copper solution, which is held at 45°C for 30 minutes to deposit a thin layer of uniformly distributed copper. The flow diagram for the process is shown in Fig. 63.

The plating process was done in the author's lab. Due to the toxic vapor of the solution that is generated during the plating process, the process was done in a fume hood lab. A hot plate with a temperature probe is used to control the reaction temperature. The chemical solution for the copper plating is purchased from Transene Company, Inc. The

result of the copper plating part is shown in Fig. 65. The picture of the plating reaction on the hot plate is shown in Fig. 64.

Figure 63. Process of electroless copper plating

Figure 64. Electroless plating in action

Figure 65. Parts after electroless copper plating

Figure 66. Test sample for copper thickness

As it is shown in the Fig. 65, the part is uniformly covered with copper. In order to determine the thickness of the plating, a partially plated sample is 3D printed and plated with the same process as shown in Fig. 66 with an optical image of the boundary of the plated area.

The thickness of the plated copper is measured by a contact profilometer (Fig. 67). The average measured thickness of the copper plating is 0.5 micrometer. It should be noted that at 15 GHz, the skin depth of the copper is 0.53 micrometer. Usually, a metal with a thickness of at least 5 times skin depth can act as a perfect electrical conductor of the electromagnetic wave. However, with the electroless copper plating, only one skin depth of copper can be deposited on the plastic. The insufficient thickness of the copper will increase the conductive loss of the waveguide. This impact will be further discussed in chapter 5.

Due to the thickness limit of electroless copper plating, other approach of metal plating has been investigated. Nickel plating usually has better thickness and better adhesion to the plastic. Thus, nickel electroless plating was done as the alternative approach for metallization.

Figure 67. Measured copper thickness of the electroless plated copper using profilometer

4.2.2 Electroless Ni Plating

Similar to electroless copper plating, the Ni plating process consists of four major steps: etching, activating, fixing, and electroless nickel deposition. The etching solution is composed of chromium trioxide and sulfuric acid. The etching solution etches the plastic and creates micro-pores which acts as the bond between the substrate and subsequent metal layer. This step is held at 60 °C for 20 minutes for the desired etching. The second activating solution was composed of hydrochloric acid and palladium chloride. The hydrochloric acid (HCL) acts as the chrome reducer removing the Cr^{6+} ions residual on the plastic substrate, which will prevent the subsequent metal deposition. The palladium chloride will then be deposited in the micro-pores that were created in the etching process. The palladium acts as a seed layer for the Ni deposition. After activating, the part was then moved to the fixing solution. The fixing solution is composed by sodium hypophosphite monohydrate, which helps to reduce the Ni ions into nickel metal on the plastic substrate. Finally, the part was moved to the electroless Ni plating solution which was composed of a mixture of citric acid monohydrate, ammonium hydroxide, hydrochloric acid, nickel sulfate, and sodium hypophosphite. The solution was held at a temperature of 41 °C for 15 minutes. Then, the part is covered with a uniformly distributed Ni plate.

The experiment is done in the fume hood with a hot plate to control the temperature. Fig. 68 shows the process of etching, activating, fixing and Ni deposition. The RTM Ni Electroless chemical solution is purchased from Transene Company Inc. The detailed plating process is documented in table 2.

Figure 68. (a) Etching, (b) activating, (c) fixing, and (d) nickel deposition

In order to test the performance of the Ni plating, a test sample was 3D printed with FDM 3D printer in author's lab and then electroless plated using the above-mentioned process. It should be noticed that the determining factor of the Ni plating was the surface roughness of the substrate. Thus, the sample was partially polished with a sand paper. As shown in Fig. 69, the top side of the rectangular sheet sample is partially sanded. The bottom of the sample was not sanded. However, since the bottom side of the sample was attached to the build plate of the 3D printer, it was very smooth. Thus, the smoothness level of the samples can be classified into three tiers: unsanded top side (very rough), sanded top side (medium rough), and unsanded bottom side (very smooth). As it is shown in Fig. 69, only the sanded top part was well plated. This indicates that the substrate cannot be too rough or too smooth in order to get an optimal nickel plating.

Table 2. Process for electroless Ni plating

Step	Detail	Temp. (°C)	Agitation	Time
Dish soap Soak	Wash with water and dish soap to remove oil	RTM	no	5 minutes
	Dry			
Water rinse		RTM	no	1 min
IPA Rinse		RTM	no	20 secs
Water rinse		RTM	no	1 min
Etching	RTM A for Plastic	60 °C	no	20 minutes
Sensitizing	RTM B	RTM	no	5 minutes
	No rinse, move to RTM C			
Fixing	RTM C	RTM	no	5 minutes
	DL water rinse		no	15 secs
Nickel Plating	150 ml RTM D + 5 ml Additive (2 stage)			
	First stage: bubbles appear 20 secs after dip into solution	43 °C	no	5 minutes
	DL water rinse, thin nickel plate, uniform plate			
	Second Stage: More bubbles appear	43° C	56 rpm	10 minutes
	DL water rinse: Thicker nickel, good adhesion			
Running water rinse	No nickel particles coming off. Very good plate

Figure 69. Test sample of the nickel electroless plating

Figure 70. Nickel electroless plated power divider

The nickel plated power divider is shown in Fig. 70. The flange was polished with the sand paper while other area was not polished. As is shown in the Fig. 70, the flange area is well plated, while other area is not plated well. This indicates that a smoother surface

is required for plating. Thus, PolyJet 3D printing was used as an alternative technique with better surface roughness.

4.2.3 Copper Electroplating¹

The copper electroplating was used to build a thick layer of conductive copper on top of the electroless plated metal. The copper electroplating was necessary because the Ni developed with the electroless process was too thin, which does not provide enough conductivity as well as thickness. Thus, copper electroplating must be performed after electroless plating process.

The electroless plating utilizes a copper electrolyte solution. The 3D printed part was connected to the cathode while a copper chip was connected to the anode. After applying DC power, the copper in the metal was dissolved into the electrolyte solution and the Cu^{2+} ions were reduced to the 3D plated part in the anode. The electroplating solution was bought from Transene Company Inc. Preelectrolyzation is needed for the solution. The process was done in author's lab with a DC power supply and a hot plate to maintain the temperature of the solution. The preelectrolyzation was done by connecting both anode and cathode with copper chip and applying DC power. This process increases the conductivity of the electrolyte solution and the density of the Cu^{2+} ions. Fig. 71 shows the electroplating process on the waveguide array. The measured sheet resistance is $0.05 \Omega/\text{sq}$ using four point probe.

¹ The copper electroplating process in section 4.2.3 is developed by Ph.D. student Jordan A. Ramsey.

Figure 71. Electroplating on the waveguide array

4.3 PolyJet Printing of the SWA Antenna

Due to the unsuccessful plating on FDM part, an alternative 3D printing technique with lower surface roughness must be adopted. PolyJet printing technique usually has better surface roughness compared to FDM. Thus, all the new parts were printed with the PolyJet technique. The material used is Vero White Plus. All the PolyJet printed parts are purchased from Stratasys, Ltd. Nickel electroless plating followed by copper electroplating are used to deposit the conductive layer on the waveguide as mentioned in

section 4.2.2 and 4.2.3. The pictures of the 3 parts before plating, after nickel plating and after copper plating is shown in Fig. 72, 73, and 74, respectively. Both the nickel plating and the electroplating is successful after using the PolyJet printed parts. A microscope view of a slot is shown in fig. 75, the measured slot dimension is $8.45 \times 1.64 \text{ mm}$.

Figure 72. Divider (a) before plating, (b) after Ni plating, and (c) after Cu plating with microscope view of the inside surface

Figure 73. Termination (a) before plating, (b) after Ni plating, and (c) after Cu plating

Figure 74. Waveguide array (a) before plating, (b) after Ni plating, and (c) after Cu plating

Figure 75. Microscope view of a slot (slot dimension $8.45 \times 1.64 \text{ mm}$) (a) left side, (b) right side

4.4 Conclusion

This chapter presents the fabrication and the metallization of the proposed SWA antenna. The fabrication of all parts is performed by PolyJet 3D printing technique. Metallization is done by nickel electroless plating followed with copper electroplating. Attempt has been given for FDM 3D printing using ABS plastic for better power handling and mechanical robustness, but the plating on ABS failed due to the high surface roughness of the FDM printing.

Chapter 5. Measurement of the SWA Antenna

In order to precisely predict the antenna performance, the simulation model is modified to include the impact of the surface roughness and the coaxial to waveguide adapter. Before pattern measurement, several one and two port measurements are conducted to verify the antenna performance. The 1/2-port measurement results show good agreement with simulation after applying surface roughness and adapter. Then, pattern measurement is conducted at the anechoic chamber at the ElectroScience Lab. It is observed that the measured scan range matches well with simulation value. However, the measured gain is larger than simulation at high frequencies. The reason is identified as the fabrication error on the slot dimension. The simulation is then modified based on the fabricated dimension. The modified simulation results show better agreement with measurement value.

5.1 Characterization of the Coax-WG Adapter & Surface Roughness

In 3D printed structure, the surface roughness is the major cause of conductive loss [23]. In order to characterize its effect on antenna performance, a contact profilometer is used to measure the surface roughness. The Causal Huray model is used to predict the antenna performance in the simulation. In this model, the rough metal surface is characterized as equal sized spheres tightly stacked on a smooth conductor to form a pyramidlike structure [32, 63]. The radius of the sphere is calculated to be 2.5 μm using the equation from [64].

A standard Ku-band coaxial to waveguide adapter (62AC86-1) with a maximum VSWR of 1.25 is used to feed the antenna, which is also characterized in the simulation as shown in Fig. 76.

Figure 76. (a) Coax-WG adapter, (b) 3D schematic of the adapter, and (c) simulated S_{11}

5.2 One/two Ports Measurement of the SWA Components

The S -parameter of the component is measured using Agilent N5230A PNA-L network analyzer. First, the two ports measurement is performed to characterize the return loss and the insertion loss of the power divider. The two power dividers are configured back to back as shown in Fig. 77. (a). Fig. 77 (b) and (c) show the measured S -parameters and comparison with the simulation results with and without the adapter and surface roughness. The measured $|S_{21}|$ (in this thesis, the $|S|$ symbol indicates the amplitude of the S -parameter) for two adapters is approximately -5 dB across the Ku-band, which matches well with the simulation result after applying the surface roughness.

Figure 77. (a) Model of the two ports measurement, (b) measurement result and simulation result without adapter & roughness, and (c) with adapter & roughness

The one port measurement is performed on both the divider and the array. Fig. 78 shows the configuration of the one port measurement on the divider and the measured $|S_{11}|$ compared with simulation results. The measured $|S_{11}|$ matches well with the simulation result after adding the adapter and surface roughness, with the exception that the measured $|S_{11}|$ is larger than -10 dB around 13.5 GHz. Fig. 79 shows the measurement configuration of the SWA antenna and the measured $|S_{11}|$, which matches well with the simulation result after adding the adapter and surface roughness.

Figure 78. One-port measurement configuration and measured S_{11} of the power divider

Figure 79. One-port measurement configuration and measured S_{11} of the SWA antenna

5.3 Pattern Measurement of the Assembled SWA Antenna

The pattern measurement is conducted in the anechoic chamber at the ElectroScience Lab. A 2-18 GHz standard horn antenna H-1498 is used for the gain calibration. The specification of the measurement setup is shown in Table. 3. Fig. 80 shows the measurement setup of the H -plane and E -plane pattern. A 3D printed mount is used to fix the antenna for E -plane measurement. Fig. 81 shows the measured H -plane pattern from 12 GHz to 18 GHz compared with the simulation result. The measured scanning range is from -15.0° to $+17.0^\circ$. The measured patterns are shifted by 2° to compensate for the misalignment during the measurement. Fig. 82 shows the measured pattern at broadside. The measured side lobe level and $HPBW$ for E -plane and H -plane is -11.1 dB, 16.0° and -13.2 dB, 4.4° , respectively. The measured peak gain value from 12 to 18 GHz of the boresight is shown in Fig. 83.

Table 3. Specification of pattern measurement

Feed Horn Freq.	2-18 GHz
Angular Step Size	1°
Frequency Step	5 MHz

Figure 80. Pattern measurement at anechoic chamber for (a) H -plane and (b) E -plane

Figure 81. Measured and simulated H -plane pattern from 12 GHz to 18 GHz showing scanning range from -15° to $+17^\circ$, for those plots, measured pattern is shifted for 2° to compensate for the misalignment during the measurement.

Figure 82. (a) Measured E -plane pattern, (b) Measured H -plane pattern, both at 14.6 GHz showing beam at the broadside

Figure 83. Peak gain at boresight from 12 to 18 GHz

5.4 Discussion of Pattern Measurement Result

The one/two ports measurement and pattern measurement of the fabricated SWA antenna are presented in this chapter. The measured beam scan angle is -17° to $+15^\circ$ across the Ku-band. The measured side lobe level and *HPBW* for *E*-plane and *H*-plane at 14.6 GHz is -11.1 dB, 16.0° and -13.2 dB, 4.4° , respectively. The measured *S*-parameters show good agreement with simulation results after applying surface roughness and coax to waveguide adapter. However, the measured gain is larger than simulation value at high frequencies. The reason is identified as the fabrication error on slot length. Measured slot length using digital caliper shows that the average slot length is 8.45mm whereas the simulation model is 8.5mm. Thus, simulation model is modified to reflect the fabrication error.

5.5 Modified Simulation with 8.4 mm Slot

As noted earlier, the slot length is decreased in the simulation model. The simulation result with the shorter slot is shown in Fig. 83. After the reducing the slot length to 8.4 mm, the peak gain is shifted to higher frequency compared to previous simulation, which shows better agreement with measurement result. Fig. 85 shows the simulated peak gain without adding the surface roughness and adapter. The *H*-plane pattern at different frequency and comparison with modified simulation result is shown in Fig. 86.

Figure 84. Simulated gain over frequency with shorter slot

Figure 85. Gain over frequency with shorter slot

Figure 86. Measured and simulated H -plane pattern (with 8.4 mm slot) from 12 GHz to 18 GHz showing scanning range from -15° to $+17^\circ$

Figure 87. Simulated radiation efficiency

5.6 Radiation Efficiency of the SWA Antenna System

As shown in Fig. 87, the simulated radiation efficiency is reduced by approximately 40% at center frequency after applying the surface roughness.

5.6 Conclusion

The one/two ports measurement and pattern measurement of the fabricated SWA antenna are presented in this chapter. The measured beam scan angle is -17° to $+15^\circ$ across the Ku-band. The measured side lobe level and *HPBW* for *E*-plane and *H*-plane at 14.6 GHz is -11.1 dB, 16.0° and -13.2 dB, 4.4° , respectively. The measured gain ranges from 4.6 dBi to 21.4 dBi at the frequency band. The measured gain is higher than simulation value at high frequencies. The discrepancy is caused by the fabrication error on the slot length. This is verified by reducing the slot length to 8.4 mm in the simulation. As noted in section 5.5, the modified simulation result shows better agreement with measurement result. At all frequency, the cross-pol is 20 dB less than co-pol at boresight. A comparison of the measurement and simulation result is shown in Table. 4.

Table 4. Comparison between simulation and measurement results.

	Simulation	Measurement
Scanning range	-14.4° to $+16.8^\circ$	-15° to $+17^\circ$
Peak Gain	18.6 dBi (@ 15.6 GHz)	21.4 dBi (@ 15.8 GHz)
Lowest Gain	5.1 dBi (@ 12 GHz)	4.6 dBi (@ 12.16 GHz)
Maximum SLL (<i>H</i> -plane)	-11.2 dB (@ 18 GHz)	-10.68dB (@ 12 GHz)
<i>E</i> -plane SLL at broadside	-13.4 dB	-11.2 dB

Chapter 6. Conclusion and Future Work

6.1 Summary

A 3D printed Ku-band frequency scanning SWA antenna is designed, fabricated and measured. A wide band power divider is designed to feed the antenna. The inductive walls are added to the power divider to improve the broadband impedance matching. The antenna parameters are optimized to achieve symmetrical beam scan range and optimal gain flatness. Non-radiating slots are added to the array to enable metallization. A Ku-band waveguide matched load is used to terminate the array, which is also characterized in the simulation. 3D printed waveguide termination has been investigated and discussed. The proposed is fabricated using PolyJet 3D printing technique and metallized using nickel electroless plating followed with copper electroplating. Attempts have been given on FDM 3D printing for better power handling and mechanical robustness, but the plating is failed due the surface roughness of FDM printing. The measured scanning range is from -15.0° to $+17^\circ$. The measured side lobe level and *HPBW* for *E*-plane and *H*-plane at 14.6 GHz is -11.1 dB, 16.0° and -13.2 dB, 4.4° , respectively. The maximum measured gain is 21.4 dBi at 15.8 GHz. Overall, the measurement matches well with the simulation result. Table. 5 shows the comparison of this work and reported 3D printed SWAs. At the best of our knowledge, this is the first 3D printed non-resonant SWA antenna.

Table 5. Comparison of this work and state-of-the-art

	[7]	[31]	[32]	[8]	This work
Type	Resonant	Resonant	Resonant	Non-resonant	Non-resonant
Freq. (GHz)	12.5	15	78.7	130-180	12-18
Feed	Bottom feed	Bottom feed	Bottom feed	End feed	End feed
Gain (dBi)	25	20	22.5	6-13	4.6-21.4
Element numbers	10×8	48	10×10	20	18×4
Scan range	N/A	N/A	N/A	-38° to -5.5°	-15° to -15°
Fabrication technique	SLA 3D printing	SLA 3D printing	SLA 3D printing	PolyStrata Microfabrication	PolyJet 3D printing

6.2 Potential Application of the Proposed Antenna

This light-weight, frequency scanning antenna can be used in any non-communication applications. Particularly, the light weight property makes this antenna preferable for airborne application such as weather radar and landing radar.

6.3 Future work

One disadvantage of this antenna is that the gain difference across the frequency band is too large. This can be solved by changing the slot offset to increase the radiated power at the end of the frequency band. Another problem is that granting lobe will appear at E -plane at high frequencies due to the element spacing. The element spacing is limited by the waveguide dimension. This can be solved by replacing traditional rectangular waveguide by ridged waveguide. Another way to avoid the granting lobe is using corrugated waveguide. Corrugation can increase the dispersion of the EM wave, which increases the scanning angle. Thus, same scanning range can be achieved within smaller frequency range. In summary, we envision the following of the future work of this project:

- Increase gain flatness
 - Optimizing slot offset
- Reduce E -plane SLL
 - Use ridged waveguide
 - Use corrugated waveguide

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