

Enhanced EIRP and Reconfigurable Polarization Antenna for Millimeter-Wave Digital Beamforming using Circular Arrays

by

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Author's Declaration

I hereby declare that I am the sole author of this thesis. This is a true copy of the thesis, including any required final revisions, as accepted by my examiners.

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Abstract

The move towards millimeter wave (mm-wave) frequency bands in wireless communication has been driven by the need for higher data rates. However, some of the major challenges faced with the shift to mm-wave are high free space propagation path loss and limited output power of power amplifier (PA). As a result, phased arrays have emerged as a crucial technology in delivering higher effective isotropic radiated power (EIRP) whereby numerous antenna elements, each driven by PAs, are combined over the air. To increase the overall EIRP, the number of PAs can be increased and consequently, the number of antenna elements are also increased. However, this approach results in larger array panel sizes and narrower beamwidth, which may not be suitable in certain 5th generation (5G) scenarios, such as mobile applications. The narrow beamwidth can potentially make the alignment between the transmitter and receiver difficult and result in signal degradation. Alternatively, the output power of the transmitter can be increased by utilizing multiple PA channels in conjunction with on-chip power combining. On-chip power combining techniques such as Wilkinson power combining, direct power combining, or transformer-based power combining have been widely used in the literature. However, as the number of PA channels to be combined increases, on-chip power combining becomes less feasible due to the limited Q -factor of on-chip passive components. On-antenna power combining has been proposed to use the antenna as both a radiating element and power-combining passive. Through the use of multi-feed antennas, multiple PA channels can be used to drive a single radiating element.

This thesis presents a multi-feed active antenna module that offers enhanced EIRP and polarization reconfigurability operating from 37 GHz to 40 GHz. Additionally, the active antenna module is suitable for use in antenna arrays. First, the return loss of the multi-feed active antenna module is analyzed and the overall active antenna module design is optimized. Next, a short parametric study on planar circular array configuration is conducted to find an optimal antenna array design using the active antenna module. Lastly, a seven element fully-digital beamforming array using the active antenna module is proposed, capable of polarization reconfigurability operating from 37 GHz to 40 GHz.

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List of Abbreviations

- 5G** 5th generation iii, 1, 2, 27
- BFIC** beamforming integrated circuit 26, 30, 61, 64
- EIRP** effective isotropic radiated power iii, 3, 4
- FNBW** first null beamwidth 6
- FR1** frequency range 1 1
- FR2** frequency range 2 1
- HPBW** half-power beamwidth 6
- LNA** low noise amplifier 26
- LoS** line-of-sight 2
- mm-wave** millimeter wave iii, 1–4, 61
- NLoS** non-line-of-sight 3
- PA** power amplifier iii, 3, 26
- PCB** printed circuit board 4, 64
- RF** radio frequency 2–4
- SNR** signal-to-noise ratio 3

Chapter 1

Introduction

1.1 Motivation

Advances made in wireless communication in the last few years has not only paved the way for the introduction of the 5G mobile network, but also highlights the importance of the role of 5G in the coming decade. As the number of mobile users, Internet of Things devices, smart vehicles, virtual reality, and any other appliances that rely on some form of wireless communication link increase, the need for 5G becomes more prevalent. Not only does 5G offer significantly higher data rates, but also promises lower latency and larger bandwidths.

Similar to previous broadband cellular networks, the 5G spectrum is also divided into multiple frequency bands. Two different frequency ranges have been allocated for use in 5G spectrum: frequency range 1 (FR1) and frequency range 2 (FR2)). FR1 include frequency bands between 0.41 GHz and 7.125 GHz, whereas FR2 include frequency bands in the mm-wave frequencies. Given that most of the mm-wave frequency spectrum is still largely unused and the constraint of limited bandwidth availability in FR1, there has been enormous research efforts to make mm-wave 5G networks commercially viable. Another advantage of systems operating at mm-wave frequencies is the physical size of the antenna is much smaller than the size of the antennas operating at FR1 frequency bands. The decrease in antenna size allows for better integration with smaller, handheld devices.

However, there are many challenges encountered with the shift of wireless communica-

tion systems to mm-wave frequencies, such as high free-space path loss, limited antenna gain and limited transmit power. In order to overcome these challenges, multiple transmit chains can be used in conjunction to meet the power requirements at mm-wave frequencies, but at the expense of increasing the design complexity of the overall system, and necessitates the use of advanced fabrication processes, which may not always be commercially viable. Subsequently, there is a demand for low-cost mm-wave 5G systems that can offer higher data rates and support larger modulation bandwidths, while meeting the transmit power requirements to overcome the high free-space path loss experienced at mm-wave frequencies.

A cornerstone of mm-wave 5G systems is the use of beamforming phased-array antennas; a collection of antennas with varying phases working in conjunction to create a directional radiation beam. The concept of phased-array antennas has existed for decades and has been used in numerous military applications, such as the PAVE Phased Array Warning System in 1979 [14]. The use of phased-array antennas in 5G mm-wave systems has become commonplace in recent years due to their ability to produce highly directive radiation patterns, which is necessary to overcome the propagation losses at higher frequencies. But designing phased-array antennas at mm-wave frequencies is challenging due to factors such as fabrication tolerances, heat dissipation, and routing between front-end modules and antennas.

1.2 Problem Statement

Most phased-array antennas operating at mm-wave frequencies are implemented using analog beamforming; the phase input to each element is adjusted at mm-wave frequencies using an analog phase shifter and only one digital signal can be transmitted or received throughout the entire antenna array. While radio frequency (RF) beamforming is a sufficient solution at mm-wave frequencies for some applications, RF beamforming has many limitations. Firstly, analog beamforming can only support one transmit/receive beam at a time, which is not viable if there are multiple concurrent users. Secondly, the beamsteering resolution of an analog beamforming antenna array depends heavily on the analog phase shifters driving each element; if the phase shifter does not perform as expected, the radiation beam pattern can be severely impacted. Lastly, analog beamforming can only service line-of-sight (LoS) users.

On the other hand, digital beamforming is achieved by having the same number of digital signals as the number of antenna elements within the array and the phase adjustment is implemented in the digital domain. By having the same number of digital signals as antenna elements, advanced digital signal processing techniques can be employed to enable multiple concurrent transmit or receive beams, enable higher beamsteering resolution, mitigate the effects of multi-path fading effects, and even service non-line-of-sight (NLoS) users.

Still, there are many drawbacks of digital beamforming architecture, one being the power consumption; each antenna element requires its own RF chain which consumes much more power than just the one required RF chain in analog beamforming. Another drawback of digital beamforming architecture is the difficulty in implementation at mm-wave. Because there are far greater number of RF chains, the integration between the RF front-end modules and the antenna elements is a lot more complex, especially given the confined spacing between antenna elements in order to avoid grating lobes. Furthermore, the output power of a PA at mm-wave frequencies is limited, and consequently, the radiated power from each antenna element will also be limited. In order to realize digital beamforming at mm-wave frequencies, a novel antenna design will be required to overcome the limited output power of PAs at higher frequencies and achieve higher EIRP. To increase the EIRP, the number of elements in an antenna array can be increased, but will lead to larger array panel sizes and narrower beamwidth. The narrower beamwidth can be problematic for dynamic scenarios such as mobile application, where misalignment between the transmitter and receiver can be detrimental in terms of signal-to-noise ratio (SNR) [15].

Moreover, in scenarios where multiple wireless systems coexist, such as crowded urban areas, interference from other wireless sources or reflected signals can seriously deteriorate the quality of signals sent or received. To overcome such challenges, polarization diversity is an important feature to have, as it helps reduce interference from other sources with similar polarization orientations. Additionally, having polarization diversity can be a cost-effective solution as it enables the realization of one system to achieve multiple polarization orientations rather than multiple systems with one fixed polarization orientation.

Lastly, implementing antenna arrays at mm-wave frequencies is challenging due to the constrained spacing between antenna elements. Traditionally, linear arrays or uniform rectangular arrays are used in the implementation of antenna arrays at mm-wave frequen-

cies, however, this approach is not sustainable as antenna spacing decreases whereas RF front-ends will roughly stay the same size. Consequently, a different approach is required to relax the spacing constraints without compromising the overall radiation pattern of the antenna array.

This thesis proposes the design of an enhanced EIRP antenna in a uniform circular array configuration to enable digital beamforming at mm-wave frequencies on printed circuit board (PCB) technology. Furthermore, the enhanced EIRP antenna is designed to be polarization re-configurable and can achieve four polarization states while maintaining good matching across the different polarization states.

1.3 Thesis Organization

This thesis is organized as follows: Chapter 2 presents different beamforming architectures, the theory used in different antenna array topology, and a literature review of notable or recent works in antenna arrays, digital beamforming at mm-wave and multi-polarization antennas. Chapter 3 focuses on the designs of the novel enhanced EIRP antenna. Next, Chapter 4 presents the design of the uniform circular array used. Lastly, Chapter 5 concludes this thesis and recommends future research directions based on the work presented.

Chapter 2

Background and Literature Review

2.1 Antenna

An antenna is defined in the IEEE Standard for Definitions of Terms for Antennas as “that part of a transmitting or receiving system that is designed to radiate or to receive electromagnetic waves” [16]. The design of an antenna is a crucial part of any wireless communication system, as it is the component used to transmit or receive electromagnetic waves over free-space. If an antenna is not designed properly, radio waves cannot be radiated out or received efficiently, thereby severely degrading the performance of the entire wireless communication link.

2.1.1 Antenna Figures of Merit

An antenna’s performance can be characterized in several ways and depending on what wireless communication link the antenna will be used, different antenna properties will be desired. In the following section, different metrics describing antennas will be defined.

Radiation Pattern

The radiation pattern of an antenna is used to describe the distribution of radiated energy into free space. The radiation pattern can be a graph of the electric or magnetic field at

a constant radius, or a graph of the power density along a constant radius. Because the radiation pattern of an antenna is a three-dimensional quantity, it is usually sufficient to plot the radiation pattern as a two-dimensional plot as a function of θ and a fixed ϕ or vice versa. An example radiation pattern of an antenna is shown in Fig. 2.1. Different parts of the radiation pattern can be broken down into different radiation lobes. A radiation lobe is defined as a “portion of the radiation pattern bounded by regions of relatively weak radiation intensity” [16]. Radiation lobes can be further classified based on their characteristics, which is especially useful when describing an antenna’s radiation pattern.

A major lobe refers to the “radiation lobe containing the direction of maximum radiation”, and can be seen in Fig. 2.1 at $\theta = 0$. Two metrics can be used to describe the major lobe: half-power beamwidth (HPBW) and first null beamwidth (FNBW). The HPBW is defined as the angular separation between the two points of the major radiation lobe at which the peak radiation intensity decreases by 3 dB and is measured in degrees.

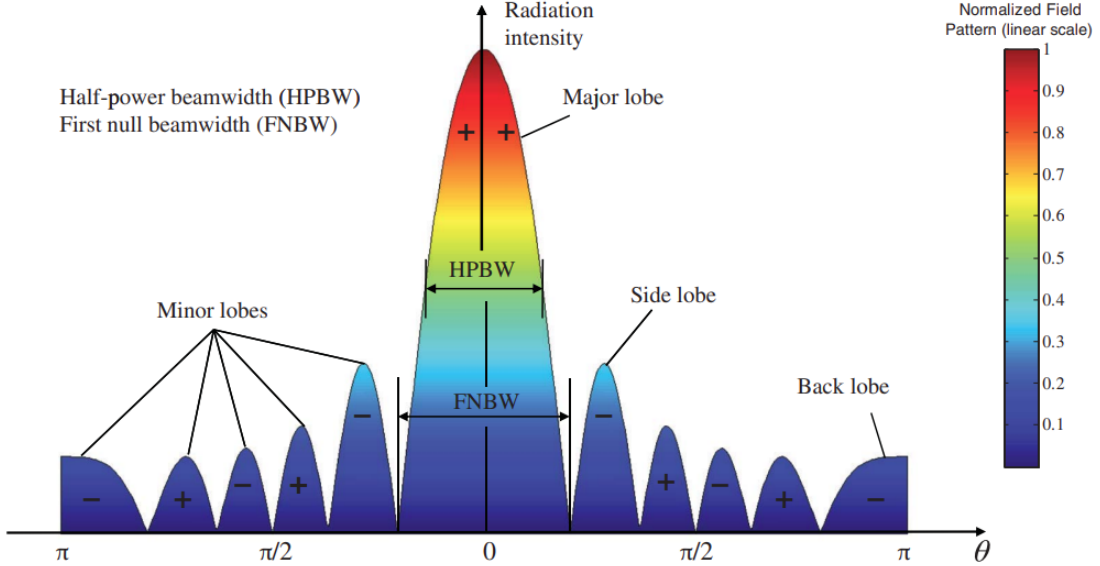


Fig. 2.1: Example radiation pattern [1]

The FNBW is defined as the angular separation between the two points of the antenna’s radiation pattern where the first null appears after the major lobe, and is also measured in degrees. The minor lobes are defined as any radiation lobes that are not defined as the major lobe. The term side lobe is usually used to describe the minor lobe adjacent to the

major lobe. Lastly, the back lobe refers to the radiation lobe that is 180° away from the direction of the major lobe.

Directivity

Usually, the antenna design in question is not an isotropic source, meaning it does not radiate in all directions equally. Thus, directivity is a metric used to quantify the radiation intensity in a specific direction by a non-isotropic source compared to the radiation intensity of an isotropic source. When a specific direction is not given, the directivity is taken to be in the direction of highest radiation intensity. Directivity can be defined by equation 2.1.

$$D_0 = \frac{U_{max}}{U_0} = \frac{4\pi U_{max}}{P_{rad}} \quad (2.1)$$

where

D_0 : maximum directivity (dimensionless)

P_{rad} : total radiated power (W)

U_{max} : maximum radiation intensity (W/unit solid angle)

U_0 : radiation intensity of isotropic source (W/unit solid angle)

It is evident that in the case of an isotropic source, equation 2.1 results in $D_{max} = 1$ when $U_{max} = U_0$. In antenna array applications, it is desirable for the antenna element to not have any back-radiation or to exhibit side-lobes in its radiation pattern. For 5G applications where beamsteering is used, having high directivity is desired, as directivity is a measure of how well the antenna can direct energy in a certain direction.

Antenna Efficiency

An antenna's overall efficiency, denoted by e_0 in equation 2.2, is taken at the terminals of the antenna structure. The total antenna efficiency considers the losses due to reflection mismatch (e_r) and losses due to conductor (e_c) and dielectric (e_d) resistivity.

$$e_0 = e_r e_c e_d \quad (2.2)$$

where

- e_0 : total efficiency (dimensionless)
- e_r : reflection efficiency (dimensionless)
- e_c : conduction efficiency (dimensionless)
- e_d : dielectric efficiency (dimensionless)

An antenna's efficiency is an important performance metric to consider; having poor antenna efficiency can be a serious bottleneck to any wireless communication link and increase overall power consumption. Usually e_c and e_d are combined into one term called antenna radiation efficiency and is defined as $e_{cd} = e_c e_d$. It is difficult to compute e_c and e_d , but can be experimentally found, however, the two quantities usually cannot be measured separately.

Gain

While the directivity of an antenna only describes the antenna's ability to focus energy in a specified direction, the gain of an antenna is a figure of merit that takes into account how directive the antenna is as well as the efficiency of the antenna. The gain of an antenna is given by the following equation:

$$G_0 = e_{cd} D_0 \quad (\text{Dimensionless}) \quad (2.3)$$

Equation 2.3 takes into consideration the losses of the antenna itself, such as the conduction losses or dielectric losses, but does not take into account the mismatch losses when the antenna is connected to a transmission line. To take into account the mismatch losses, a reflection efficiency factor can be added to Equation 2.3 to create a new metric: realized antenna gain, given by the following equation:

$$G_{re0} = e_r G_0 = e_r e_{cd} D_0 = e_0 D_0 \quad (\text{Dimensionless}) \quad (2.4)$$

Ideally, an antenna will be perfectly matched to the connected transmission line and will have zero reflection losses, which will result in $G_0 = G_{re0}$. In such a case, the only losses will be due to conduction losses or dielectric losses.

Bandwidth

The bandwidth of an antenna can be defined as the range of frequencies where the antenna's performance, such as gain or input impedance, is designed to operate optimally. Apart from frequency independent antennas, most antennas have a specified bandwidth of operation; at frequencies outside an antenna's bandwidth, the antenna will exhibit deteriorated characteristics. As a result, it is important to design an antenna to operate at and across the specified frequency band of interest.

The bandwidth of an antenna can be given as a ratio of the upper cut-off frequency to the lower cut-off frequency. For example, a 5:1 bandwidth antenna indicates the upper cut-off frequency is five times greater than the lower cut-off frequency. Alternatively, the bandwidth of an antenna can be presented as a fractional bandwidth; the difference between the two cut-off frequency divided by the centre frequency of the bandwidth. For example, a 10% bandwidth antenna operating at 30 GHz would have frequency of operation between 28.5 GHz and 31.5 GHz.

The bandwidth of an antenna can refer to either the impedance bandwidth or pattern bandwidth. When referring to the pattern bandwidth, characteristics such as gain, sidelobe level, and beamwidth are used as metrics to see how they vary across frequency. In contrast, the impedance bandwidth refers to the input impedance at the antenna's terminals across frequency.

Polarization

The polarization of an antenna refers to the orientation of the electromagnetic waves radiated or received by an antenna. The orientation of the electromagnetic field is defined by the direction of the electric field vector within the wave. Two main types of polarization exists: linear polarization and circular polarization. Moreover, there exists two types of linear polarization and two types of circular polarization. The two linear polarization types consist of vertical polarization and horizontal polarization, whereas for circular polarization, there exists left-hand circular polarization (LHCP) or right-hand circular polarization (RHCP). The polarization of an antenna is an important factor when designing antennas, as it directly affects the performance and efficiency of the overall wireless communication link. Polarization mismatch occurs when the polarization of the receiving and transmitting antennas are not aligned, which results in some transmitted energy not being efficiently

received by the antenna. This will ultimately lead to reduced signal loss and degradation in quality of signal.

Axial Ratio

Axial ratio is a parameter used to describe the quality of circular polarization in an antenna and quantifies how closely the polarization of the circularly polarized wave approaches true circular polarization. The axial ratio of an antenna is defined as the ratio between the major and minor axes of the polarization ellipse. An ideal circular polarization will have equal major and minor axes and exhibit an axial ratio of 1 (0 dB). However, if the circular polarization is poor, the major and minor axes will not be equal, the axial ratio will be greater than 1 (> 0 dB) and elliptical polarization will be observed. For truly linear polarization, the axial ratio will be infinite, since the minor axes will be zero.

Effective Isotropic Radiated Power

The effective isotropic radiated power (EIRP) is a measure of how much output power is radiated in a given direction compared to an isotropic source. EIRP is a metric that takes into account both the transmitter power and the ability of the antenna to direct the power in a specified direction.

$$EIRP = P_t + G_t \quad (\text{dBi}) \tag{2.5}$$

where P_t is the total power supplied to the antenna in dBW and G_t is the gain of the antenna in dBi. In order to increase the EIRP of an antenna, then either P_t , G_t , or both P_t and G_t has to be increased. As such, the EIRP is an important factor to consider in any wireless communication link, and will dictate the design of the antenna used.

Array Factor

So far, the aforementioned antenna figure of merits have been used to describe a single antenna element. However, when multiple individual antennas are arranged in a specific configuration to create an antenna array, the array factor is used to describe the overall

radiation pattern of the antenna array. The array factor is defined in [16] as “the radiation pattern of an array antenna when each array element is considered to radiate isotropically”. The array factor assumes each antenna element within the array are identical and have the same radiation properties. The array factor of an antenna array depends on three factors: the spatial distribution of each element, both the amplitude and phase excitation of each element, and the number of elements in the antenna array. By only varying the amplitude and phase excitation of each element, the array factor can be varied and steered in a desired direction without changing the configuration of the array. However, the expression for array factor depends on the spatial configuration of the antenna array. Examples of common antenna array configurations include uniform linear array (ULA) or uniform rectangular array (URA). Alternatively, antenna arrays can also be arranged in a circular configuration, and the array factor of a circular array will be discussed in depth in Chapter 4.

2.2 Beamforming Architecture

In an effort to achieve massive multiple-input multiple-output (MIMO) systems at mm-wave frequencies, beamforming antenna arrays have seen widespread use in both academia and industry due to beamforming antenna array’s ability to dynamically change its radiation patterns based on user location. Beamforming can help create more stable and reliable connections between base stations and end users, as wireless links at mm-wave frequencies require line-of-sight (LoS) and are a lot more susceptible to interference. Beamforming architectures can be classified into three groups: analog beamforming, digital beamforming and hybrid beamforming. The three different beamforming architectures are shown in detailed in Fig. 2.2.

Before discussing beamforming architecture, it is important to first discuss the different building blocks of an RF chain. A typical RF chain includes: digital-to-analog converter (DAC) or analog-to-digital converter (ADC), phase shifter, power amplifier (PA) or low-noise amplifier (LNA), filters, upconverter or downconverter, and switches. At mm-wave frequencies, the output power of PAs are limited, which is why phased array antennas are employed to overcome the free space propagation path loss by combining over the air.

Analog Beamforming

In analog beamforming, only one common RF source is fed into the antenna array, which is then split accordingly to each antenna element within the array using passive power splitters. Shown in Fig. 2.2c, analog phase shifters are used to apply the required phase shifts at RF frequency. Because only one common RF source is required, only one frequency conversion block and DAC/ADC is required to implement analog beamforming. Due to only requiring one common RF source, analog beamforming has the lowest design complexity and has the lowest power consumption compared to the other two beamforming architectures. However, a major disadvantage of analog beamforming is line-of-sight requirements; there should be a clear, unobstructed path between the analog beamforming array and the end-user.

Digital Beamforming

In digital beamforming, shown in Fig. 2.2a, each antenna has a dedicated RF chain; if the antenna array has n antenna elements, then n number of RF chains are used to excite the array. Because of the large number of RF chains used, digital beamforming offers the highest degree of freedom in terms of digital pre-coding and allows for frequency selective phase shifting of each antenna element. Digital beamforming arrays are able to be used in multi-users, multi-stream scenarios or non-line-of-sight scenarios, of which analog beamforming arrays are unable to be used. Even though multibeam can be achieved with analog beamforming arrays, multi-streams of data cannot be performed with analog beamforming

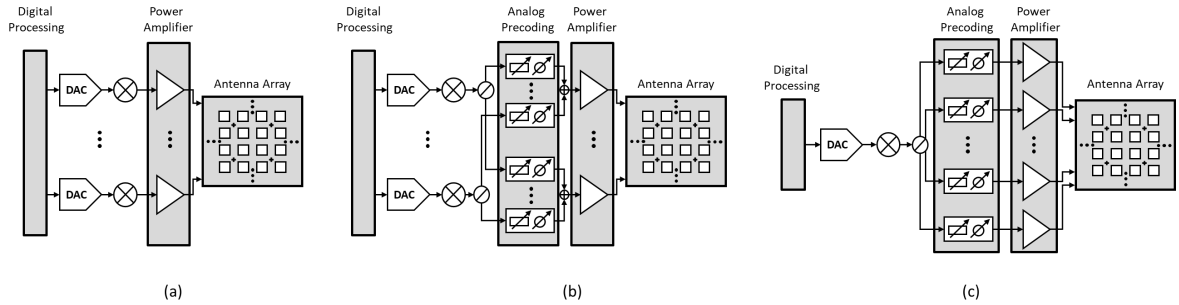


Fig. 2.2: Different beamforming architectures: (a) digital beamforming, (b) hybrid beamforming, and (c) analog beamforming.

arrays. However, these benefits come at the expense of higher power consumption, higher design complexity, and higher hardware costs.

Hybrid Beamforming

Lastly, in hybrid beamforming arrays, as the name suggests, is a solution between digital beamforming architecture and analog beamforming architecture; more than one RF chain is used to excite the array, unlike analog beamforming, but the number of RF chains used is less than the number of antenna elements, unlike digital beamforming. Hybrid beamforming architecture is shown in Fig. 2.2b. With hybrid beamforming architecture, there is a compromise between the number of data streams available and the power consumed. Compared to a fully-digital beamforming architecture, fewer DAC/ADCs are used, which equates to lower hardware cost and power consumed; conversely, compared to an analog beamforming architecture, more than one DAC/ADC is used, so more than one data stream is available.

Currently, most works operating at the mm-wave frequencies have implemented analog beamforming architecture or hybrid beamforming architecture; largely due to the lower design complexity, lower cost of manufacturing and lower power consumption. However, to further investigate the viability of massive multi-user MIMO systems, the need for antenna arrays utilizing fully-digital beamforming architecture is needed.

2.3 Literature Review

In the following section, a literature review of existing works implementing a fully-digital beamforming architecture is presented. Afterwards, multi-polarization antenna designs are reported and compared.

2.3.1 Digital Beamforming at mm-wave Frequencies

The advantages of digital beamforming, compared to other beamforming architectures, includes: higher channel capacity due to the ability to have multiple data streams, realization of multiple beams, and high beamsteering resolution. However, to implement a

fully digital beamsteering array requires many complete transmitter/receiver chains; the number of transceiver chains is equal to the number of antenna elements in the array. For this reason, it is incredibly costly to implement digital beamforming arrays, especially if high-resolution or wideband ADCs/DACs are used. In the following section, several notable works implementing fully digital beamforming architectures are presented.

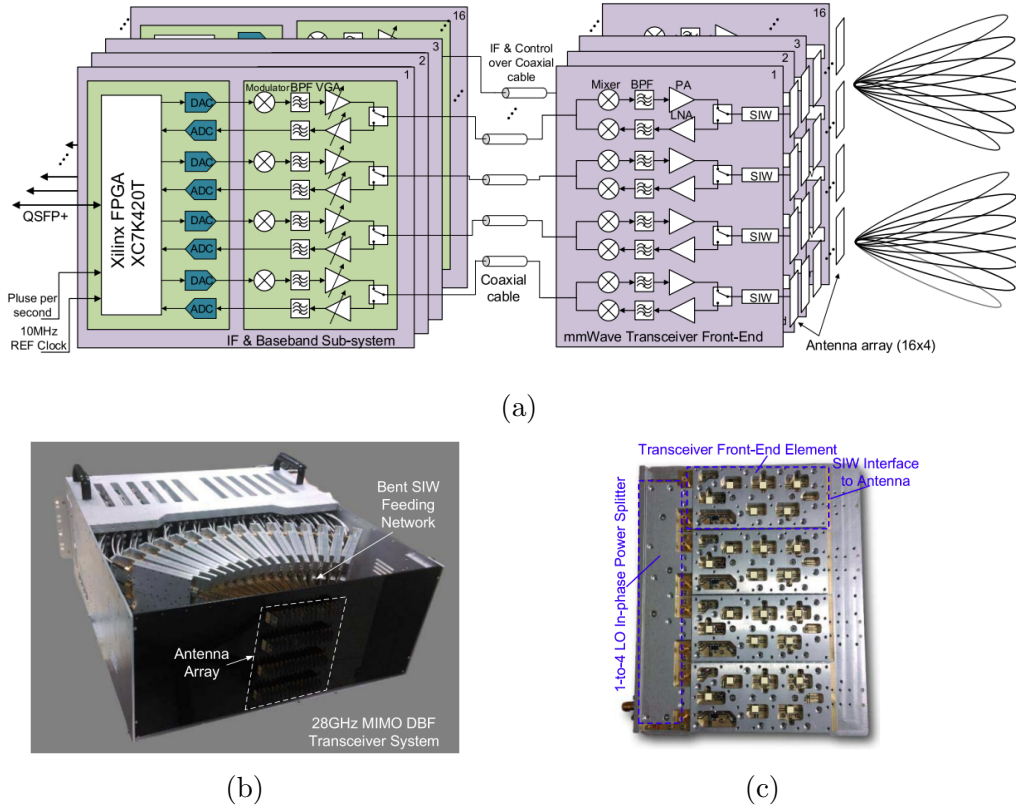


Fig. 2.3: Fully digital beamforming-based MIMO transceiver operating at the 28 GHz band. (a) Block diagram of entire transceiver system. (b) Photograph of fabricated and assembled transceiver system. (c) Fabricated mm-wave transceiver front-end board in metal housing. [2]

In 2018, Yang et al. presented a 64-channel massive MIMO transceiver with a fully digital beamforming architecture at mm-wave frequency [2]. This MIMO transceiver operates in time division duplex mode at a centre frequency of 28 GHz with a 500 MHz signal bandwidth. The design of the transceiver system is split into two parts: the mm-wave

transceiver front-end array (mmWFEA) and the IF-baseband subsystem, which can be seen in the block diagram shown in Fig. 2.3a. The 64-element antenna array configuration is 4 rows of 16 elements, as shown in Fig. 2.3b. The mmWFEA consists of five amplifiers, four filters, two passive sub-harmonic mixers, and one PIN switch. The mmWFEA uses a CPWG-to-SIW transition from the TX/RX chain to the antenna the SIW feed of the antenna element and each mmWFEA board consists of four RX/TX chain, as shown in Fig. 2.3c. The reason behind using a SIW feed is due to the limited antenna spacing of $\lambda/2$ and the thickness of the mmWFEA board, the boards needed to be bent and arranged in a arc-like fashion, as shown in Fig. 2.3b. The RF performance of the transceiver system was measured using both continuous wave (CW) and quadrature amplitude modulated-64 (QAM-64) signals. Using CW measurements, a transmitter (Tx) gain of 29 dB and a Tx gain flatness of 1.05 dB was measured across 500 MHz bandwidth. Using a single-carrier QAM-64 signal, the gain imbalance and quadrature error were measured to be -0.002 dB and -45.676 m-degree, respectively.



Fig. 2.4: Fully digital beamforming 32 elements antenna array. [3]

Next, two over-the-air (OTA) measurements were conducted using the presented transceiver system as a base station (BS) and a four-element antenna array as the user equipment (UE). For both OTA tests, a 500 MHz wide signal using OFDM QAM-64 modulation scheme was used for both OTA tests. The first OTA test consisted of a single moving UE and the beam-

tracking abilities of the transceiver BS. The second OTA test conducted was the peak data rate test with multiple UEs. Up to 20 non-coherent data streams were transmitted to multiple UEs at the same time, and a peak data rate of 50.73 Gbps was achieved.

Similarly, Chopra et al. presented a novel fully digital beamforming based receiver architecture with 32 channels operating at 25.1 GHz [3]. The antenna array is planar and arranged as 4×8 elements, which is shown in Fig. 2.4. The MIMO testbed has an IF of 3 GHz and a maximum baseband capture bandwidth of 1 GHz. OTA measurements were also conducted using this 32-element MIMO prototype; the MIMO testbed was placed on the rooftop of a parking lot 50 m high, while a transmitter system was placed on the roof of a van. The required baseband and up conversion modules were placed inside the van and measurements were collected while the van drove, up to 40 miles per hour (64.37 kilometres per hour), around the receiver MIMO testbed.

Liang et al. presented in 2017 a real-time massive MIMO prototype with 64 independent channels [17]. This MIMO prototype operates at a centre frequency of 4.64 GHz and is able to support 200 MHz data streams to a total of 12 UEs concurrently. Using this MIMO prototype as a BS, both single-user MIMO and multi-user MIMO experiments were conducted in both LoS and NLoS circumstances in a real urban application scenario in Yokohama, Japan. For the LoS single-user MIMO and multi-user MIMO scenario, a throughput of 1.54 Gbps and 11.29 Gbps, respectively, was achieved. However, no detail was reported in the implementation of this 64 channel MIMO prototype.

Another fully-digital beamforming approach was presented by Lu et al. in 2021 [4]. Lu et al. presented a sixteen-element fully integrated digital beamformer fabricated using 40 nm-CMOS technology also operating at the 28 GHz band. The overall IC system architecture can be seen in Fig. 2.5a. The presented work incorporates the entire RX chain within the IC, including: the mm-wave front end, ADCs, clock generation and bit stream processing. The CMOS die is flipped chip onto an eight layer 4×4 patch antenna array realized on LTCC substrate. The antenna element used in this work consists of a patch antenna, which is aperture coupled. The feed lines from the flipped-chip CMOS die to the antenna elements and the antenna structure are shown in Fig. 2.5b. OTA measurements were also conducted using a 100 MHz wide QAM-4 modulated signal and an EVM of -18 dB was reported. A narrow bandwidth modulated signal was used due to the ADC limited bandwidth of 100 MHz. Overall, the CMOS die itself occupies 7.73 mm^2 and consumes a total power of 2.8 W.

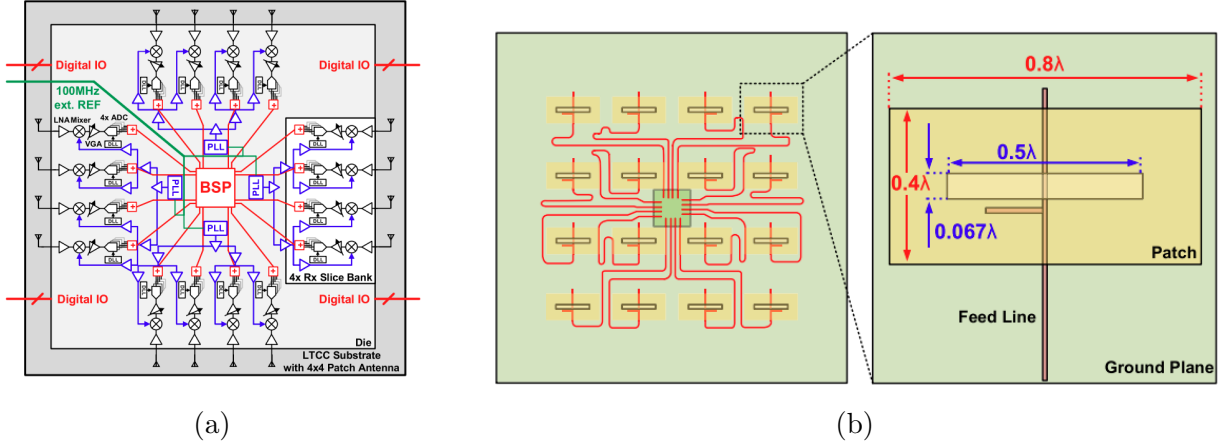


Fig. 2.5: Fully integrated digital RX beamforming IC operating at 28 GHz. (a) Block diagram of receiver architecture. (b) Aperture-coupled microstrip patch antenna array. [4]

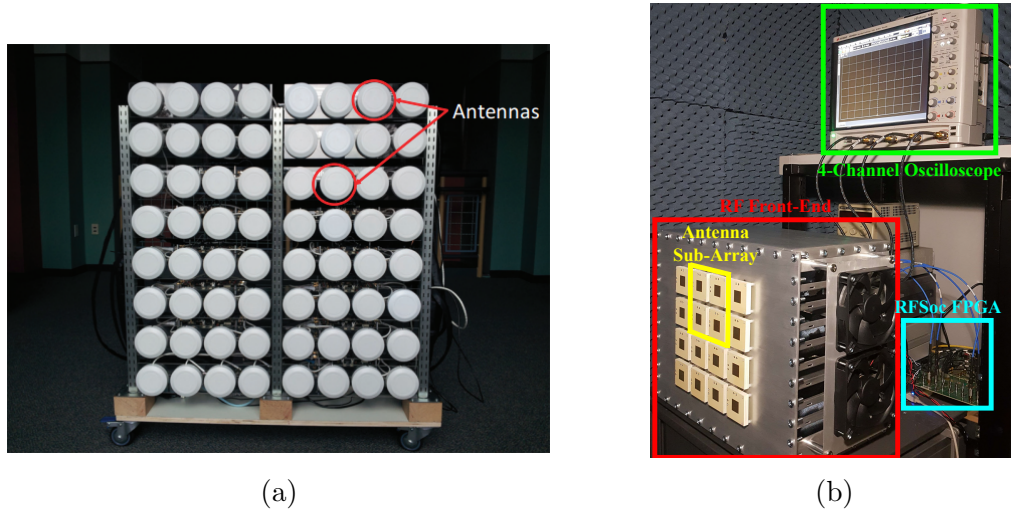


Fig. 2.6: Sub 6 GHz fully digital MIMO systems. (a) “Argos” MIMO testbed from Rice University [5]. (b) 4×4 MIMO FPGA testbed from University of Waterloo [6].

Other fully-digital MIMO testbeds have also been realized, however, at the sub- 6 GHz frequency band. For example, Shepard et al. presented in 2012 a base station architecture that is scalable up to thousands of antennas, shown in Fig. 2.6a [5]. This work was named “Argos” and an initial prototype of 64 antennas was realized. The “Argos” prototype was

capable of serving 15 terminals concurrently at the 2.4 GHz frequency band. Experiments were conducted in several indoor locations to determine the cell capacity of improvement of the Argos testbed. Another fully-digital MIMO testbed shown in Fig. 2.6b is from the University of Waterloo; a Xilinx Ultrascale+ RFSoc ZCU216 evaluation board is used to feed a custom-built 4×4 antenna array at a centre frequency of 3.5 GHz [6]. To validate the MIMO FPGA implementation, a 2×2 subset of the entire antenna array was excited with four uncorrelated 75 MHz orthogonal frequency-division multiplexing (OFDM) signals.

2.3.2 Multi-Polarization Antenna

Antennas that can attain different polarization states, also known as multi-polarization antennas or polarization re-configurable antennas, are highly sought after for scenarios where polarization diversity is required. For example, in crowded urban areas where multiple wireless communication systems coexist with each other and signal reflections can occur, polarization diversity can be invaluable. Polarization diversity can reduce losses due to multi-path fading [18], increase channel capacity in MIMO systems [19], improve signal-to-noise ratio due to interference from other wireless systems or polarization mismatch between transmitter and receiver.

Generally, there are two approaches to designing multi-polarization antennas: by using a re-configurable feeding network to the antenna or by using electrical switches on the antenna itself to change the antenna structure between different polarization states. In the second approach, radio-frequency microelectromechanical systems (RF-MEMS) or PIN diodes are usually employed as electrical switches on the antenna structure and are controlled by a DC bias to switch on or off depending on the desired polarization state.

PIN diodes have two states: “on” state and “off” state. When the PIN diode is in “on” state, it exhibits very low resistance and essentially acts as an electrical short. Conversely, when the PIN diode is in “off” state, the PIN diode exhibits very high resistance and acts as an electrical open. An early work using PIN diodes to realize polarization re-configurable antennas was presented by Kim et al. in 2008 [7]. The work consisted of a circular patch antenna that was able to achieve different polarization states using a single microstrip feed. In total, 5 PIN diodes were used in the antenna design to operate at 2.46 GHz for WLAN applications. In this antenna design, a “U” shaped slot was cut into the circular patch and two of the PIN diodes were used to connect the slot to the rest

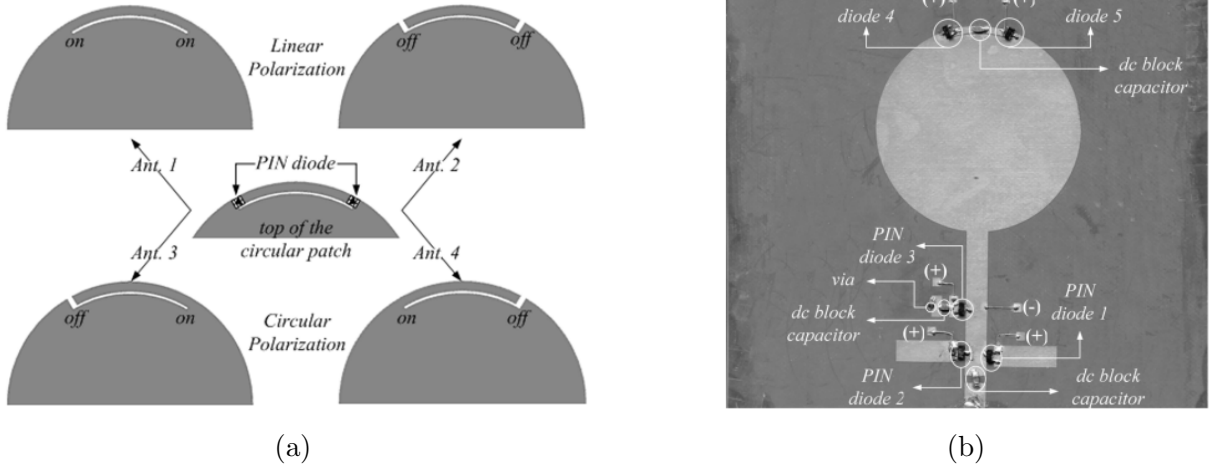


Fig. 2.7: Circular patch antenna with single microstrip feed with re-configurable polarization capability on PCB. (a) Antenna in different polarization states. (b) Photograph of the fabricated antenna. [7]

of the antenna. Using the two PIN diodes to alter the antenna structure, four states can be achieved: high-frequency linear polarized, low-frequency linear polarized, right-hand circular polarized and left-hand circular polarized. Shown in Fig. 2.7b are the different configurations of the antenna structure between different polarization states. However, a limitation of this work is the input impedance changes as the polarization states changes. Which is why three PIN diodes (PIN diode 1-3) are used in a re-configurable stub matching network, as seen in Fig. 2.7b.

Another example of multi-polarization antennas using PIN diodes is presented by Tran et al. in 2017 [8]. In this work presented by Tran et al., a crossed bow tie dipole antenna utilizing 6 PIN diodes and a single probe feed was able to achieve five polarization states: two circular polarization states and three linear polarization states. Shown in Fig. 2.8b is the bow tie antenna design and the location of the six PIN diodes used. Unlike the work presented by Kim et al., a matching network is not needed, and all six PIN diodes are only used to alter the antenna structure between different polarization states, shown in Fig. 2.8a. The input impedance of this antenna design is wideband; able to achieve fractional bandwidth of 37.1% across all polarization states, due to the inherent wideband property of bow tie antennas. Although this work does not require a matching network,

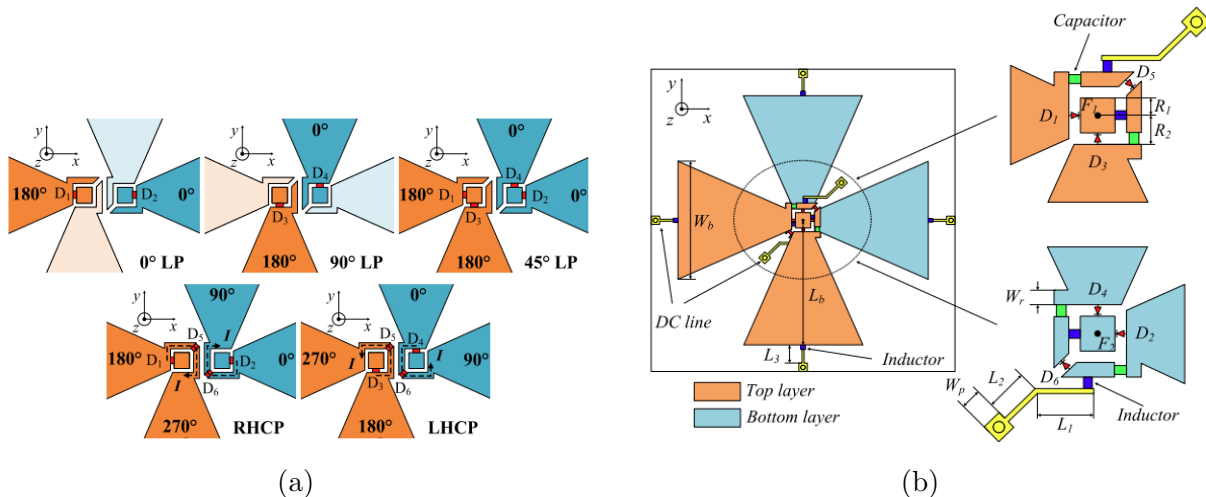


Fig. 2.8: Crossed bowtie dipole antenna realized on a two layer PCB. (a) Antenna configurations for different polarization states. (b) Details of crossed bowtie dipole antenna design. [8]

this work still suffers from changing input impedance, and somewhat unstable radiation pattern across different polarization states. Other works utilizing PIN diodes to achieve multiple polarization states [20, 21, 22] also suffer from varying input impedance or unstable radiation pattern as the polarization states are changed.

All works presented so far using PIN diodes operate at the sub-10 GHz range and might not be suitable for use in mm-wave frequencies, especially if an integrated antenna array package on PCB solution is desired. Additionally, PIN diodes suffer from lower isolation in the open state as the frequency of operation increases and parasitic elements from mounting the PIN diode itself become more pronounced, which will invariably introduce undesired perturbations on the antenna structure itself when tuning between different polarization states. Lastly, PIN diodes require higher DC power supplies and a complex bias circuit to isolate the RF signal on the antenna from the power supplied to the PIN diodes. These problems will be compounded at higher frequencies, as parasitic elements in surface mount components used in the bias circuit will make designing these bias circuits more difficult.

Alternatively, RF-MEMS based switches can be used instead of PIN diodes to realize polarization re-configurable antennas. RF-MEMS switches, compared to PIN diodes, have much higher isolation when operated at higher frequencies and do not consume as much

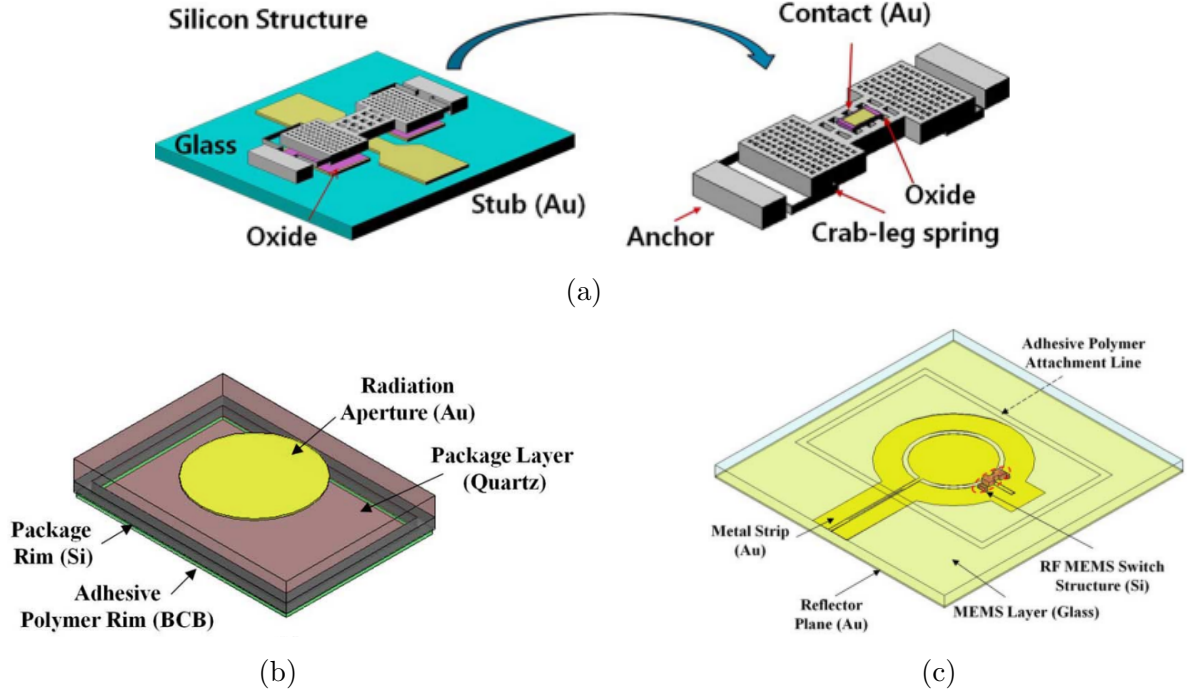


Fig. 2.9: Circular/linear polarization re-configurable antenna using RF-MEMS. (a) Structure of RF-MEMS switch used for polarization tuning within the antenna. (b) Top layer consisting of circular patch. (c) Bottom layer consisting of slot ring and RF-MEMS switch [9]

power. In 2012, Jung et al. Presented a polarization re-configurable antenna using RF-MEMS operating at the K-band [9]. This work consisted of a ring slot which is aperture coupled to a disc patch located above, which acts as the radiating element. The RF-MEMS switch, shown in Fig. 2.9a, is located on the ring slot and connected to a stub, which is orthogonal to the feed line and the ring slot. In total, two polarization states were achieved with this work: left-hand circular polarization and linear polarization. When the RF-MEMS switch is closed, the stub is connected to the ring slot and left-hand circular polarization can be achieved. Conversely, when the RF-MEMS switch is open, the stub is disconnected from the ring slot and linear polarization is achieved.

A drawback of this work is the packaging solution; this antenna is not built on PCB and requires special manufacturing consideration. The antenna consists of two layers:

a top layer containing the circular patch acting as the radiating element, shown in Fig. 2.9b, and the bottom layer containing a slot ring structure, shown in Fig. 2.9c. The top layer consists of a quartz substrate and gold, whereas the bottom layer consists of a glass substrate and gold. The top layer also acts as a protective cover for the RF-MEMS switch. The motivation behind this two layer approach is to shield the RF-MEMS switches from environmental factors such as temperature, humidity, physical shock and dust. Additionally, by having the RF-MEMS switches incorporated below the radiating disc, any unwanted parasitic behavior caused by poor soldering or wire bonding of the RF-MEMS switch will not have significant impact on the radiation performance of the antenna. Lastly, the RF-MEMS switch used in this work [Fig. 2.9a] is not commercially available; the fabrication of this specific RF-MEMS switch is based on a special SiOG process specially developed previously by the same group [23].

In 2004, Cetiner et al. presented a multi-functional re-configurable antenna using RF-MEMS switches [10] compatible with PCB technology. The proposed antenna structure is a “pixel-patch” antenna fed by a diagonal microstrip with a quarter-wave transformer and radial stubs, shown in Fig. 2.10. In Fig. 2.10, the blue boxes indicate the locations of the RF-MEMS turned on, and the red boxes indicate the location of the RF-MEMS switches turned off. The device is built on PCB and incorporates a novel RF-MEMS switches compatible with microwave laminates. The technology behind the PCB-compatible RF-MEMS switches and the relevant fabrication process was also developed by the same authors, which was presented previously in [24, 25]. This work by Cetiner et al. was able to achieve five polarization states at two frequencies (4.1 GHz and 6.4 GHz); a total of ten polarization states were realized, including: dual linear, linear X, linear Y, right-hand circular polarization and left-hand circular polarization. Similar to the work presented by Kim et al., a re-configurable matching input network is implemented to manage the changing input impedance across the different polarization states. Due to the microstrip feed inserted to the antenna structure at a 45° direction and by having all the RF-MEMS switches on, all the pixels are connected, and dual linear polarization states can be achieved. To achieve a single linear polarization state, shown in Fig. 2.10a and Fig. 2.10b, the RF-MEMS switches in the x-direction or the y-direction are turned off. Lastly, to realize circular polarization states, “slots” were created in the pixel antenna to generate X-polarized and Y-polarized modes of equal magnitudes and 90° phase difference. These “slots” were created by turning off some RF-MEMS switches and disconnecting certain pixels within the pixel-patch antenna, shown in Fig. 2.10c and Fig. 2.10d. Again, a limitation of this work is

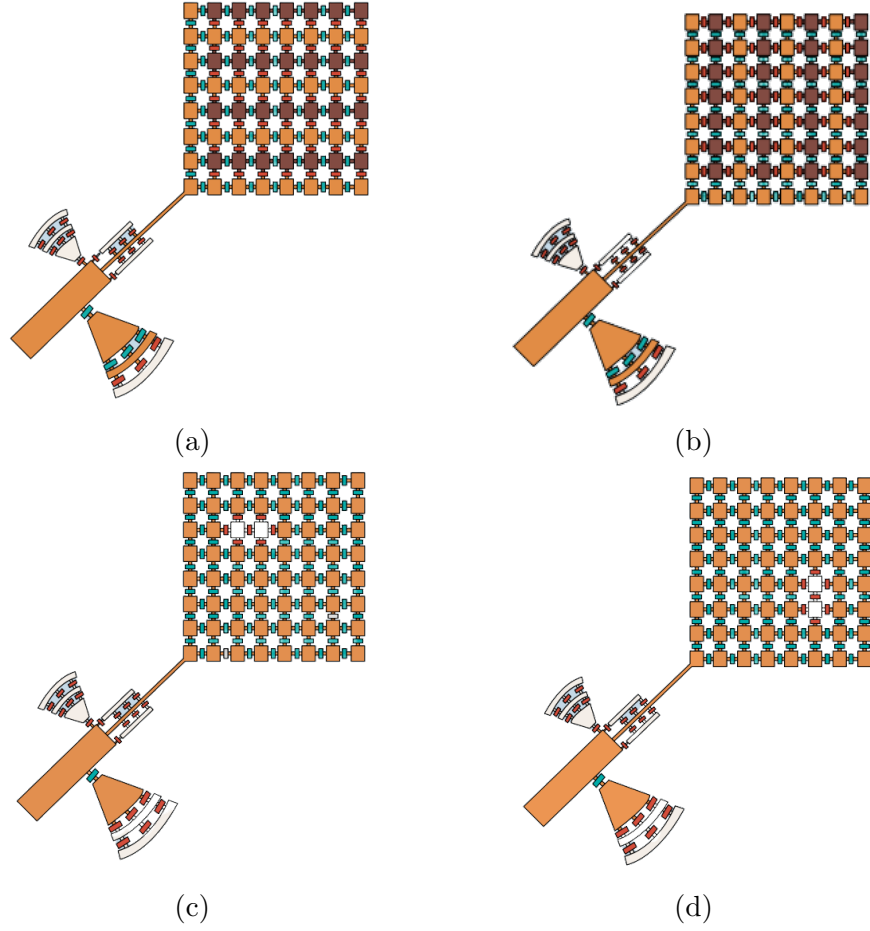


Fig. 2.10: Pixel patch antenna utilizing RF-MEMS switches. (a) Linear X-polarization state. (b) Linear Y-polarization state. (c) Right-hand circular polarization state. (d) Left-hand circular polarization state. [10]

the process behind the RF-MEMS switches used in this work is not commercially available and will have higher manufacturing costs.

Other works on polarization re-configurable antennas can be seen in Fig. 2.11. In Fig. 2.11a, Jaiswal et al. presented a 77 GHz RF-MEMS single-pole single-throw (SPST) based polarization re-configurable antenna able to achieve linear polarization and circular polarization states [11]. The design consists of a rectangular patch excited by a microstrip feed with four triangular shaped patches acting as parasitic elements for different polariza-

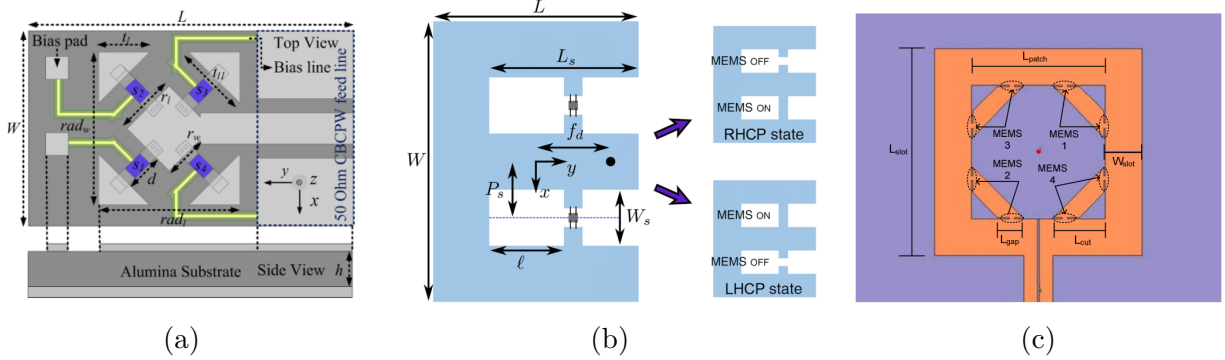


Fig. 2.11: Polarization re-configurable antenna works utilizing RF-MEMS switches. (a) Jaiswal et al. [11]. (b) Kovitz et al. [12]. (c) Balcells et al. [13].

tion states. Four SPST switches are used to connect the main rectangular patch with the adjacent four triangular parasitic patches. The work was fabricated on a $635\text{ }\mu\text{m}$ alumina substrate and the total area of the device is 0.794 mm^2 . Turning all the SPST switches on or off will produce linear polarization at 80 GHz or 70.5 GHz, respectively. The two circular polarization states are achieved at 77GHz by turning on a pair of the diagonally opposite switches, and turning off the other two switches. In Fig. 2.11b, Kovitz et al. presented a novel broadband re-configurable patch antenna array in 2015. The antenna element itself was an “E” shaped patch utilizing two RF-MEMS switches to switch between right-hand circular polarization and left-hand circular polarization, as shown in Fig. 2.11b [12]. This work was designed to operate at 2.4 GHz and was realized on a two layer PCB stack-up. However, because of the asymmetric “E” shape design of the antenna patch, the radiation pattern was not stable across the two polarization states and beam squint was observed in $\phi=0$ radiation plane. Lastly, seen in Fig. 2.11c is a polarization re-configurable MEMS-integrated patch antenna operating at 60 GHz presented by Balcells et al. [13]. The antenna was able to achieve linear polarization, right-hand circular polarization and left-hand circular polarization. In total, four pairs of RF-MEMS switches were incorporated into the design and were realized on a $200\text{ }\mu\text{m}$ synthetic quartz glass substrate. To obtain linear polarization, all the RF-MEMS switches are turned on or off; whereas to obtain circular polarization, the diagonal pairs of RF-MEMS switches are turned off and the other two are turned on. Nevertheless, the return loss of this antenna structure varies considerably across the different polarization states.

While many of the aforementioned works detailing polarization re-configurable antennas utilizing RF-MEMS switches have been reported to work in the mm-wave frequencies, these works still suffer from changing radiation pattern, return loss issues, or require special microfabrication facilities to realize.

Chapter 3

39 GHz Dual-Differential Feed Antenna Element

3.1 Project Scope

Following the discussions from previous chapters, it is evident a novel polarization re-configurable antenna operating at mm-wave frequency with stable input impedance and radiation pattern is needed. If RF front-end circuitry is used, a stable return loss of the novel antenna design is vital, otherwise antenna efficiency will be severely deteriorated, and worse, there is risk of permanently damaging PAs or LNAs connected to the antenna. The PA or low noise amplifier (LNA) within the beamforming integrated circuit (BFIC) are designed to perform optimally to a specific output impedance, and manufacturers will usually provide a VSWR rating where the BFICs can operate before being permanently damaged.

3.1.1 Design Requirements

The following provides a list of requirements regarding the performance of the proposed antenna design:

- The proposed antenna design must be able to achieve multiple polarization states: vertical-polarization state, horizontal-polarization state, left-hand circular polariza-

tion state and right-hand circular polarization state. Circular polarization has been shown to have greater immunity to interferences and multipath fading effects, which can be beneficial in 5G scenarios [26].

- The proposed antenna design must have a reflection coefficient less than -10 dB across the frequency range of 37 GHz to 40 GHz for all four different polarization states.
- The radiation pattern of the proposed antenna design should be stable across the different polarization states.
- The proposed antenna design when operating in circular polarization states should achieve 3 dB axial ratio bandwidth from 37 GHz to 40 GHz.
- The proposed antenna structure must be built on PCB technology that is commercially viable.
- The antenna structure should be somewhat compact to ensure enough PCB layers can be used for power and digital routing for RF-front end circuitry.

Given the above requirements, the following section describes the design process of the proposed novel antenna design to meet said requirements.

3.2 Antenna Design

The design of the following antenna structure was built on a 12-layer PCB using Isola Astra MT77 dielectric material and high-density interconnect (HDI) PCB technology. Astra MT77 was chosen as the PCB laminate and prepreg as it is compatible with HDI technology. Additionally, Astra MT77 is compatible for use at mm-wave frequency due to its of low dissipation factor $D_f = 0.0035$ and permittivity $D_k = 2.9$ at 40 GHz [27].

Shown in Fig. 3.1 is the 12-layer stack-up proposed by the PCB manufacturer using HDI any-layer technology. The proposed stack-up is defined as “5+N+5 any-layer”; the “5+N+5” indicates the stack-up consists of a single core and five layers of prepreg used above and below the core substrate, resulting in 5 cycles of lamination and both mechanically drilled via and laser drilled via are used. A via is an electrical connection between

two or more layers, and is essentially a small drilled hole, which is then plated with copper. “Any-layer” describes the stack-up as able to connect signals on any two layers together. Furthermore, a minimum area of copper is reserved to drill the via and is referred to as a

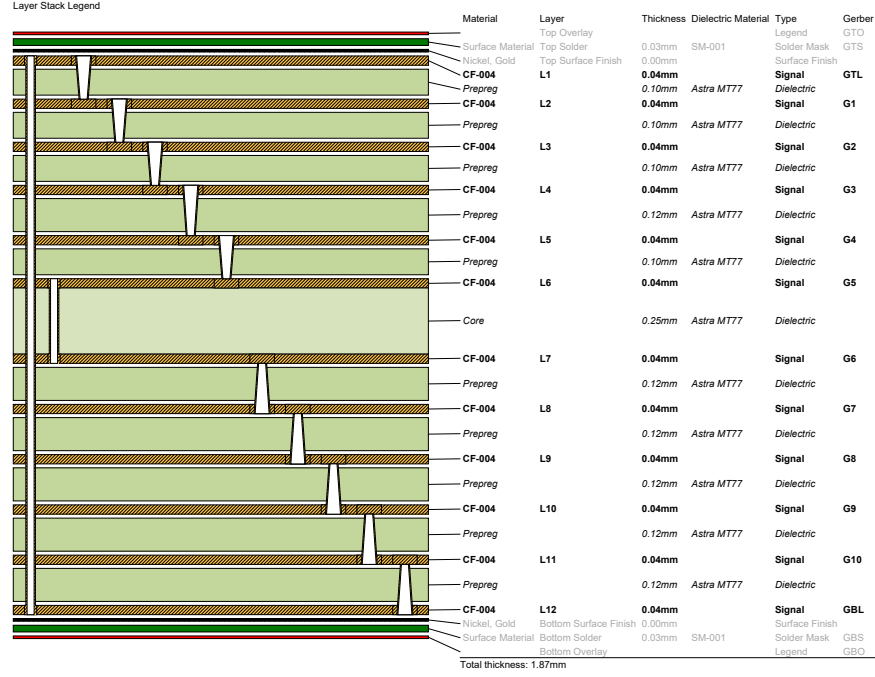


Fig. 3.1: 12-layer PCB stack-up using HDI process and stacked micro-via

via pad. Only two sets of via transitions are mechanically drilled: L1-L12 via and L6-L7 via, while all other sets of via transitions are laser-drilled micro-via. For example, in order to create a via transition from layer 8 to layer 6, a L8-L7 microvia is stacked onto a L7-L6 microvia to create a L8-L6 via transition. As a result, any via transition can be realized by stacking micro-via on top of each other. However, by stacking multiple micro-via together to create “one” via connection, there is the potential danger of misalignment between the stacked micro-via and seriously deteriorated RF performance. Another challenge faced by using the proposed stack-up is the low copper percentage in layer 2 to layer 6; the lower copper percentage will cause the stack-up to be asymmetrical and increase the chances of warpage during fabrication. Other additional constraints from the manufacturer include the size of the via; for the micro-via, the radius must be $3mil$ and via pad radius must be $7mil$, whereas for the through-hole via, the radius must be $4mil$ and via pad radius must

be $9mil$. The pad to copper clearance for through-hole via is $3mil$ and for micro-via is $6mil$ for inner layers of the stack-up and $8mil$ for layer 1 and layer 12.

An initial fabrication of an antenna structure with no RF front-end circuitry was proposed for two reasons: to test the viability of the dual-differential feed antenna concept, but equally important, to also test the reliability of the new PCB manufacturer and HDI any-layer process, especially using stacked micro-via. The term “passive antenna” is used in this thesis to refer to antenna structures without RF front-end circuitry connected. In contrast, the term “active antenna” refers to an antenna structure with RF front-end circuitry connected. The following section describes the design methodology of the passive antenna design, which is directly connected to surface mount connectors. Afterward, the passive design is modified to an active design by modifying the antenna feeds to connect directly to RF front-end circuitry.

To design the proposed novel antenna structure, Ansys high-frequency structure simulator (HFSS) was used. In HFSS, cylinders were used to model the via transitions. For the micro-via, the cylinders had a radius of $6mil$ and a pad size of $12mil$. For the through-hole via, the cylinders had a radius of $8mil$ and a via pad of $18mil$. Realistically, the micro-via would resemble a cone more so than a cylinder, as seen in Fig. 3.1. However, within this work, the micro-via was assumed to be more cylindrical than conical and was modelled as such. Lastly, the properties of the dielectric material Astra MT77 were initially assumed to be the same as the Astra MT77 material properties found within HFSS default materials, where the dissipation factor $D_f = 0.0017$ and permittivity $D_k = 3$. The assumption is shown to be not a valid approximation by the work of Salski et al. [27]. In this work, the complex permittivity of various dielectrics were measured with a Fabry-Pérot open resonator in the frequency range of 20-110 GHz. Salski et al. reported that at 40 GHz, the dissipation factor $D_f = 0.0035$ and permittivity $D_k = 2.9$; these were the values used for the properties of the dielectric when designing the antenna in HFSS.

3.2.1 Passive Antenna Design

As previously mentioned, RF-MEMS and PIN-diodes can be used to create polarization reconfigurable antennas, but is not a suitable approach as a commercially available solution or for use at mm-wave frequencies. Alternatively, many antenna designs utilizing multi-feed to realize different polarization states have been reported in the literature by changing

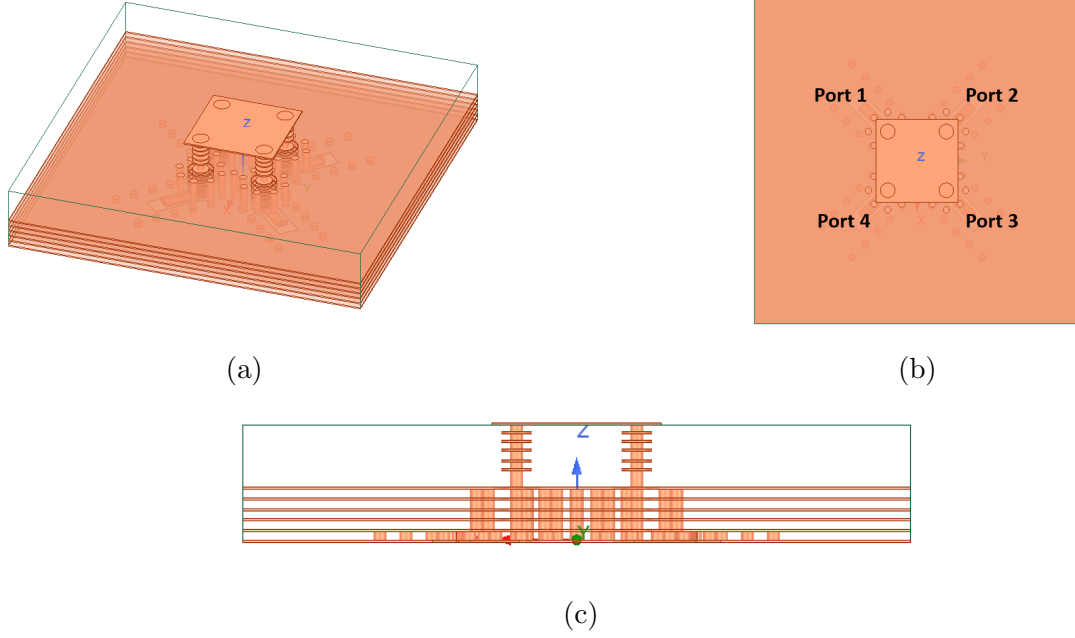


Fig. 3.2: Passive dual-differential feed antenna HFSS model. (a) Isometric view. (b) Top view with ports labelled. (c) Side view.

the phase delay to the individual feeds. Some works have realized the phase delays using LC technology [28, 29, 30] or BST technology [31, 32]. However, in this work, BFICs were used to realize the necessary phase delays due to availability.

Shown in Fig. 3.2 is the passive novel antenna design; a dual differential probe feed patch antenna constructed on the 12-layer HDI stack-up. The antenna structure consists of a square patch located on layer 1 and is fed by four coaxial feeds realized by four mechanically drilled L1-L12 through-hole. Lastly, the four through-hole via transitions are connected to coplanar waveguide with ground (CPWG) on layer 12, which are then routed out to surface-mounted 2.4mm connectors (not shown in Fig. 3.2) at 45° angles to create an “X” shape feed on layer 12.

Fig. 3.2c depicts the side view of the passive antenna structure. The radiating square patch is located on layer 1, while layer 7 is used for antenna ground and the CPWG routing is realized on layer 12. The copper on layer 2 to layer 6 was etched away to create the necessary clearance from the antenna ground on layer 7 and the radiating square patch on

layer 1. Three sets of via transitions are used in this design: L1-L12, L7-L12, L11-L12. Via transition L1-L12 is used for the coaxial feed from the CPWG on layer 12 to the radiating patch on layer 1. The via transition L7-L12 is used as shielding for the coaxial feeding and to connect the RF grounds on different layers together. Lastly, the via transition L11-L12 is used as via shielding for the CPWG line on layer 12.

Because the antenna has multiple feeds, the input impedance S_{11} is no longer a sufficient metric, as it does not take into account the influence of the other antenna feeds. Accordingly, a new overall input impedance expression is required to take into account the effects of the other antenna ports. The following derivations are from by Duan et al. on multi-port antenna design [33].

The S-parameter of the four port antenna structure can be described by Equation 3.1 as a function of frequency f . The S_{ii} denotes the single-ended return loss of each antenna input when the other three ports are terminated with 50Ω matched loads.

$$S_{antenna}(f) = \begin{bmatrix} S_{11}(f) & S_{12}(f) & S_{13}(f) & S_{14}(f) \\ S_{21}(f) & S_{22}(f) & S_{23}(f) & S_{24}(f) \\ S_{31}(f) & S_{32}(f) & S_{33}(f) & S_{34}(f) \\ S_{41}(f) & S_{42}(f) & S_{43}(f) & S_{44}(f) \end{bmatrix} \quad (3.1)$$

Seen in Fig. 3.2b, the locations of the four feeds are located and labelled. Equations 3.2a–3.2c are used to describe the symmetric and reciprocal nature of the four port antenna.

$$S_{11}(f) = S_{22}(f) = S_{33}(f) = S_{44}(f) \quad (3.2a)$$

$$S_{13}(f) = S_{31}(f) = S_{24}(f) = S_{42}(f) \quad (3.2b)$$

$$S_{12}(f) = S_{21}(f) = S_{23}(f) = S_{32}(f) = S_{34}(f) = S_{43}(f) = S_{14}(f) = S_{41}(f) \quad (3.2c)$$

Equation 3.1 can be rewritten as:

$$S_{antenna}(f) = \begin{bmatrix} S_{11}(f) & S_{12}(f) & S_{13}(f) & S_{12}(f) \\ S_{12}(f) & S_{11}(f) & S_{12}(f) & S_{13}(f) \\ S_{13}(f) & S_{12}(f) & S_{11}(f) & S_{12}(f) \\ S_{12}(f) & S_{13}(f) & S_{12}(f) & S_{11}(f) \end{bmatrix} \quad (3.3)$$

Assuming the input power to each antenna port is the same, the input voltage to the four port antenna can be written as:

$$V^+ = [e^{j\theta_1} \quad e^{j\theta_2} \quad e^{j\theta_3} \quad e^{j\theta_4}]^T \quad (3.4)$$

Consequently, using Equation 3.3 and Equation 3.4, the reflected voltage V^- of the four port antenna can be written as:

$$\begin{aligned} V^- &= S_{antenna}(f) \times V^+ \\ &= \begin{bmatrix} e^{j\theta_1} S_{11}(f) + (e^{j\theta_2} + e^{j\theta_4}) S_{12}(f) + e^{j\theta_3} S_{13}(f) \\ e^{j\theta_2} S_{11}(f) + (e^{j\theta_1} + e^{j\theta_3}) S_{12}(f) + e^{j\theta_4} S_{13}(f) \\ e^{j\theta_3} S_{11}(f) + (e^{j\theta_2} + e^{j\theta_4}) S_{12}(f) + e^{j\theta_1} S_{13}(f) \\ e^{j\theta_4} S_{11}(f) + (e^{j\theta_1} + e^{j\theta_3}) S_{12}(f) + e^{j\theta_2} S_{13}(f) \end{bmatrix} \end{aligned} \quad (3.5)$$

If the input signal to the four port antenna is dual differential, then the phase difference between signals at port 1 and port 3 will be 180° . Similarly, the signals at port 2 and port 4 will also be 180° out of phase. This relationship can be described by the following equations:

$$\theta_1 = \theta_3 + (2n - 1)\pi, \quad \theta_2 = \theta_4 + (2m - 1)\pi \quad m, n = 1, 2, 3 \dots \quad (3.6)$$

Simplifying Equation 3.5 using Equation 3.6 yields:

$$\begin{aligned}
V^- &= \begin{bmatrix} e^{j\theta_1} S_{11}(f) - e^{j\theta_1} S_{13}(f) \\ e^{j\theta_2} S_{11}(f) - e^{j\theta_2} S_{13}(f) \\ e^{j\theta_3} S_{11}(f) - e^{j\theta_3} S_{13}(f) \\ e^{j\theta_4} S_{11}(f) - e^{j\theta_4} S_{13}(f) \end{bmatrix} \\
&= (S_{11}(f) - S_{13}(f)) \begin{bmatrix} e^{j\theta_1} \\ e^{j\theta_2} \\ e^{j\theta_3} \\ e^{j\theta_4} \end{bmatrix} \\
&= (S_{11}(f) - S_{13}(f)) V^+
\end{aligned} \tag{3.7}$$

From Equation 3.7, the return loss of the four port antenna is shown to be:

$$S_{sd}(f) = \frac{V^-}{V^+} = S_{11}(f) - S_{13}(f) \tag{3.8}$$

where S_{sd} is the differential return loss of the four port antenna seen by each antenna port.

The result obtained in Equation 3.8 highlights two important factors when designing the proposed antenna regarding the differential return loss. Firstly, the differential return loss seen at any antenna port only depends on the single-ended return loss of the respective port and coupling of the antenna port located diagonally across. Secondly, because the signal to adjacent ports are differential, the coupling from one adjacent port will cancel out the coupling to the other adjacent port. Even if the adjacent port coupling is high, as long as both adjacent port couplings are equal in magnitude, the differential nature of the signal will negate the effects and not affect the overall differential return loss.

The locations of the four antenna coaxial feed on the square-shaped antenna patch can be seen in Fig. 3.2b. The overall design of the dual-differential antenna patch was kept as symmetric as possible so that the assumptions made in Equations 3.2a - 3.2c stay valid. Furthermore, by exciting the antenna with four signals with varying phases, different polarization states can be achieved by the antenna. The input signal for different polarization states are dual-differential; the relative phase difference between the adjacent ports can be varied, while the phase difference of 180° between the diagonal ports must

always be maintained. As long as the phase excitations satisfy the assumptions made in Equations 3.6, S_{sd} can be used to describe the differential return loss seen at all four antenna ports at every polarization state. Additionally, S_{sd} does not vary when switching across different polarization states.

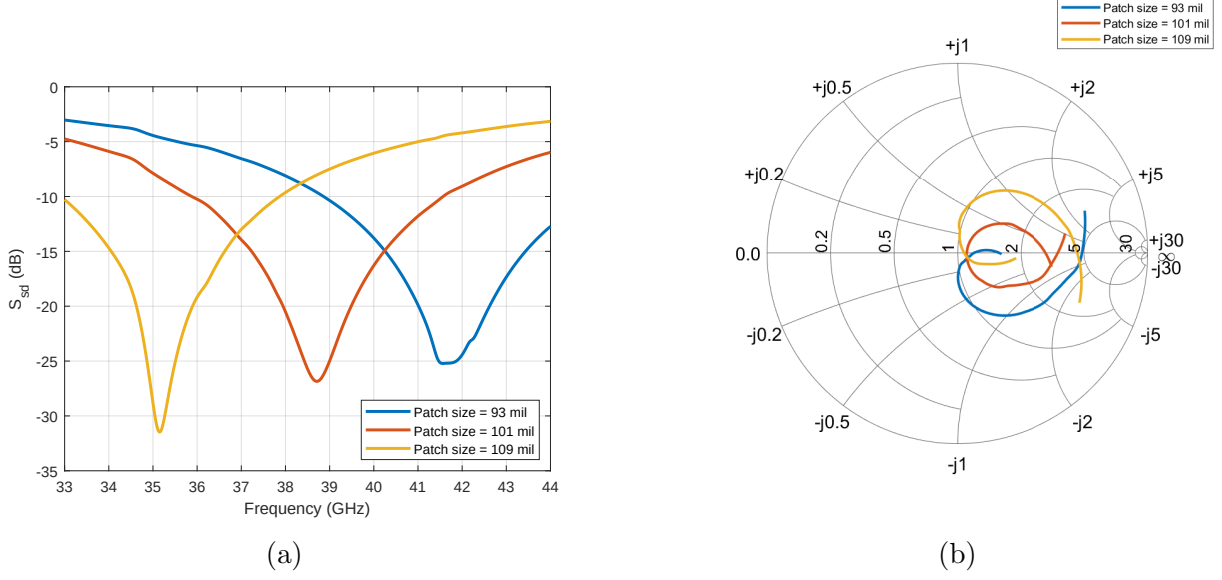


Fig. 3.3: S_{sd} of varying patch sizes with fixed coaxial feed positions. (a) S_{sd} in decibels. (b) S_{sd} on smith chart.

The design of the passive four feed antenna was optimized by tuning two parameters: the size of the patch antenna and the location of the four coaxial feeds on the patch

Parameter description	Value (mil)
Patch width	101
Patch length	101
Coaxial feed diameter	8
Coaxial feed pad diameter	18
CPWG feed width	10
Height between patch and ground	37.165

Table 3.1: Passive antenna design parameters

antenna. The antenna design has to maintain symmetry across the x-axis and y-axis, so the patch antenna was chosen to be square shaped. Additionally, the four antenna feeds were kept in the corners of the square patch. The optimized dimensions of the passive antenna structure are listed in Table 3.1.

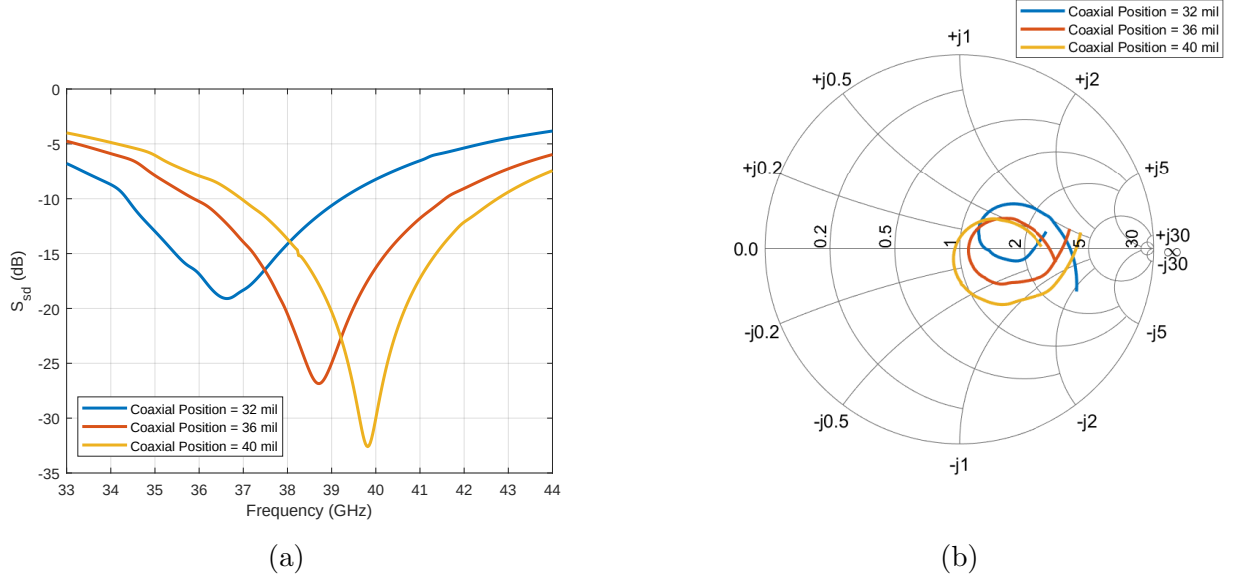


Fig. 3.4: S_{sd} of varying coaxial feed positions and fixed patch size. (a) S_{sd} in decibels. (b) S_{sd} on Smith chart.

By varying the size of the patch while keeping the location of the four coaxial feeds fixed, the center frequency of the differential return loss can be varied and tuned. In Fig. 3.3, the effects of varying the patch size is shown. By increasing the size of the patch and fixing the location of the four feeds, the center frequency of the differential return loss decreases. Whereas, if the size of the patch is made smaller, the center frequency of the differential return loss increases. The size of the ground plane is 200 mil, which is approximately, $1.3 \times \lambda_0$ and the edge of the ground plane is 15 mil from the edge of the patch antenna, which is approximately $0.5 \times \lambda_0$.

If the locations of the four coaxial feeds are varied and keeping the size of the square patch fixed, similar effects can also be seen in Fig. 3.4. By increasing the distance the location of the four feeds are from the center of the patch, the center frequency of the differential loss decreases. Additionally, changing the location of the coaxial feeds on the

patch antenna also affects how well-matched the differential return loss is, shown in Fig. 3.4a; on the Smith chart, increasing the coaxial feed location from the center of the patch results in a slight shift to the left, seen in Fig. 3.4b. However, there is a limit to how much the coaxial feed can be increased; the manufacturer requires a minimum of $4mil$ between the edge of the square patch and the edge of the coaxial feed.

The current distribution on the antenna was plotted to verify the correct polarization was realized. In Fig. 3.5, the current distributions of the two linearly-polarized states are shown. For the horizontally-polarized state, the phase input to the four ports are:

$$\theta_1 = \theta_4 = 0, \theta_2 = \theta_3 = 180^\circ \quad (3.9)$$

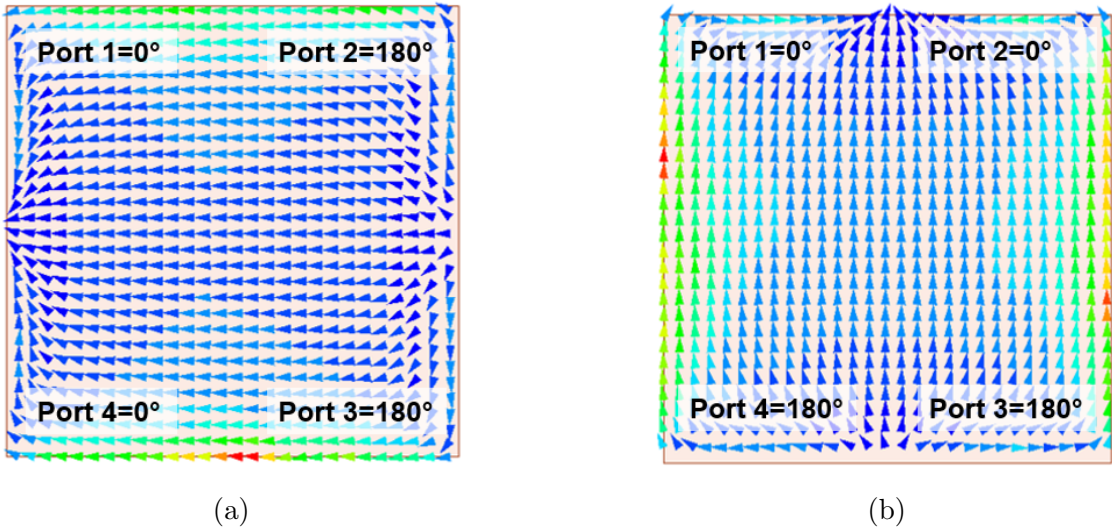


Fig. 3.5: Current distribution of patch antenna for linearly polarized states. (a) Vertically polarized. (b) Horizontally polarized.

The current flows along the x-direction in Fig. 3.5a, which is desired. Alternatively, to excite the vertically-polarized state, the phase input to the four ports are:

$$\theta_1 = \theta_2 = 0, \theta_3 = \theta_4 = 180^\circ \quad (3.10)$$

Referring to Fig. 3.5b, the direction of current flow is in the direction parallel to the y-axis, which is the desired mode of excitation. By changing the excitation phase to each of the

four ports, two different linearly polarized states can be achieved by the dual-differential antenna design and without the use of PIN diodes or RF-MEMS.

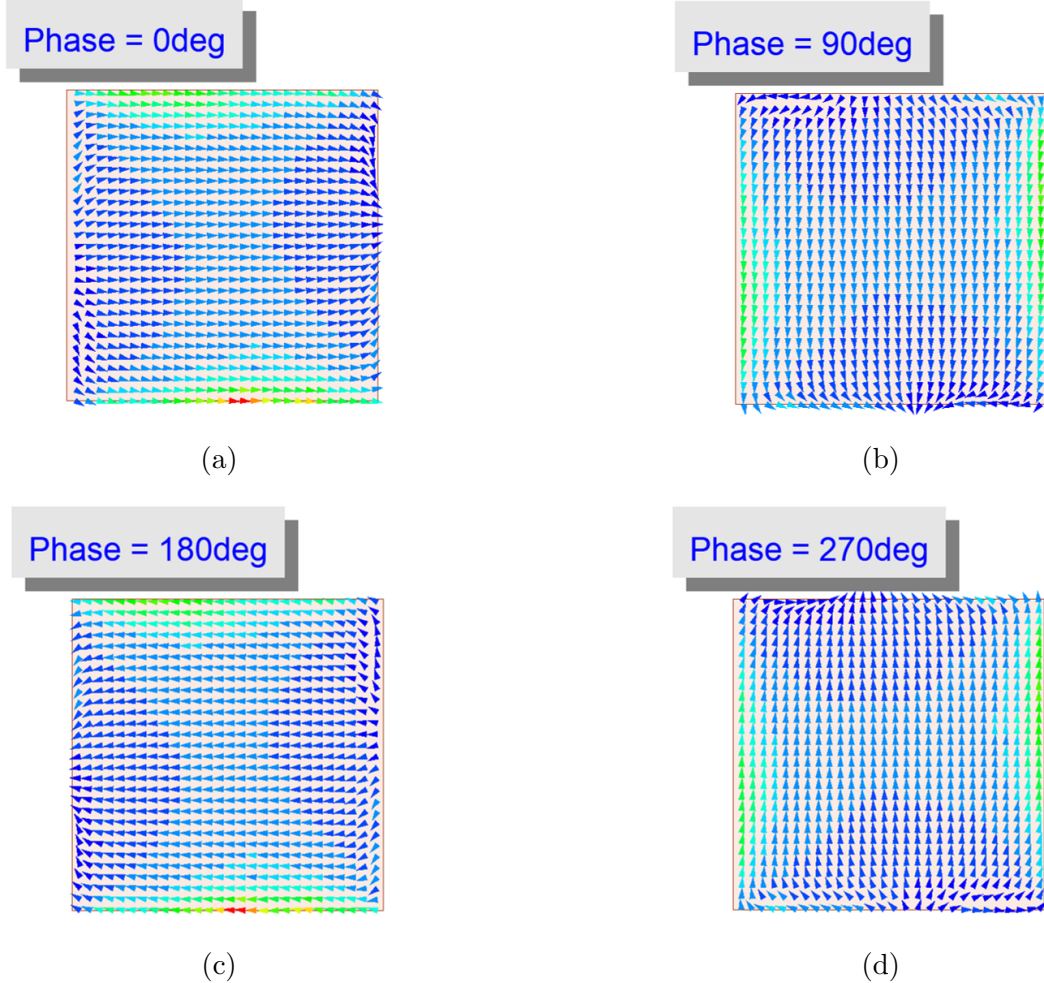


Fig. 3.6: Current distribution of patch antenna for LHCP state at different phases.

To realize circular polarization states, the phase inputs to the four port antenna is again dual-differential; however, the phase difference between adjacent ports is 90° . For example, to realize LHCP using the four feed antenna, the input phase to each of the four ports is given as:

$$\theta_1 = 0^\circ, \theta_2 = 90^\circ, \theta_3 = 180^\circ, \theta_4 = 270^\circ \quad (3.11)$$

In Fig. 3.6, the distribution of the surface current of the patch is plotted across different points in time to verify the circular nature of the polarization and the correct direction of circular polarization. In Fig. 3.6a, the direction of current flow is along the positive x-direction. When the phase is 90° (or $t = T/4$), the direction of current flow has changed and now flows in the negative y-direction, shown in Fig. 3.6b. Next, at $t = T/2$, the direction of current flow changes from the negative y-direction to the negative x-direction in Fig. 3.6c. Lastly, at $t = 3T/4$, the direction of current flow is along the positive y-direction, which can be seen in Fig. 3.6d. The rotational nature of the current flow in the clock-wise direction over time indicates the LHCP mode was realized.

For completeness, the distribution of surface current of the antenna in RHCP is shown in Fig. 3.7. To realize RHCP using the four feed antenna, the input phase to each of the four ports:

$$\theta_4 = 0^\circ, \theta_3 = 90^\circ, \theta_2 = 180^\circ, \theta_1 = 270^\circ \quad (3.12)$$

where the differential property of the input signal between ports 1 and ports 3 is maintained, and also between port 2 and port 4 is maintained. In Fig. 3.7, the distribution of the surface current of the patch antenna in RHCP is plotted at various times. Similar to Fig. 3.6, the rotational nature of the surface current across time can be seen in Fig. 3.7; however, the surface currents rotate in the counter-clockwise direction, which is the desired mode for RHCP excitation. To further verify the circular polarization states achieved by the dual-differential antenna, the axial ratio is plotted in Fig. 3.8a. The axial ratio achieved by the dual-differential antenna is below 0.05 dB across the entire frequency range of operation. The axial ratio exhibits identical performance between LHCP and RHCP because the two circular polarization states are achieved by changing the phases and the antenna structure itself does not change.

The required input phase excitation to the four feed antenna for the different polarization states is summarized in Table 3.2. The required phase excitation to each antenna port is dual-differential; the phase difference between port 1 and port 3 is 180° , while the phase difference between port 2 and port 4 is 180° .

Because the input signals for each polarization state satisfy the assumptions made in Equation 3.3, it is valid to use Equation 3.8 to describe the differential return loss seen by

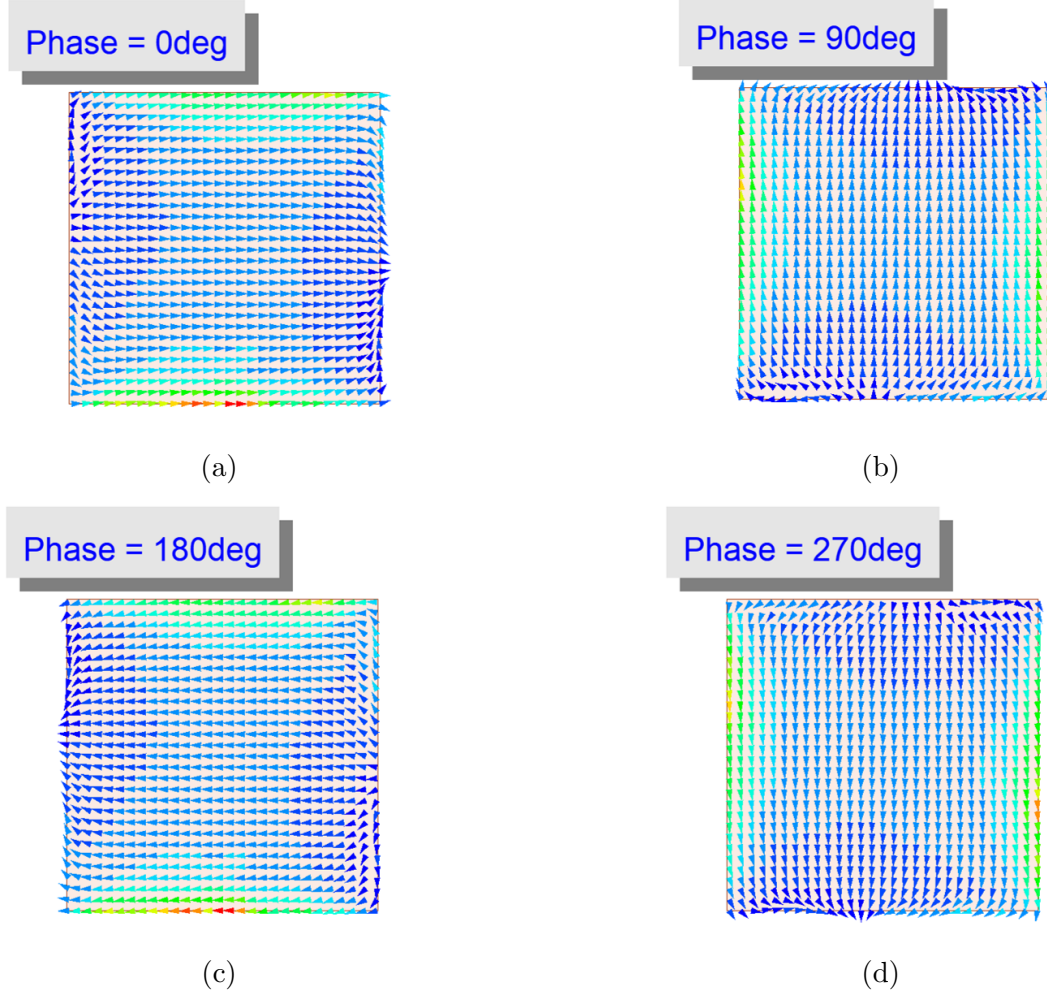


Fig. 3.7: Current distribution of patch antenna in RHCP state at different phases.

each antenna port in all polarization states. It is important to highlight that Equation 3.8 can be used in every polarization state as it is not input phase dependent, as long as the input signal is dual-differential.

Next, the S-parameters of the antenna are reported in Fig. 3.9a, where only the S_{1i} values are shown due to the symmetric nature of the antenna design. Only three traces can be seen in Fig. 3.9a because S_{12} and S_{14} are identical.

Because the proposed antenna design is optimized for dual-differential, the single ended

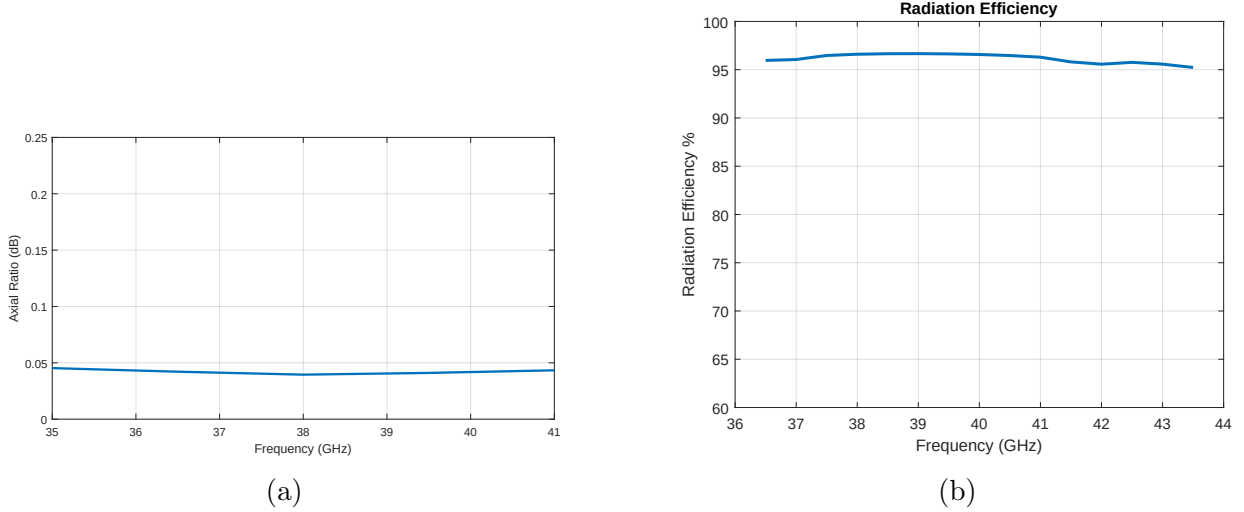


Fig. 3.8: (a) Axial ratio in LHCP and RHCP state. (b) Radiation efficiency for all four polarization states.

Polarization state	Port 1	Port 2	Port 3	Port 4
Vertically polarized	0°	0°	180°	180°
Horizontally polarized	180°	0°	0°	180°
LHCP	0°	90°	180°	270°
RHCP	270°	180°	90°	0°

Table 3.2: Antenna port excitation phases for different polarization states

S_{11} shown in Fig. 3.9a has poor matching (at most -11 dB) and narrow -10 dB bandwidth (36–39 GHz). The isolation between port 1 and the adjacent ports (port 2 and port 4) is poor, at -12 dB across the frequency range of operation. The isolation with the diagonal port, port 4, is also poor, which ranges from -10 dB and -14 dB. However, using Equation 3.8 to plot the S_{sd} of the antenna design, the antenna design exhibits good matching at -26 dB and the -10 dB bandwidth ranges from 36–41.5 GHz, shown in Fig. 3.9b. Lastly, the axial ratio of the antenna design in circular polarization state is plotted in Fig. 3.8a. Both LHCP and RHCP exhibit identical axial ratio performance across frequency; due to the multiple feeds, the antenna was able to achieve an axial ratio of less than 0.05 dB between 35–41 GHz.

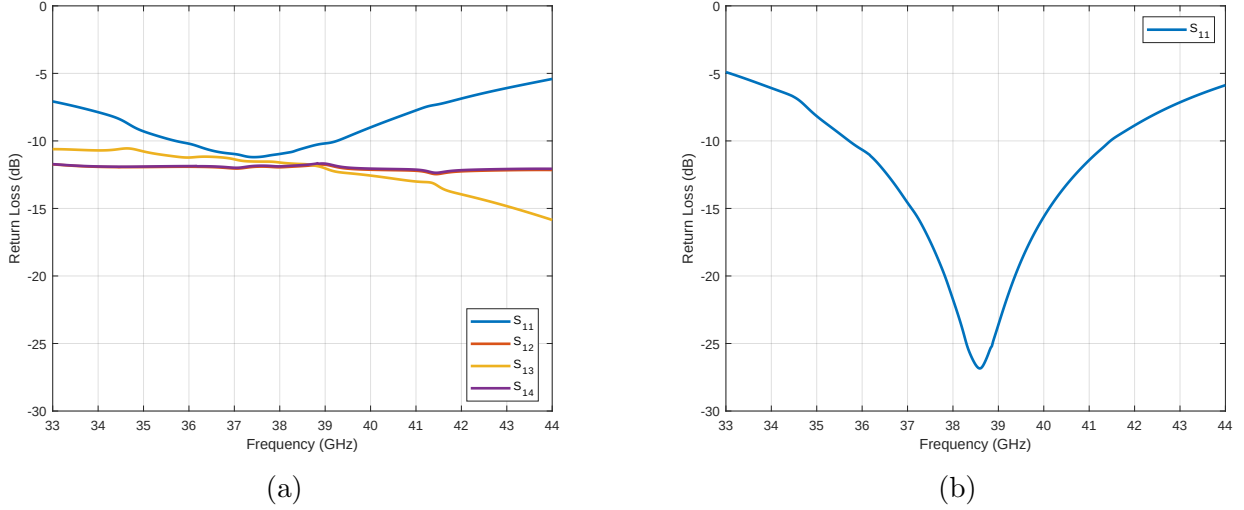
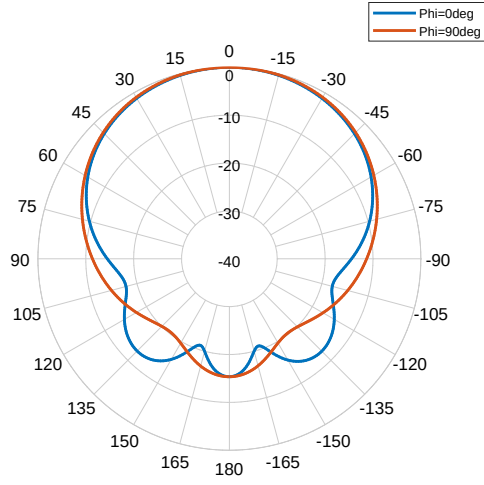


Fig. 3.9: (a) Single ended S_{1i} . (b) Differential return loss S_{sd} .

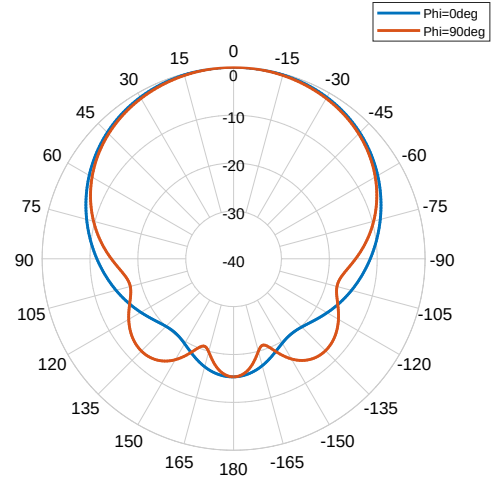
The normalized radiation pattern of the four feed antenna in the four different polarization states are shown in Fig. 3.11 at $\phi = 0^\circ$ plane and $\phi = 90^\circ$ plane. For the vertically polarized state in Fig. 3.11b, the HPBW in the $\phi = 0^\circ$ plane is 96° and in the $\phi = 90^\circ$ plane is 84° . The sidelobe level in the $\phi = 90^\circ$ plane is -13.2 dB and in the $\phi = 0^\circ$ plane there is no sidelobe level, only the existence of the backlobe at -16 dB.

Radiation patterns for the horizontally polarized state are shown in Fig. 3.11b and are similar to the radiation pattern when in vertically polarized state; the only difference is the radiation pattern of vertically polarized state in the $\phi = 0^\circ$ plane is equivalent to the radiation pattern in horizontally polarized state in the $\phi = 90^\circ$ plane and vice versa. In Fig. 3.11c, the radiation pattern for RHCP state in $\phi = 0^\circ$ plane and $\phi = 90^\circ$ plane are shown to be identical. The HPBW of the radiation pattern in RHCP state is 92° and the sidelobe level is -15.5 dB in both ϕ planes. The radiation pattern for LHCP state is shown in Fig. 3.11d and is identical to the radiation pattern in RHCP state. From Fig. 3.11, the dual-differential antenna exhibits stable radiation pattern across the different polarization states; the radiation pattern does not change significantly when operating in different polarization states.

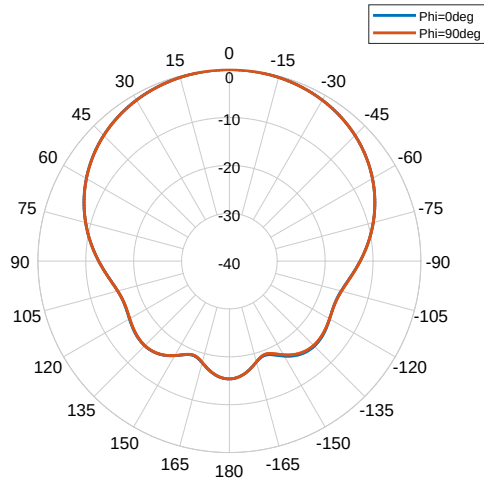
To conclude this section, a passive patch antenna with four feeds was designed and optimized. The four feed antenna was designed with a dual-differential input signal and by changing the phase excitation of the input signal, different polarization states were



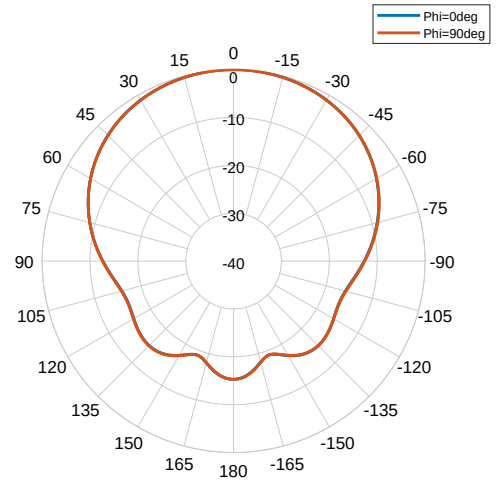
(a)



(b)

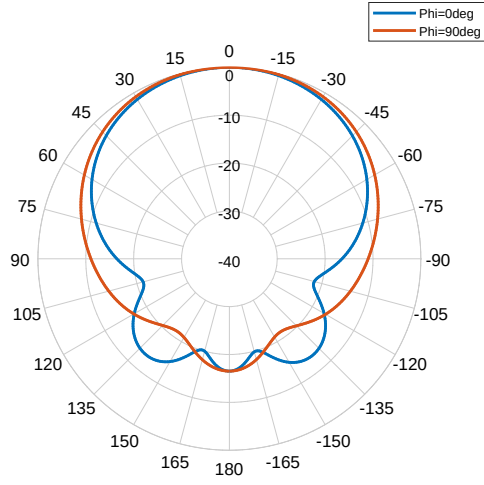


(c)

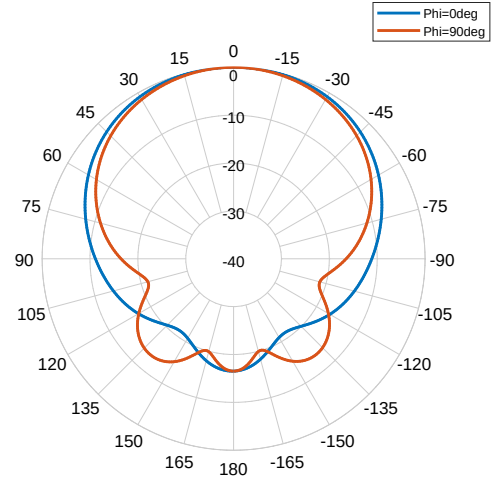


(d)

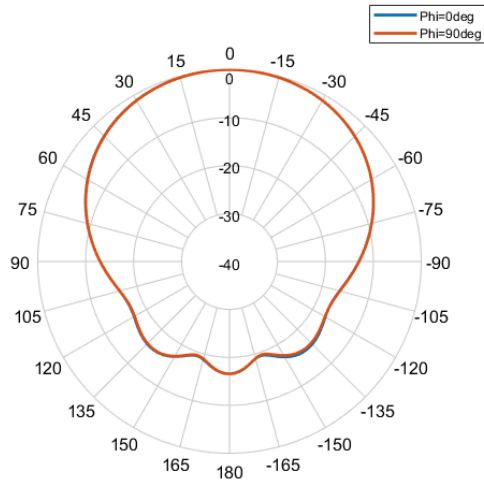
Fig. 3.10: Radiation pattern of dual-differential feed antenna at 37 GHz in (a) horizontally polarized state, (b) vertically polarized state, (c) RHCP state, and (d) LHCP.



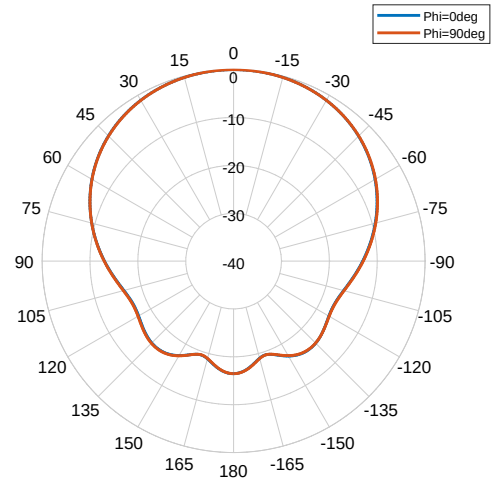
(a)



(b)

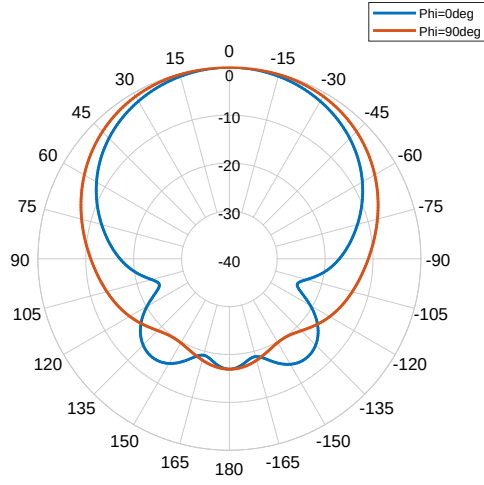


(c)

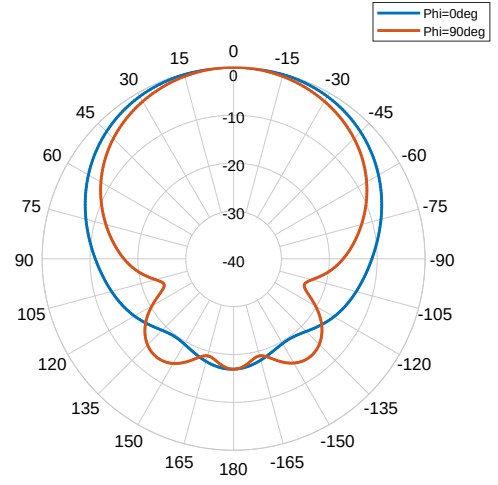


(d)

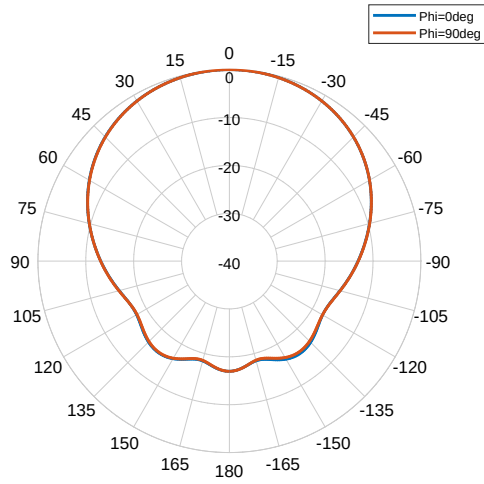
Fig. 3.11: Radiation pattern of dual-differential feed antenna at 38.5 GHz in (a) horizontally polarized state, (b) vertically polarized state, (c) RHCP state, and (d) LHCP.



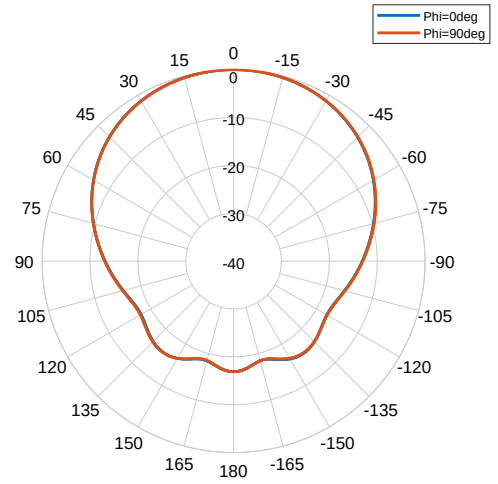
(a)



(b)

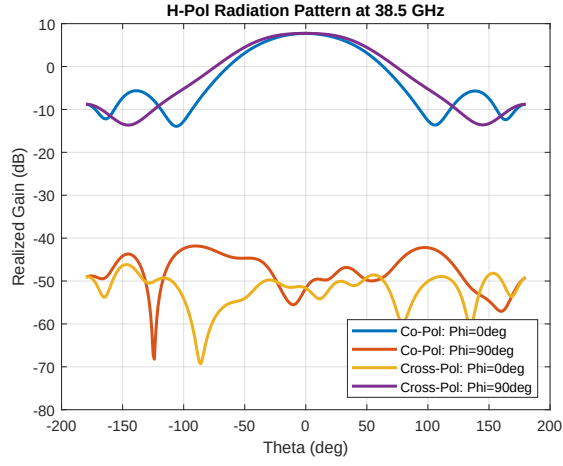


(c)

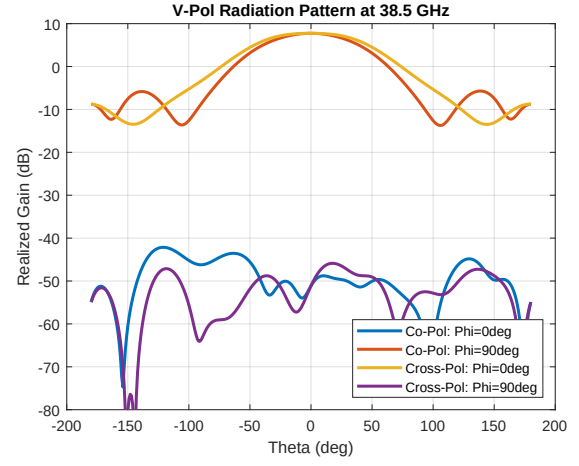


(d)

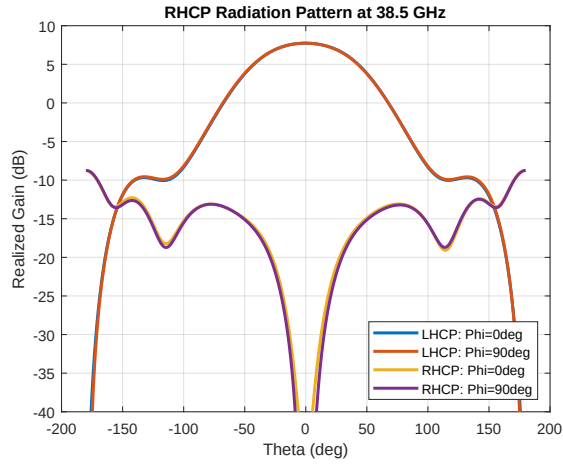
Fig. 3.12: Radiation pattern of dual-differential feed antenna at 40 GHz in (a) horizontally polarized state, (b) vertically polarized state, (c) RHCP state, and (d) LHCP.



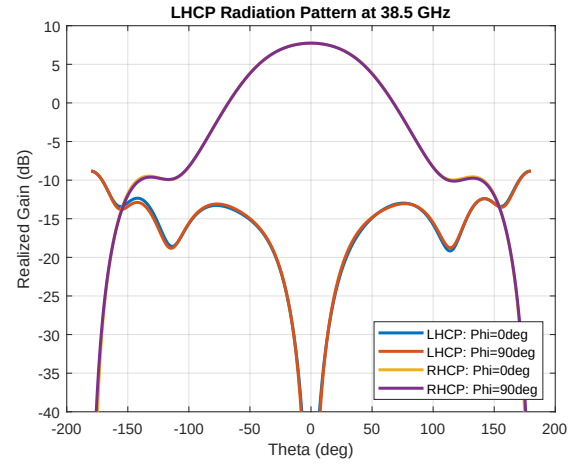
(a)



(b)



(c)



(d)

Fig. 3.13: Radiation pattern of passive dual-differential feed antenna at 38.5 GHz in (a) horizontally polarized state, (b) vertically polarized state, (c) RHCP state, and (d) LHCP.

achieved. In the next section, an active version of the four feed antenna will be presented and analyzed. The challenges faced in modifying the passive antenna design to an active antenna design will be discussed in the next section.

3.2.2 Active Antenna Design

A commercially available beamforming integrated circuit (BFIC) from Anokiwave was used; the AWMF-0169 is a 39 GHz Quad Core IC that operates at a frequency range of 37-40 GHz, supports four radiating elements and operates in Tx/Rx half duplex operation. The AWMF-0169 IC total footprint size is 3.6 mm x 3.6 mm, requires two different supply voltages for power and serial-peripheral-interface (SPI) protocol for digital control. The AWMF-0169 uses wafer level chip scale package (WLCSP) and will be mounted on layer 12 of the PCB. Unlike the traditional use case of the AWMF-0169, where it is used to connect to a 2×2 sub array of single-feed antennas, a single AWMF-0169 will be used to feed the four feeds of the dual-differential antenna design.

The active design uses the same stack-up as the passive design; a 12-layer 5+N+5 Astra MT77 HDI any-layer HDI stack-up will also be used for the active design. Because the BFIC will be flip-chip assembled on layer 12 directly beneath the patch antenna located on layer 1, the passive antenna design presented in the previous chapter cannot be used because the coaxial feeds from layer 1 to layer 12 will be located within the BFIC footprint. Consequently, the coaxial feeds must transition from layer 1 to any layer between layer 8 to layer 10, to avoid the BFIC on layer 12. Lastly, the antenna feed will transition to layer 12 from the intermediary layer and routed to the RF pins of the BFIC. Conversely, because the design involves a BFIC, other challenges are presented, such as the routing of other signals, including power supplies and digital signals. Because the AWMF-0169 requires two different voltage supplies, two of the layers will be used as power planes. Layer 11 must be used as RF ground for routing any CPWG lines on layer 12.

The proposed active antenna design is shown in Fig. 3.23. The design is a modification of the passive antenna design presented in the previous chapter; a square patch is located on layer 1 and four coaxial feeds are located in the corners of the square patch. However, instead of a through-hole via from layer 1 to layer 12, seven micro-vias are stacked to create a transition from layer 1 to layer 8. The motive behind this is to create space on layer 12 to allow the BFIC to be assembled. On layer 8, the four antenna feeds are routed out at

angles of 45° to form an ‘X’ shape, as highlighted in Fig. 3.14c. The length of this routing is a new parameter that can be used to optimize the active antenna design. Next, three micro-vias are stacked to create a transition from layer 8 to layer 12 whereby on layer 12, the four antenna feeds are routed into the pins of the BFIC to form a ‘H’ shape.

From Fig. 3.14, it is evident that the design is no longer fully symmetric due to the CPWG routing on layer 12 to the RF pins of the BFIC; the symmetry is only maintained across the x-axis or y-axis but not along the 45° line of symmetry. As a result, the differential return loss derived in Equation 3.8 is no longer applicable to the active antenna and needs to be modified to account for the asymmetry of the design. The following is similar to the derivation presented in the previous chapter, except the coupling between adjacent ports are no longer all equal. The following equations are used to describe the symmetric and reciprocal nature of the active antenna design.

$$S_{11}(f) = S_{22}(f) = S_{33}(f) = S_{44}(f) \quad (3.13a)$$

$$S_{13}(f) = S_{31}(f) = S_{24}(f) = S_{42}(f) \quad (3.13b)$$

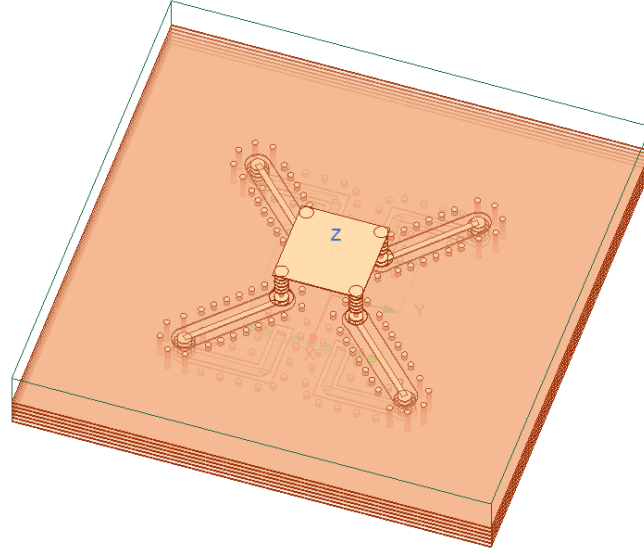
$$S_{12}(f) = S_{21}(f) = S_{34}(f) = S_{43}(f) \quad (3.13c)$$

$$S_{23}(f) = S_{32}(f) = S_{14}(f) = S_{41}(f) \quad (3.13d)$$

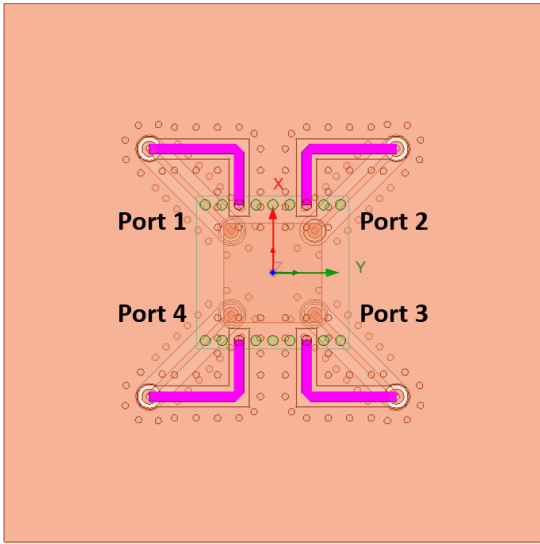
Equation 3.13c describes the horizontal port coupling, which includes the coupling between port 1 and port 2 or the coupling between port 3 and port 4. Conversely, Equation 3.13d describes the vertical port coupling, which includes the coupling between port 1 and port 4 or the coupling between port 2 and port 3.

The following S-parameters can be used to describe the four-port active antenna.

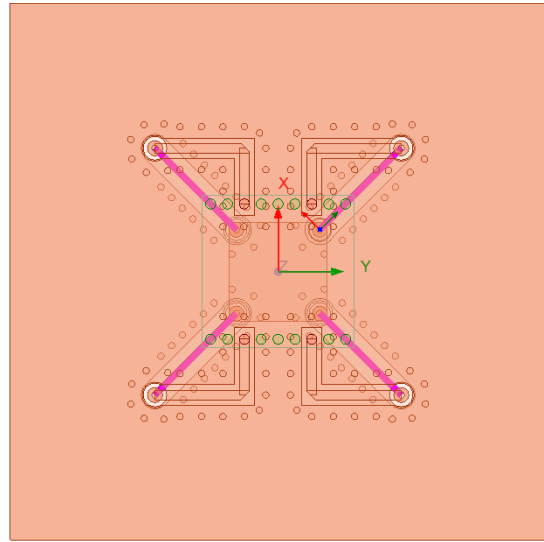
$$S_{antenna}(f) = \begin{bmatrix} S_{11}(f) & S_{12}(f) & S_{13}(f) & S_{14}(f) \\ S_{12}(f) & S_{11}(f) & S_{14}(f) & S_{13}(f) \\ S_{13}(f) & S_{14}(f) & S_{11}(f) & S_{12}(f) \\ S_{14}(f) & S_{13}(f) & S_{12}(f) & S_{11}(f) \end{bmatrix} \quad (3.14)$$



(a)



(b)



(c)

Fig. 3.14: Active dual-differential feed HFSS model. (a) Isometric view. (b) Layer 12 routing. (c) Layer 8 routing.

The input signal V^+ to the four-port antenna is the same as the input signal used in the passive antenna analysis and also can be described by Equation 3.4. To find the reflected wave V^- from the four-port active antenna design, the S-parameter matrix in Equation 3.14 in conjunction with Equation 3.6, since the input signal will also be dual-differential, to form the following:

$$\begin{aligned}
V^- &= S_{\text{antenna}}(f) \times V^+ \\
&= \begin{bmatrix} e^{j\theta_1} S_{11}(f) + e^{j\theta_2} S_{12}(f) + e^{j\theta_3} S_{13}(f) + e^{j\theta_4} S_{14}(f) \\ e^{j\theta_2} S_{11}(f) + e^{j\theta_1} S_{12}(f) + e^{j\theta_4} S_{13}(f) + e^{j\theta_3} S_{14}(f) \\ e^{j\theta_3} S_{11}(f) + e^{j\theta_4} S_{12}(f) + e^{j\theta_1} S_{13}(f) + e^{j\theta_2} S_{14}(f) \\ e^{j\theta_4} S_{11}(f) + e^{j\theta_3} S_{12}(f) + e^{j\theta_2} S_{13}(f) + e^{j\theta_1} S_{14}(f) \end{bmatrix} \\
&= \begin{bmatrix} (S_{11}(f) - S_{13}(f) + (S_{12}(f) - S_{14}(f))e^{j(\theta_2-\theta_1)})e^{j\theta_1} \\ (S_{11}(f) - S_{13}(f) + (S_{12}(f) - S_{14}(f))e^{j(\theta_1-\theta_2)})e^{j\theta_2} \\ (S_{11}(f) - S_{13}(f) + (S_{12}(f) - S_{14}(f))e^{j(\theta_4-\theta_3)})e^{j\theta_3} \\ (S_{11}(f) - S_{13}(f) + (S_{12}(f) - S_{14}(f))e^{j(\theta_3-\theta_4)})e^{j\theta_4} \end{bmatrix} \\
&= \begin{bmatrix} S_{11}(f) - S_{13}(f) + (S_{12}(f) - S_{14}(f))e^{j(\theta_2-\theta_1)} \\ S_{11}(f) - S_{13}(f) + (S_{12}(f) - S_{14}(f))e^{j(\theta_1-\theta_2)} \\ S_{11}(f) - S_{13}(f) + (S_{12}(f) - S_{14}(f))e^{j(\theta_4-\theta_3)} \\ S_{11}(f) - S_{13}(f) + (S_{12}(f) - S_{14}(f))e^{j(\theta_3-\theta_4)} \end{bmatrix} \circ V^+ \tag{3.15}
\end{aligned}$$

where $\circ$ is the Hadamard product. Equation 3.15 is similar to the result found in Equation 3.5; however, the assumption that S_{12} and S_{14} is no longer made and resulted in the extra term in Equation 3.15. Accordingly, a new differential return loss can be defined as

$$\begin{aligned}
S_{sd}^{active} &= \frac{V^-}{V^+} \\
&= S_{11}(f) - S_{13}(f) + (S_{12}(f) - S_{14}(f)) \begin{bmatrix} e^{j(\theta_2 - \theta_1)} \\ e^{j(\theta_1 - \theta_2)} \\ e^{j(\theta_4 - \theta_3)} \\ e^{j(\theta_3 - \theta_4)} \end{bmatrix} \\
&= S_{11}(f) - S_{13}(f) + (S_{12}(f) - S_{14}(f)) \begin{bmatrix} e^{-j(\Delta\theta_{12})} \\ e^{j(\Delta\theta_{12})} \\ e^{-j(\Delta\theta_{12})} \\ e^{j(\Delta\theta_{12})} \end{bmatrix} \tag{3.16}
\end{aligned}$$

where

$$\Delta\theta_{12} = \theta_1 - \theta_2 = \theta_3 - \theta_4 = 2k\pi \quad k = 1, 2, 3 \dots$$

Unlike the final result presented previously for the passive antenna design in Equation 3.8, the result presented in Equation 3.16 is no longer just a complex constant, but a vector. For simplicity, let

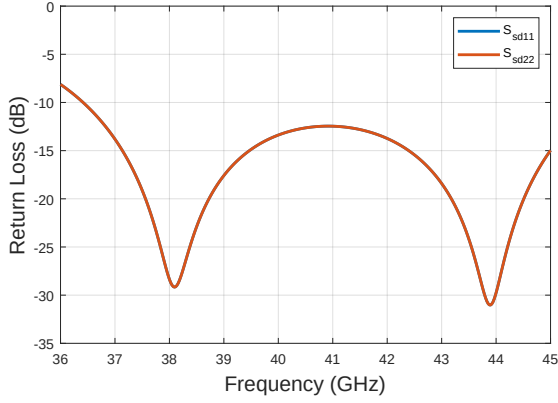
$$S_{sd11}^{active} = S_{sd,1}^{active} = S_{sd,3}^{active} = S_{11}(f) - S_{13}(f) + \Delta V^+(S_{12}(f) - S_{14}(f)) \tag{3.17a}$$

$$S_{sd22}^{active} = S_{sd,2}^{active} = S_{sd,4}^{active} = S_{11}(f) - S_{13}(f) + \Delta V^{+*}(S_{12}(f) - S_{14}(f)) \tag{3.17b}$$

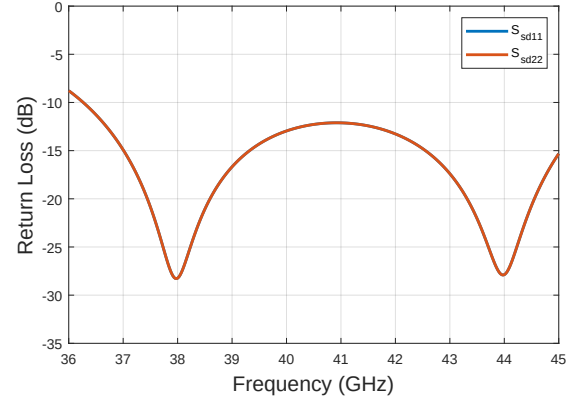
where $\Delta V^+ = e^{-j(\Delta\theta_{12})}$.

From Equation 3.17, two differential return loss will be seen by the BFIC depending on which port is being referred to. For port 1 and port 3, the differential return loss will be described by S_{sd11}^{active} , whereas for port 2 and port 4, the differential return loss will be described by S_{sd22}^{active} . As a consequence of the asymmetry of the active antenna design, each pair of the dual-differential ports will experience different return loss depending on the polarization state, unlike the passive antenna design where every antenna port experiences the same differential return loss, regardless of polarization state.

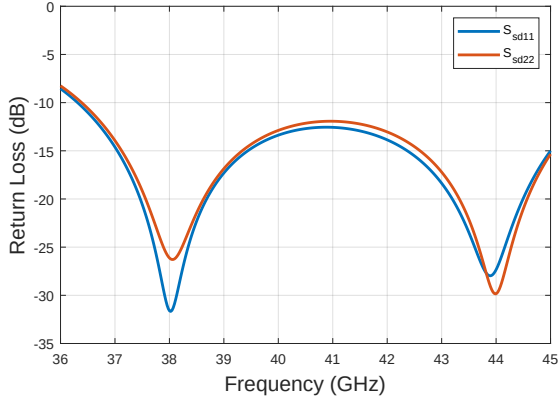
Shown in Fig. 3.15 are the two differential return loss, shown in Equation 3.17, of the active antenna design in different polarization states. Because the active antenna design



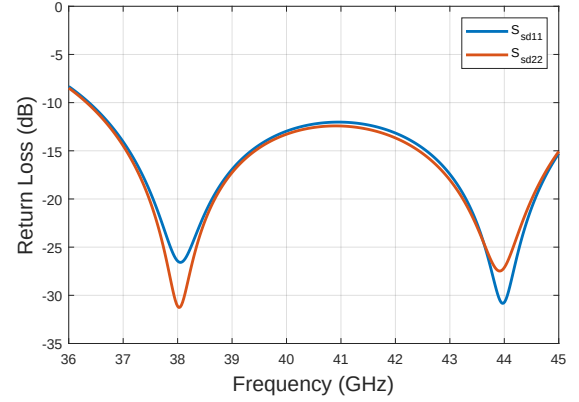
(a)



(b)



(c)



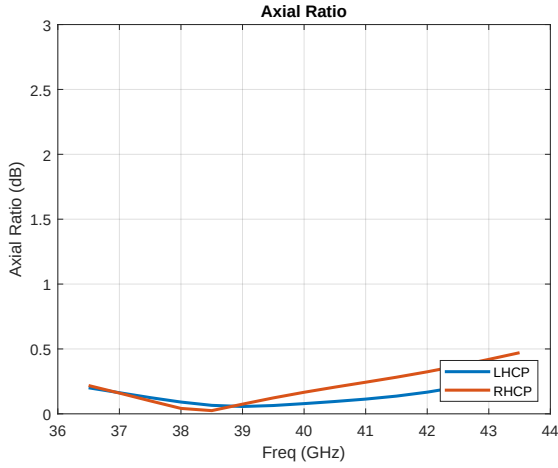
(d)

Fig. 3.15: Differential return loss of active antenna design in (a) horizontally polarized state, (b) vertically polarized state, (c) LHCP state, and (d) RHCP state.

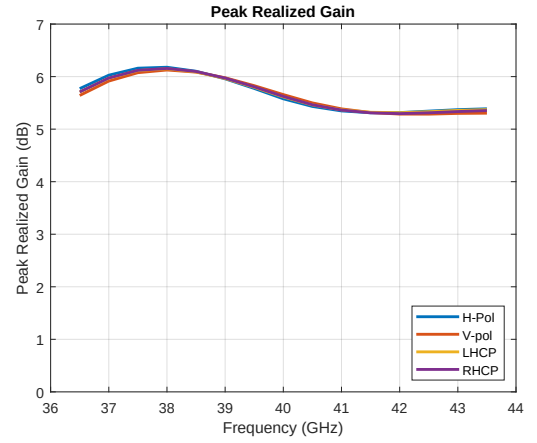
is no longer fully symmetrical due to the feeding on L12, the differential return loss varies depending on the different polarization states.

For the two linearly polarized states, the differential return loss varies slightly between the horizontally polarized state and the vertically polarized state, shown in Fig. 3.15a and Fig. 3.15b. However, within each linearly polarized state, the differential return loss for all four ports is the same. This is due to Equation 3.17a and Equation 3.17b being equal when the phase difference between port 1 and port 2 is either 0° or 180° , which results in ΔV^+ having a zero imaginary value. Alternatively, for the two circularly polarized states, two differential return loss traces can be seen in Fig. 3.15c and in Fig. 3.15d. The presence of

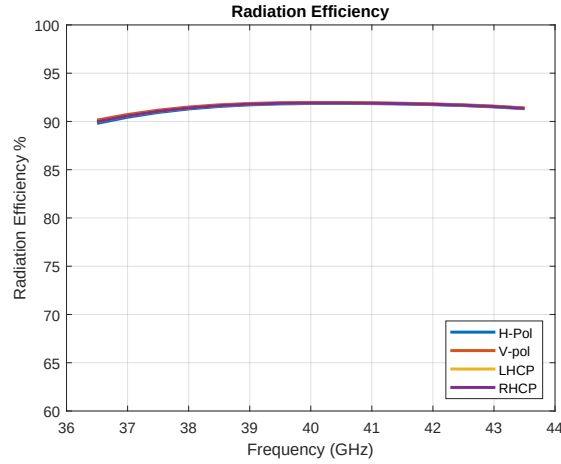
two different return loss when in either circular polarized states can also be explained by Equation 3.16: when the phase difference between port 1 and port 2 is either -90° or 90° , then ΔV^+ is a purely imaginary number, which means S_{sd11}^{active} and S_{sd22}^{active} will be different values. However, the differential return loss seen by each pair of ports are swapped when switching between the two circularly polarized states. For example, the S_{sd11}^{active} in LHCP state is the same as S_{sd22}^{active} in RHCP state, and vice versa.



(a)



(b)



(c)

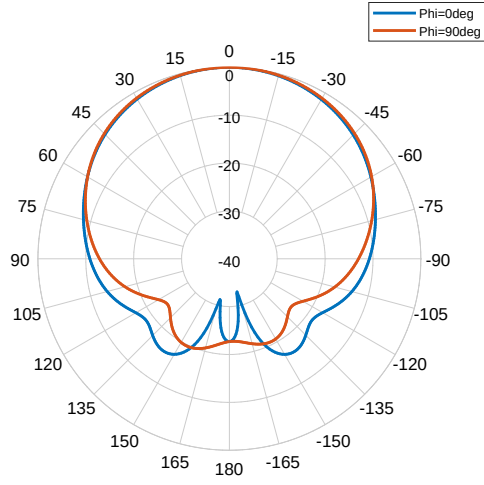
Fig. 3.16: (a) Axial ratio. (b) Peak antenna gain. (c) Radiation efficiency

The axial ratio of the active antenna design is plotted in Fig. 3.16a when operating in the circularly polarized states. The axial ratio at bore sight achieved by the design is below 0.5 dB across the entire frequency band of operation. In Fig. 3.16b, the peak antenna gain is shown for the four different polarization states. The peak antenna gain reaches 6 dB at 38 GHz and decreases to 5.3 dB at 41 GHz.

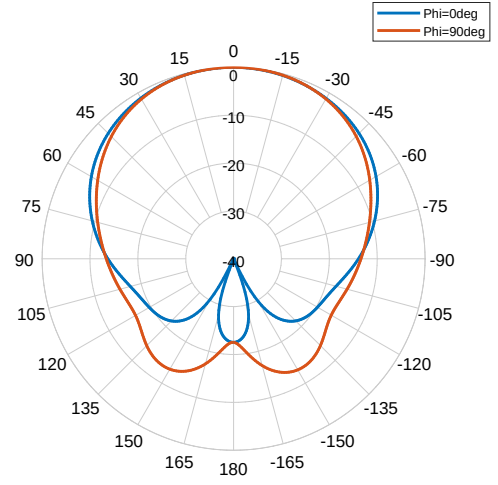
In Fig. 3.18, the radiation pattern of the active antenna is shown in the four different polarization states. In Fig. 3.18a, the radiation pattern of the active antenna in horizontally polarized state is shown. For the vertically polarized state, the radiation pattern is shown in Fig. 3.18b. The HPBW of the radiation pattern in Fig. 3.18b is 86° in the $\phi = 0^\circ$ plane, whereas in the $\phi = 90^\circ$ plane, the HPBW is 80° . For the horizontally polarized state, the radiation pattern is shown in Fig. 3.18a, where the HPBW is 88° in the $\phi = 0^\circ$ plane, whereas in the $\phi = 90^\circ$ plane, the HPBW is 96° . Unlike the passive antenna radiation pattern, the radiation patterns of the active antenna when operating in linear polarization states are no longer mirrors of each other. The radiation patterns in circularly polarized states are shown in Fig. 3.18c and in Fig. 3.18d. For both circularly polarized states, the HPBW is 91° and 88° in the $\phi = 0^\circ$ plane and $\phi = 90^\circ$ plane, respectively. Unlike the passive antenna design, the radiation pattern along $\phi = 0^\circ$ plane and $\phi = 90^\circ$ plane when operating in circular polarization are no longer the same. The radiation pattern of the proposed active design does not vary a lot across the different polarization states, which is ideal.

The current distribution of the antenna on the surface of the patch is shown in Fig. 3.21. For the vertically polarized state, the current distribution is shown in Fig. 3.21b, whereas the surface current distribution for the horizontally polarized state is shown in Fig. 3.21a. The same phase excitation is used to achieve the different polarization states in the active antenna design as the passive antenna design, as shown in Table 3.2.

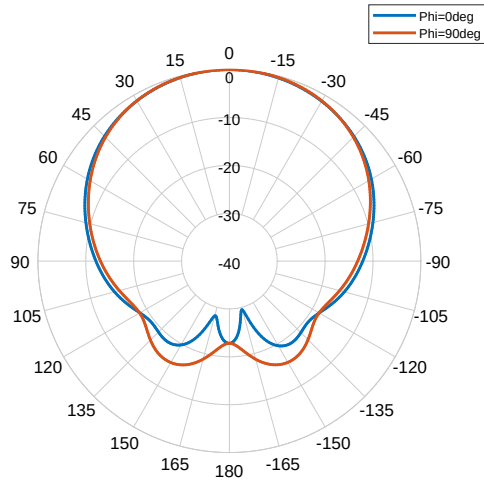
The current distribution of the patch in LHCP state is shown in Fig. 3.22 and also demonstrates the same performance as the passive antenna design when operating in LHCP state. Shown in Fig. 3.22 are the surface currents of the patch at different times; throughout one time period, the current distribution can be seen rotating in the clockwise direction, which is desired for LHCP mode. The current distribution of the patch in RHCP state is not shown, but exhibits similar performance as the LHCP state, except the current distribution rotates in the counter-clockwise direction. For completeness, the current distributions of the active antenna patch are included to ensure the modifications to the



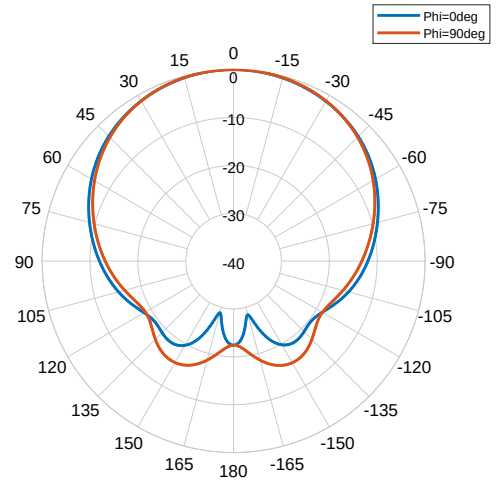
(a)



(b)

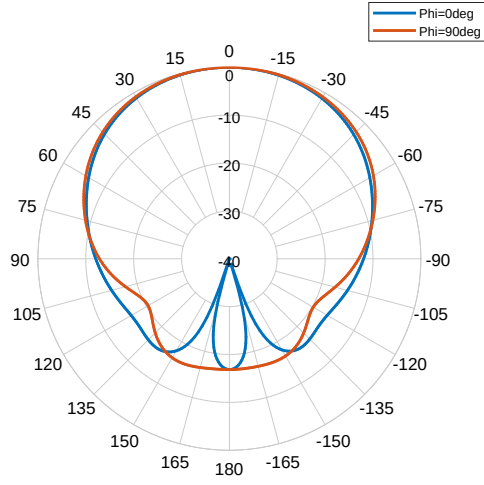


(c)

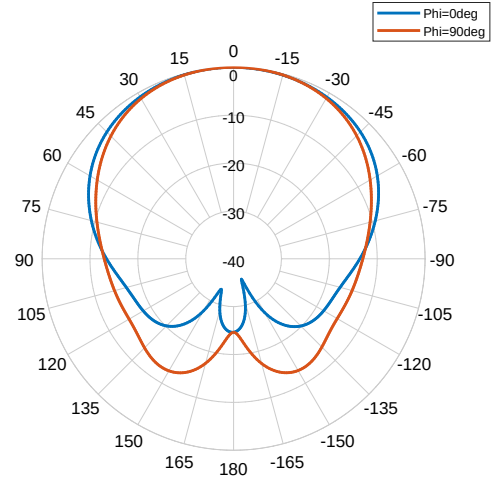


(d)

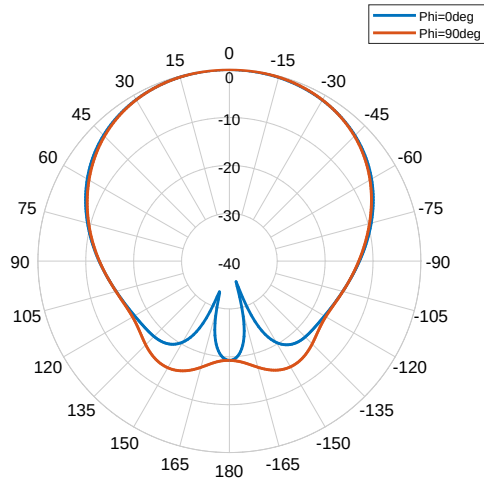
Fig. 3.17: Radiation pattern of active antenna design at 37 GHz in (a) horizontally polarized state, (b) vertically polarized state, (c) LHCP state, and (d) RHCP state.



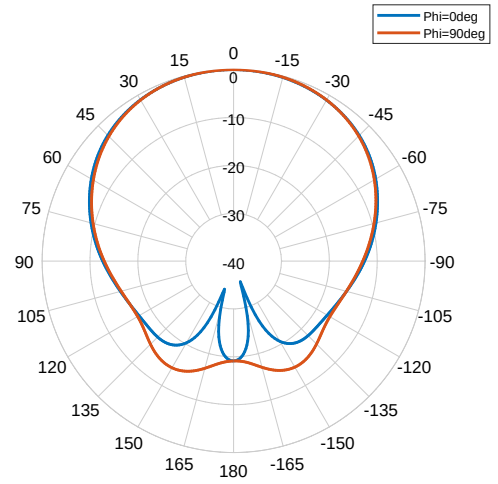
(a)



(b)

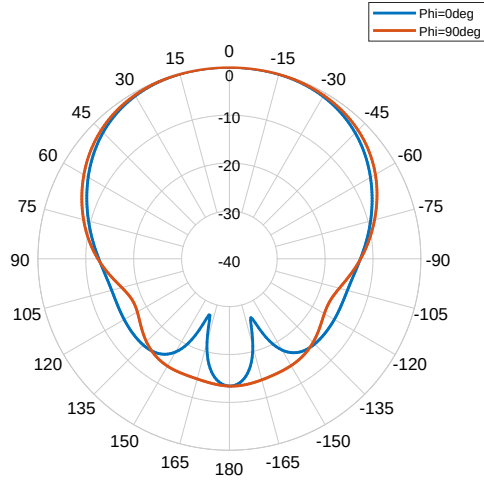


(c)

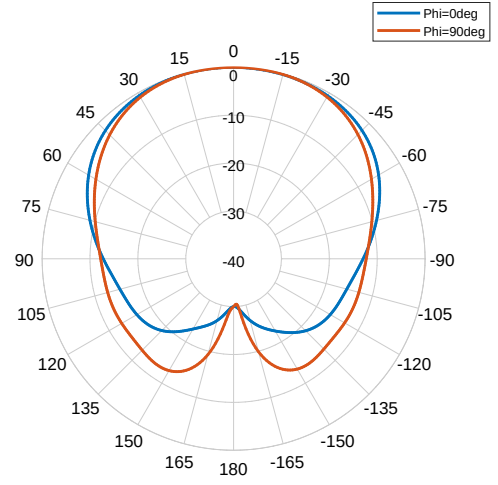


(d)

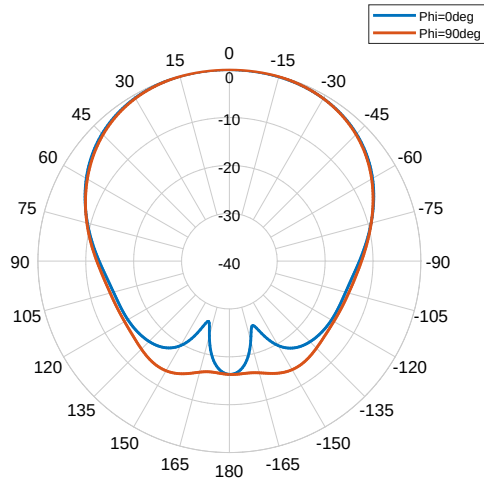
Fig. 3.18: Radiation pattern of active antenna design at 38.5 GHz in (a) horizontally polarized state, (b) vertically polarized state, (c) LHCP state, and (d) RHCP state.



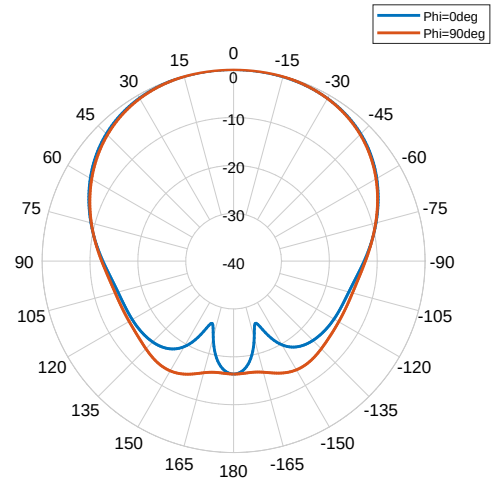
(a)



(b)



(c)



(d)

Fig. 3.19: Radiation pattern of active antenna design at 40 GHz in (a) horizontally polarized state, (b) vertically polarized state, (c) LHCP state, and (d) RHCP state.

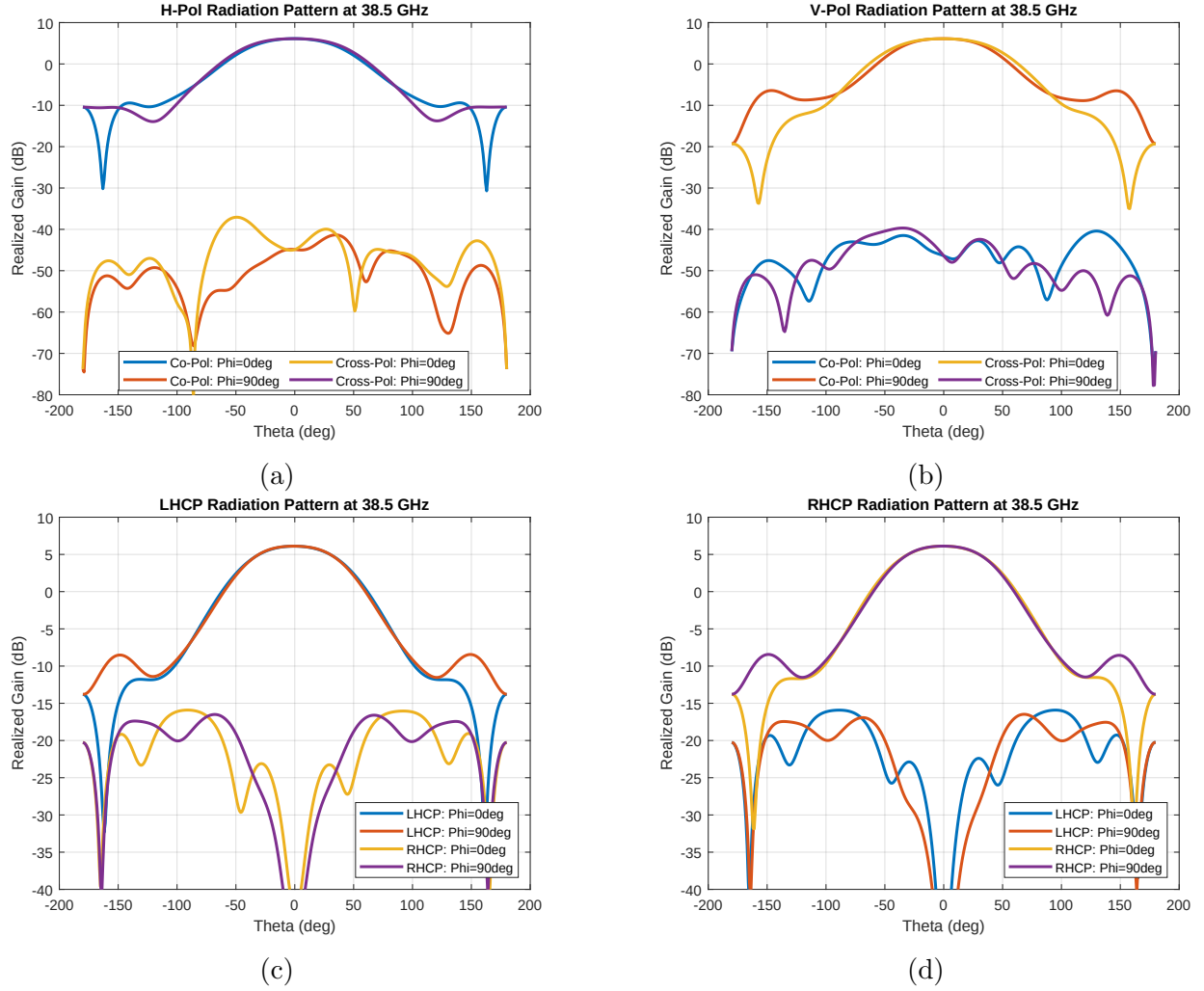


Fig. 3.20: Differential return loss of active antenna design in (a) horizontally polarized state, (b) vertically polarized state, (c) LHCP state, and (d) RHCP state.

antenna feeding on L8-L12 does not affect the polarization. From Fig. 3.21a - Fig. 3.22, the surface currents are similar to the surface currents of the patch of the passive antenna design presented previously, and can conclude the modified feeding for the active design does not affect the antenna polarization.

Lastly, the 3D view of bottom view and top view of the final PCB design is shown in Fig. 3.23. In Fig. 3.23a, the top side of the PCB is shown. The connector J3_X is used

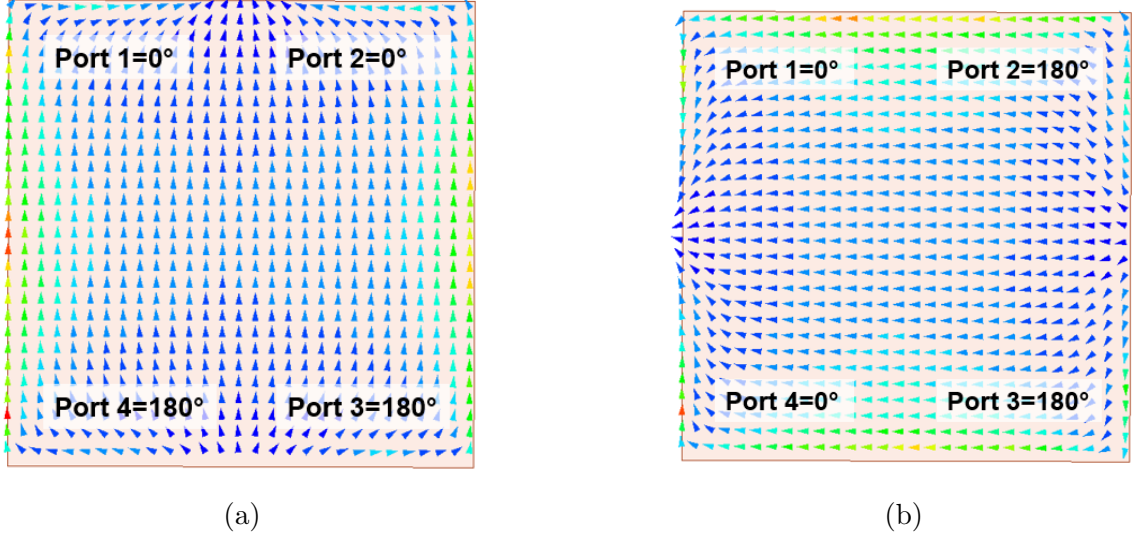


Fig. 3.21: Current distribution of patch antenna of active design in (a) vertically polarized state and (b) horizontally polarized state.

to route all the necessary digital signals required to control the BFIC, including SPI, Tx enable, and Rx enable. The connector J2_X is used to supply the two supply voltages, 1.8V and 2.5V, to the BFIC. Close to the aforementioned connector are decoupling capacitors, where the values were specified by the datasheet. Additional decoupling capacitors are located closer to the BFIC to ensure clean voltage supplies are provided to the BFIC.

A blue heat sink is attached on the topside of the PCB, where the BFIC is located. Two through holes are drilled to allow screws to slot through and fix the heat sink on top of the BFIC. Lastly, the connector J1_X is a 2.4 mm vertical launch RF connector used to feed the BFIC and uses two screws to fix the connector in place.

Moreover, shown in Fig. 3.23b, additional copper was added along the edges of layer 2 to layer 6 to increase the copper percentages in these layers in order to make the stack-up more symmetrical and reduce the chances of warpage during manufacturing. The additional copper on layer 2 to layer 6 is located more than $4\lambda_0$ distance away from the edge of the dual-differential antenna patch. The additional copper was simulated in HFSS and was confirmed to not affect the RF performance of the active antenna structure.

To conclude this section, an active antenna design based on the passive antenna design presented in the previous section was presented. The antenna was designed to operate

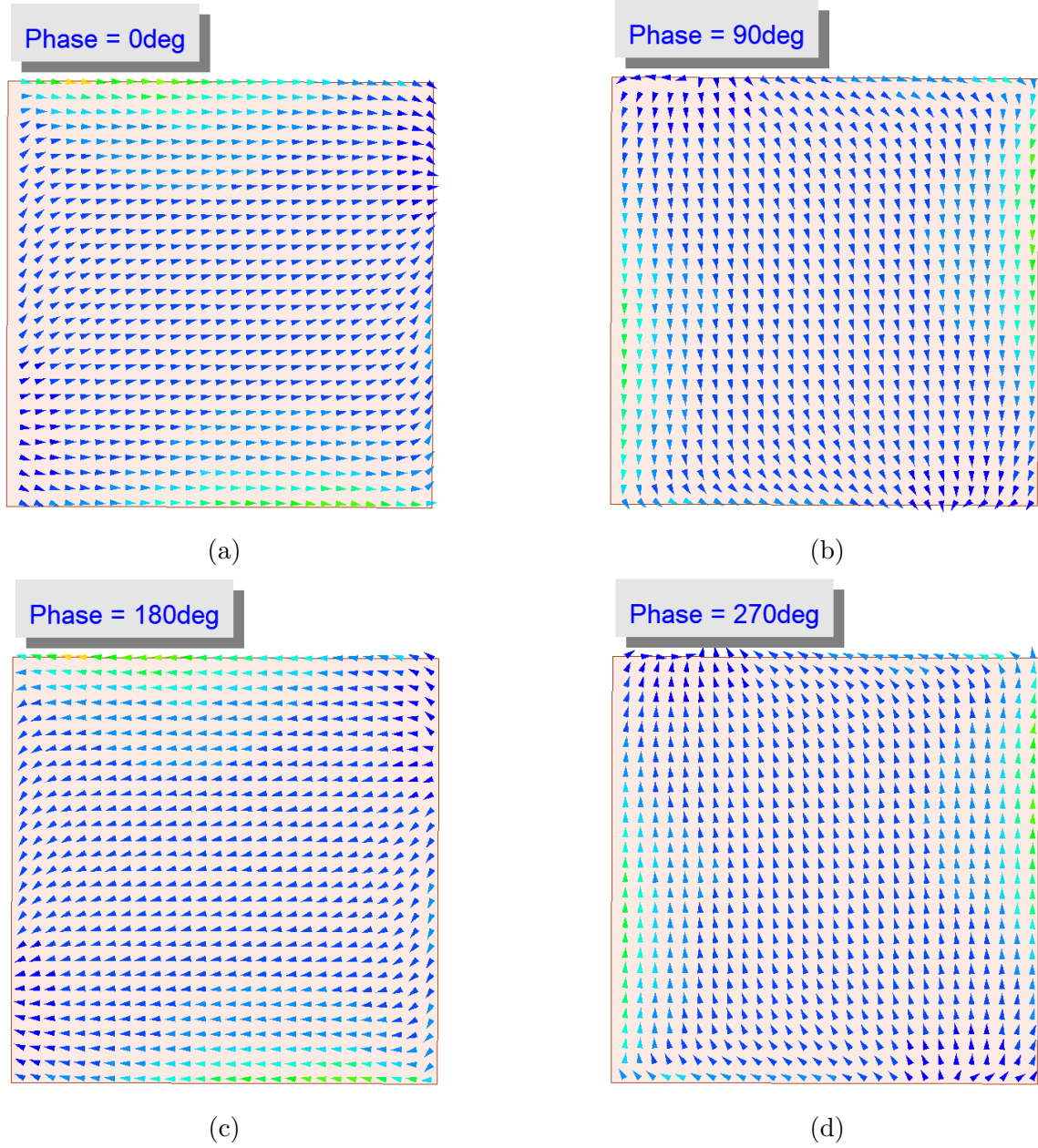


Fig. 3.22: Current distribution of patch antenna of active antenna for LHCP state at different phases.

between 37–40 GHz and is able to realize four different polarization states without significant changes to its radiation pattern. To accommodate the pin layout of the chosen BFIC, the proposed active antenna design is no longer symmetrical, which resulted in variance in the differential return loss across the different polarization states. Even though the differential return loss varied slightly between polarization states, the proposed active antenna design still exhibited good differential return loss across the frequency band of operation and stable radiation patterns.

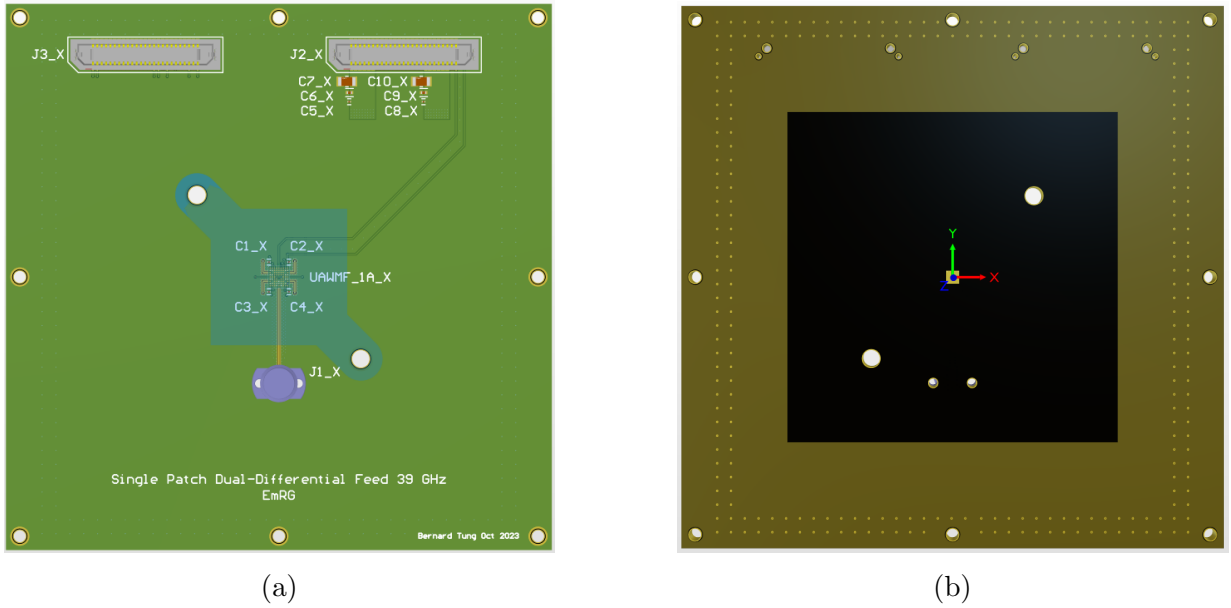


Fig. 3.23: 3D model of dual-differential antenna PCB. (a) Backside of assembled PCB with heat sink attached. (b) Front side of PCB with single dual-differential feed patch in the center.

Chapter 4

39 GHz Circular Phased Antenna Array using Dual-Differential Feed Antenna

To create an antenna array in a uniform rectangular configuration using the novel antenna design presented in the previous chapter proved to be difficult. To avoid the presence of grating lobes, the distance between the centers of each array element should be less than $\lambda_0/2$ in order to scan from 0° to 90° . However, because each array element is excited by a single quad-channel BFIC, the required clearance between adjacent BFIC required by the manufacturer is greater than the $\lambda_0/2$ at 38.5 GHz. For comparison, a 4×4 antenna array of single feed antenna elements requires only 4 quad-channel BFIC, the spacing between adjacent BFIC is approximately λ_0 , assuming the BFIC will be located at the center of each 2×2 subarray, which is more than the required manufacturer clearance.

Consequently, the use of circular array topology was explored to realize an antenna array of dual-differential antenna elements presented in the previous chapter to overcome the constrained antenna spacing without sacrificing the radiation pattern of the array by introducing the presence of grating lobes. Some works have used circular array topology to generate OAM signals at boresight [34] or radiate in the $\phi = 90^\circ$ direction for radar/sensing applications [35, 36, 37]. In [38], a 16-element uniform circular array operating at 2.4 GHz was realized for digital beamforming to generate N -beams. However, the use of uniform circular array topologies for beamforming applications at mm-wave frequencies is not as

widespread as uniform linear arrays or uniform rectangular arrays.

4.1 Circular Array

A diagram of N elements arranged in a circular array is shown in Fig. 4.1. The N elements are all located on the xy -plane and the origin is located at the center of the circular array. The radius a of the circular array describes the distance the center of the array is to the antenna elements.

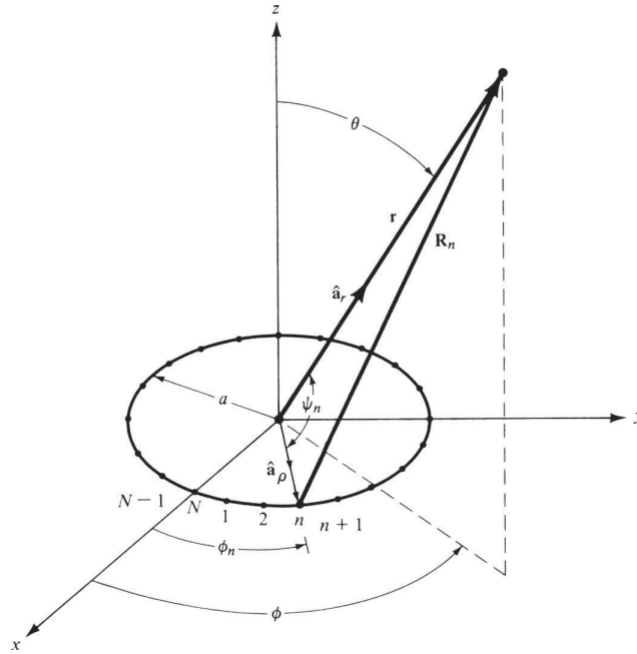


Fig. 4.1: Configuration of an N -element circular array [1]

The location of the n th element is located at an angular position of ϕ_n in the counter-clockwise direction from the positive x -axis, whereas ϕ and θ describes the direction the main beam is steered to.

The array factor of a circular array shown in Fig. 4.1 is given by

$$AF(\theta, \phi) = \sum_{n=1}^N I_n \exp(j[ka \sin \theta \cos(\phi - \phi_n) + \alpha_n]) \quad (4.1)$$

where I_n is the amplitude excitation of the n th element and α_n is the phase excitation of the n th element given by

$$\alpha_n = -ka \sin \theta_0 \cos(\phi_0 - \phi_n) \quad (4.2)$$

If assuming the circular array is uniform, or the elements are spaced equally on the circle, then $\phi_n = \frac{2n\pi}{N}$, which describes the angular location of the n th element on the circle. The first element is located at $\phi_n = \frac{2\pi}{N}$ and the last element is located at $\phi = 0$, or the positive x-axis.

Equation 4.1 can be modified to include the presence of a centre element located at the origin:

$$AF(\theta, \phi) = I_0 \exp(\alpha_0) + \sum_{n=1}^N I_n \exp(j[ka \sin \theta \cos(\phi - \phi_n) + \alpha_n]) \quad (4.3)$$

Moreover, Equation 4.3 can be further modified to describe multi-ring circular array configurations:

$$AF(\theta, \phi) = I_0 \exp(\alpha_0) + \sum_{m=1}^M \sum_{n=1}^N I_{n,m} \exp(j[ka_m \sin \theta \cos(\phi - \phi_{n,m}) + \alpha_{n,m}]) \quad (4.4)$$

where M is the total number of rings in the circular array, a_m is the radius of the m th ring, $\phi_{n,m}$ is the angular location of the n th element on the m th ring, and $I_{n,m}$ and $\alpha_{n,m}$ is the amplitude excitation and phase excitation of the n th element on the m th ring, respectively.

At a glance, an antenna array in a circular configuration can be seen to offer some advantages over uniform rectangular antenna arrays. Firstly, because the antenna elements are arranged in a circular configuration, none of the elements are considered edge or corner elements, which implies the coupling between the antenna elements are equal and the active return loss of every antenna element is equal. Another advantage of a circular antenna array is the ability to scan in the azimuth plane. A circular array is able to steer, for

example, in the $\phi = 0^\circ$ plane as well as the $\phi = 45^\circ$ plane without significant reduction in performance. This is due to the inherent symmetry of the circular array, which allows it to scan in the azimuth plane, unlike a rectangular array where beamsteering in the $\phi = 0^\circ$ plane and in the $\phi = 90^\circ$ plane results in similar performance, but not in the $\phi = 45^\circ$.

4.2 Array Design

In the following section, a parametric study is presented to explore the viability of planar circular arrays. Circular planar arrays for beam steering applications have not been widely implemented in the literature, unlike rectangular or linear arrays. As a result, the following section aims to explore how different parameters of a circular arrays affects the array factor, particularly when scanning in the azimuth. A circular planar array was chosen in order to relax the inter-element spacing to allow the placement of the BFIC on the bottom-side of the PCB without the presence of grating lobes when steering in the θ direction.

Previous works have studied the sidelobe levels of uniform circular arrays at boresight compared to the uniform rectangular arrays and linear arrays [39, 40, 41]. However, the behavior of sidelobe levels are not the same as uniform rectangular array or uniform linear array when beamsteering. Thus, further investigation into sidelobe levels when beamsteering using a circular array topology was required. A paper involving the optimization of multi-ring uniform circular array showed it was possible to beam steer while having low sidelobe levels [42]. However, this work involves many elements and large aperture size, which would not be useful for low number of elements and small sized prototype. An initial investigation into the array factor of uniform circular arrays, where the element spacing is $1.1 \times \lambda_0$ (required spacing due to BFIC size), was conducted and showed restricted scanning capabilities for low element count. However, as shown later, increasing the number of elements can reduce the sidelobe levels. Because the prototype is proof-of-concept, the number of elements will be low (6 elements). The goal of the following parametric studies is to determine whether the limited scanning angle and high sidelobe levels is due to the small number of elements, and if so, can it be improved by increasing the number of elements. The goal of the following parametric study is also to investigate the behavior of the sidelobe levels not at boresight, as previous works have already been conducted at boresight. The following studies were done at $\theta = 30^\circ$. The studies were also extended up to $\theta = 60^\circ$ and showed similar trends as $\theta = 30^\circ$, but increasing the scanning angle past

$\theta = 60^\circ$ showed poor radiation pattern regardless of increasing element count, array size, or both.

For all the following parametric studies, the normalized array factor of different circular array configurations are calculated and plotted in Matlab. The array factors are calculated without any amplitude tapering.

4.2.1 Increasing Number of Array Elements with Fixed Radius

The effects of changing the number of elements while keeping the radius of the uniform circular array fixed are shown in Fig. 4.2 and Fig. 4.3. In Fig. 4.2, the number of elements are even, whereas the number of elements in Fig. 4.3 is odd. The circular array is steered to $\theta = 30^\circ$ and swept in the azimuth, from $\phi = 0^\circ$ to $\phi = 90^\circ$. The radius of the circular array is fixed to λ_0 . Shown in Fig. 4.2a is the array factor of the circular array when the number of elements is six. The sidelobe levels in Fig. 4.2a vary considerably and, in some ϕ planes, increase to the same level as the main beam. The ϕ planes where the high sidelobe levels occur are $\phi = 30^\circ$ and $\phi = 90^\circ$. These ϕ planes are located halfway between two adjacent array elements, which suggests that beamsteering in ϕ planes between adjacent array elements have deteriorated performance. Conversely, the sidelobe levels in the $\phi = 0^\circ$ plane and the $\phi = 60^\circ$ plane are below -10 dB. As the number of elements increases, the sidelobe levels in the different ϕ planes decrease from 0 dB to -10 dB in every ϕ plane. This study was also conducted for different fixed radius and increasing number of elements, and was found to follow similar trends, but for brevity, were not included.

As shown in Fig. 4.2, the array factor converges to a final pattern; increasing the number of elements after a certain point does not improve the sidelobe levels. In Fig. 4.2e, the number of elements is 14 and adding two more elements marginally decreases the third sidelobe level, shown in Fig. 4.2f. Indeed, in Fig. 4.2d, the first two sidelobes located at $\theta = -5^\circ$ and at $\theta = -30^\circ$ do not decrease with the addition of more array elements and are identical to the first two sidelobes in Fig. 4.2e and in Fig. 4.2f. Furthermore, increasing the number of elements does not have significant effect on the sidelobe levels located at $\theta = 75^\circ$; in Fig. 4.2b, the sidelobe level at $\theta = 75^\circ$ is identical to the sidelobe level in Fig. 4.2c and thereafter. Lastly, it is important to highlight that the main beam at $\theta = 30^\circ$ does not vary when steering in the azimuth, and only the sidelobe levels are changing.

Additionally, the width of the main beam of a circular array does not change with the

number of elements, in contrast to a uniform rectangular array, where the width of the main beam decreases as the number of elements increases.

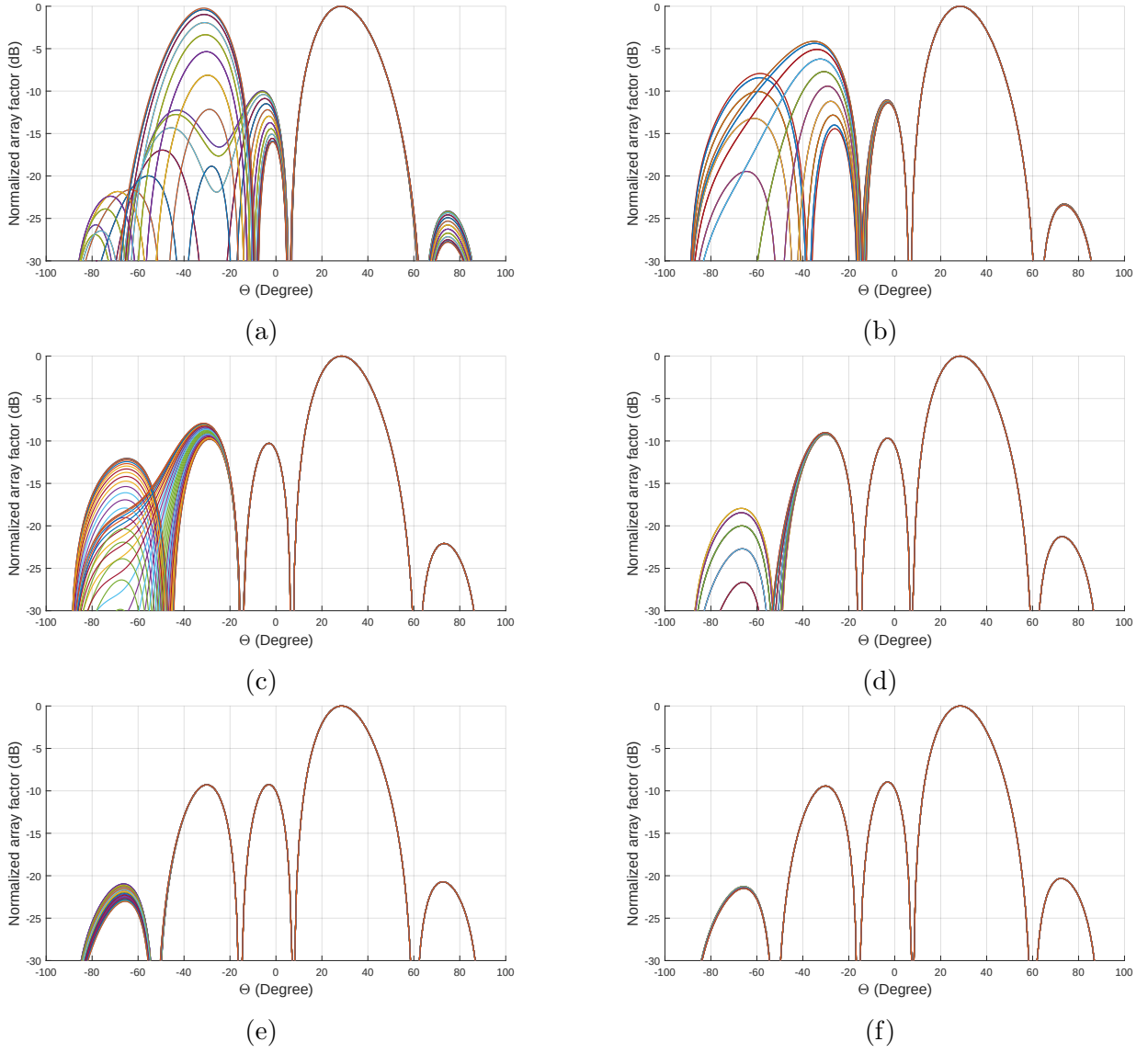


Fig. 4.2: Normalized array factor of circular array steered at $\theta = 30^\circ$ and sweeping azimuth (varying ϕ) with fixed radius λ_0 and: (a) 6 elements, (b) 8 elements, (c) 10 elements, (d) 12 elements, (e) 14 elements, (f) 16 elements.

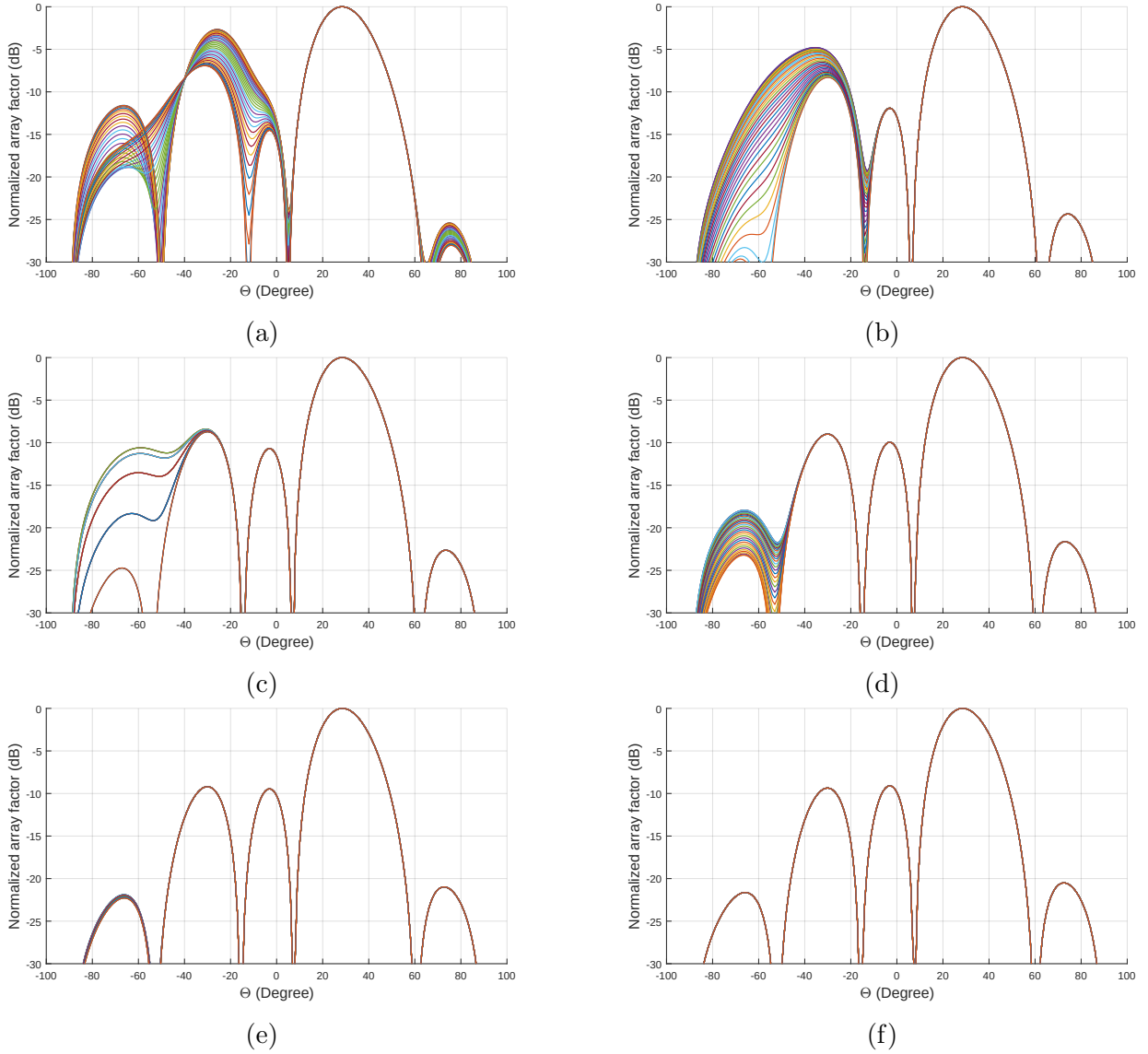


Fig. 4.3: Normalized array factor of circular array steered at $\theta = 30^\circ$ and sweeping azimuth (varying ϕ) with fixed radius λ_0 and: (a) 5 elements, (b) 7 elements, (c) 9 elements, (d) 11 elements, (e) 13 elements, (f) 15 elements.

In Fig. 4.3, the normalized array factor of a circular array with increasing odd number of elements and fixed radius of λ_0 steered to $\theta = 30^\circ$ is shown. The different traces within

each plot in Fig. 4.3 represents the array factor steered in different ϕ directions from 0° to 90° , similar to Fig. 4.2. Compared to the 6-element circular array, the sidelobe levels of the 5-element circular array and 7-element circular array are lower than the level of the main beam by at least 3 dB, shown in Fig. 4.3a and Fig. 4.3b, respectively.

A notable difference between the array factor of even number of elements and odd number of elements is the sidelobe levels of the circular array with odd number of elements is always less than the level of the main beam. Increasing the number of elements for an odd number of elements also yields the same final pattern as the case for an even number of elements, as evident in Fig. 4.3f and Fig. 4.2f.

4.2.2 Increasing Radius with Fixed Number of Array Elements

The effects of changing the radius of the circular array while keeping the number of elements fixed are shown in Fig. 4.4, Fig. 4.5, and Fig. 4.6. The various arrays are all steered to $\theta = 30^\circ$, while sweeping in the azimuth, from $\phi = 0^\circ$ to $\phi = 90^\circ$. In Fig. 4.4, the number of elements are fixed to five elements and the radius of the circular array is increased from $0.5\lambda_0$ to $1.5\lambda_0$.

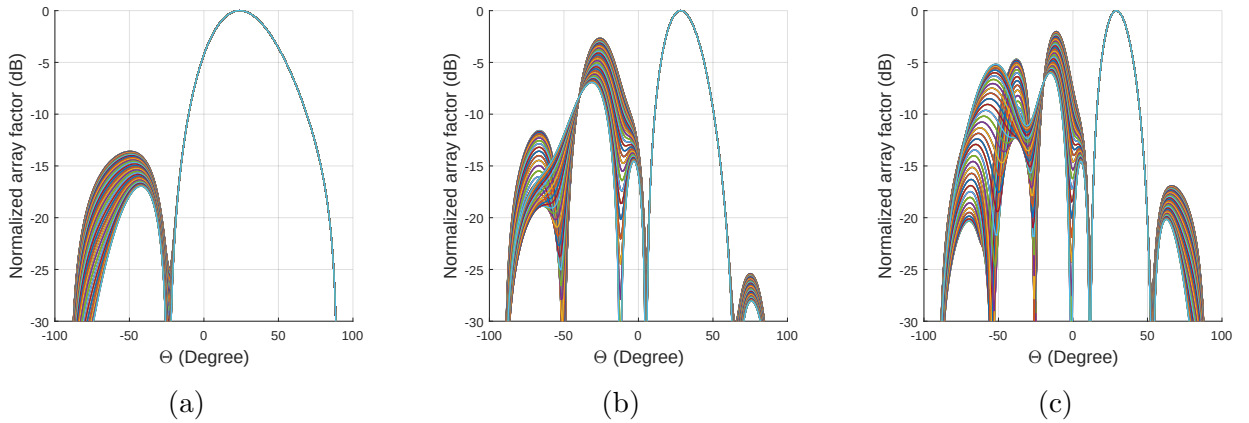


Fig. 4.4: Normalized array factor of circular array steered at $\theta = 30^\circ$ and sweeping azimuth (varying ϕ) with fixed number of 5 elements and increasing radius of (a) $0.5\lambda_0$, (b) λ_0 , and (c) $1.5\lambda_0$.

Firstly, increasing the radius of the circular while keeping the number of elements fixed increases the sidelobe levels; in Fig. 4.4a, the sidelobe levels while sweeping in the azimuth

is at -14 dB. However, when the radius of the circular array is increased to λ_0 , as in Fig. 4.4b, the sidelobe levels increase to -3 dB and in Fig. 4.4c, the sidelobe levels increase to -2 dB when the radius is increased to $1.5\lambda_0$. Secondly, as the radius increases, the width of the main beam decreases; the HPBW is 30° , 20° , and 15° for a five element circular array with radius $0.5\lambda_0$, λ_0 , and $1.5\lambda_0$, respectively.

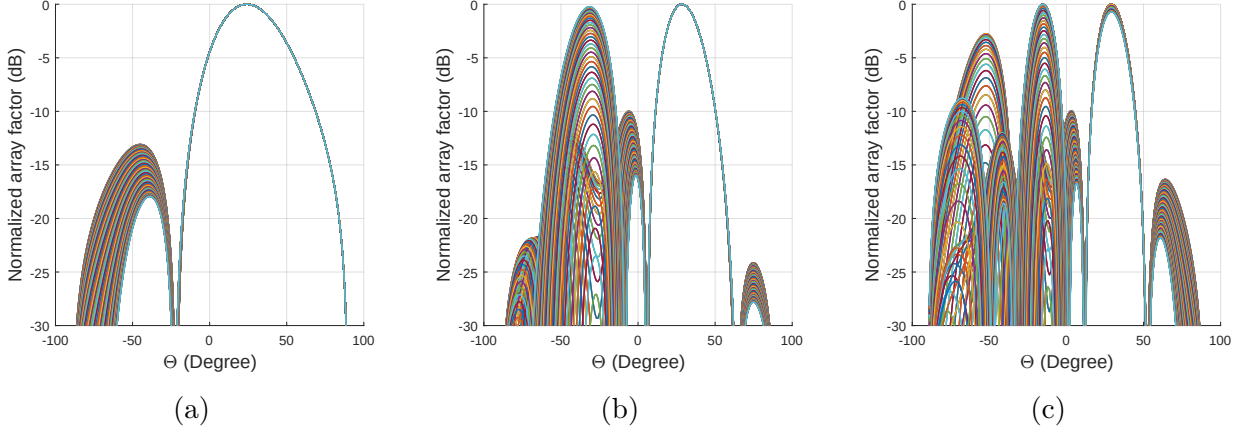


Fig. 4.5: Normalized array factor of circular array steered at $\theta = 30^\circ$ and sweeping azimuth (varying ϕ) with fixed number of 6 elements and increasing radius of (a) $0.5\lambda_0$, (b) λ_0 , and (c) $1.5\lambda_0$.

In Fig. 4.5, the array factor of a six element circular array with increasing radius from $0.5\lambda_0$ to $1.5\lambda_0$, similar to 4.4. When the radius of the six element circular array is $0.5\lambda_0$, the sidelobe levels are at -14 dB and HPBW is 30° , similar to the five element circular array with radius of $0.5\lambda_0$ in Fig. 4.4a. However, when the radius is increased to λ_0 for a six element circular array, the sidelobe levels increase significantly to the same level as the main beam, seen previously in Fig. 4.2a. Further increasing the radius of the six element circular array also increases the second and third sidelobe levels, as shown in Fig. 4.5c. Additionally, increasing the radius of the six element circular array also increases the sidelobe levels located at $\theta = 60^\circ$.

Lastly, the effects of increasing the radius of a seven element circular array is explored in Fig. 4.6. Firstly, it should be highlighted that the array factor for different ϕ scanning angles, shown in Fig. 4.6a, are identical, unlike previously in Fig. 4.4a and in Fig. 4.5a, where different array factors can be seen, depending on the different ϕ scanning angle. Increasing the radius to λ_0 , increases the sidelobe levels to -5 dB, seen in Fig. 4.6b, but

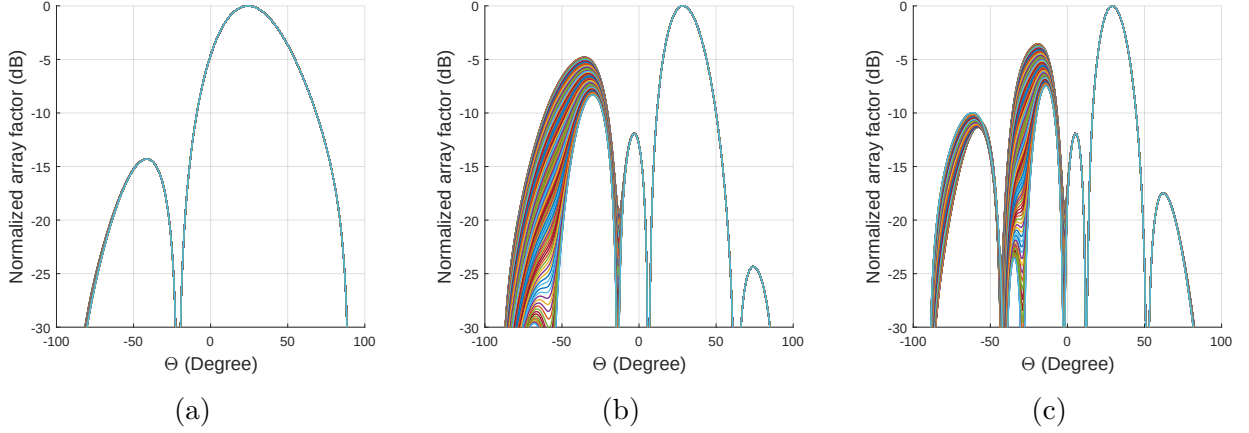


Fig. 4.6: Normalized array factor of circular array steered at $\theta = 30^\circ$ and sweeping azimuth (varying ϕ) with fixed number of 7 elements and increasing radius of (a) $0.5\lambda_0$, (b) λ_0 , and (c) $1.5\lambda_0$.

decreases the HPBW to 20° , similar to the five element and six element circular arrays. Increasing the radius further to $1.5\lambda_0$ results in higher sidelobe levels at -3 dB and smaller HPBW of 15° , shown in Fig. 4.6c.

4.2.3 Increasing Radius and Number of Array Elements

The effects of fixing the antenna array elements spacing, while increasing both the radius of the circular array and the number of elements, is explored. In Fig. 4.7, the array is steered to $\theta = 30^\circ$ and swept in the azimuth between $\phi = 0^\circ$ to $\phi = 90^\circ$. The spacing between adjacent array elements is fixed to λ_0 , while the number of elements is increased and the radius of the circular array is also increased accordingly to keep the constant inter-element spacing. In Fig. 4.7a, the number of elements is four and the radius of the circle is 0.64λ . Similar observations can be made compared to previous sections; the HPBW is high due to the small radius of the circular array and high sidelobe levels due to the low number of array elements.

Increasing the number of elements to 16 and the radius of the circular array to $2.55\lambda_0$ results in the normalized array factor shown in Fig. 4.7b. Again, increasing the radius of the circular array decreases the HPBW and increasing the number of elements in general decreases the sidelobe levels. Further increasing the number of elements to 64 and 256,

the array factors are shown in Fig. 4.7c and in Fig. 4.7d, respectively. By increasing both the number of elements and the radius of the circular array, the sidelobe levels can be decreased, shown in Fig. 4.7d. This can be especially useful at mm-wave frequencies, where λ_0 is not large and having, for example, a $40\lambda_0$ radius circular array is still relatively small.

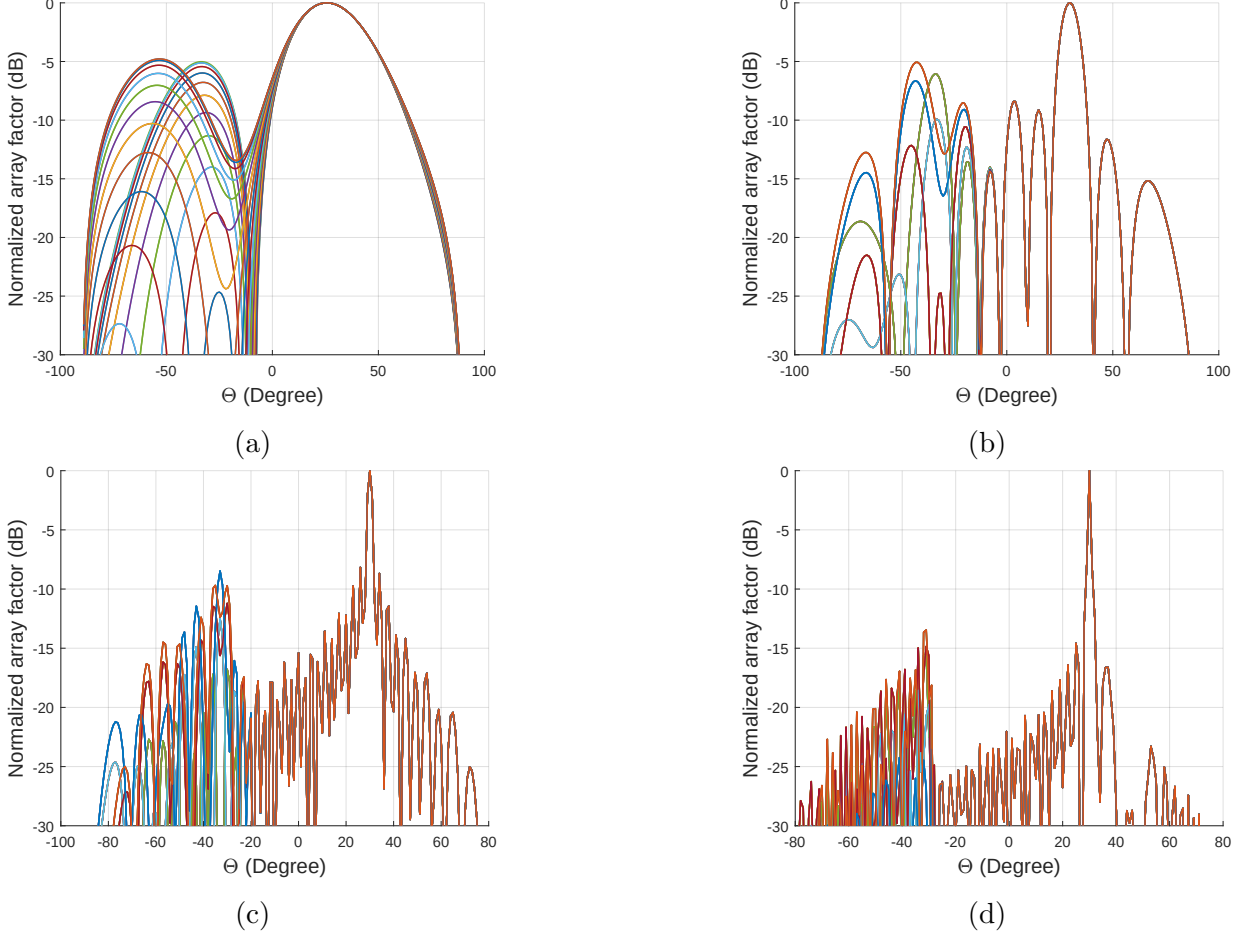


Fig. 4.7: Normalized array factor of circular array steered at $\theta = 30^\circ$ and sweeping azimuth (varying ϕ) with fixed antenna element spacing of λ_0 while increasing radius and number of elements: (a) 4 elements, (b) 16 elements, (c) 64 elements, (d) 256 elements.

Furthermore, the antenna array element spacing is λ_0 in Fig. 4.7 and is a viable array configuration for beamforming at mm-wave frequencies to overcome the $\frac{\lambda}{2}$ spacing of

uniform rectangular arrays to avoid grating lobes is a challenge to many array designs at mm-wave frequencies. Another way to reduce the sidelobe levels in, for example, Fig. 4.7d below -15 dB, the element spacing must be reduced to less than λ_0 by either increasing the number of elements if the beamwidth is to be maintained or reducing the radius of the circular array.

4.2.4 Proposed Circular Array using Novel Antenna Design

Based on the parametric study conducted in the previous chapter, an initial proof-of-concept prototype was designed and implemented on PCB, also using the same stack-up proposed in Chapter 3. Because of the size of the BFIC, the antenna element spacing was fixed to $1.1\lambda_0$, and the corresponding radius of the circular array is $1.05\lambda_0$. The number of elements within the circular array was decided to be 6+1: six elements, with an additional element in the center, to have a total of seven elements. Seven elements circular array was decided to be sufficient as an initial prototype, especially since each element is excited by a BFIC and is analogous to a 28 element single feed antenna element array. Increasing the number of elements requires a larger number of RF streams, and a custom-built solution was implemented to test and validate the 6+1 circular array, further discussed in Appendix A.

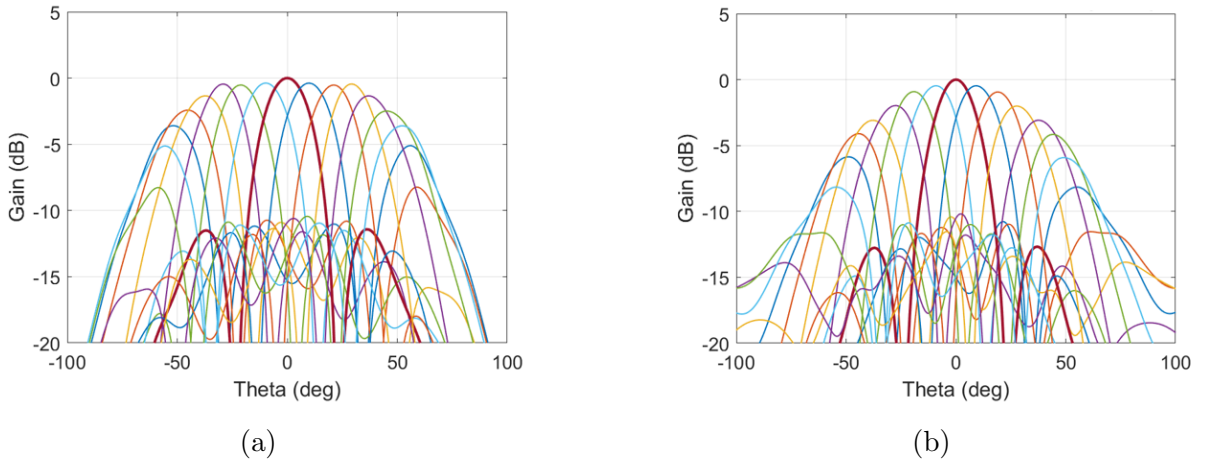


Fig. 4.8: Radiation pattern of 6 + 1 design at $\phi = 0$. (a) Horizontally polarized. (b) LHCP.

The entire circular array was simulated in HFSS and the coupling between elements was found to be -20 dB or below across the entire operating frequency range between every element. The steering performance of the circular array was also simulated in HFSS and is shown in Fig. 4.8. For brevity, only the horizontally polarized state and LHCP state is shown. The steering performance is limited; the proposed circular array can only scan $\pm 50^\circ$ before the gain drops by 3 dB when in horizontally polarized state shown in Fig. 4.8a, whereas in LHCP state, the proposed array can only scan $\pm 40^\circ$ before the gain drops by 3 dB as seen in Fig. 4.8b. The reported gain in 4.8 includes the changes in active S_{11} due to mutual coupling between elements.

The proposed 6+1 circular array utilizing the dual-differential antenna design is shown in Fig. 4.9. In Fig. 4.9b, the backside of the circular array design is shown. Six patch antennas can be seen arranged in a circular array configuration, with a seventh patch in the center. Similar as before, dummy copper was added on the edge of each layer between layer 2 to layer 6, which are all shorted to ground, in an effort to increase the copper percentage and avoid potential warping during PCB manufacturing. The effects of the added copper ground were simulated in HFSS to ensure RF performance was affected.

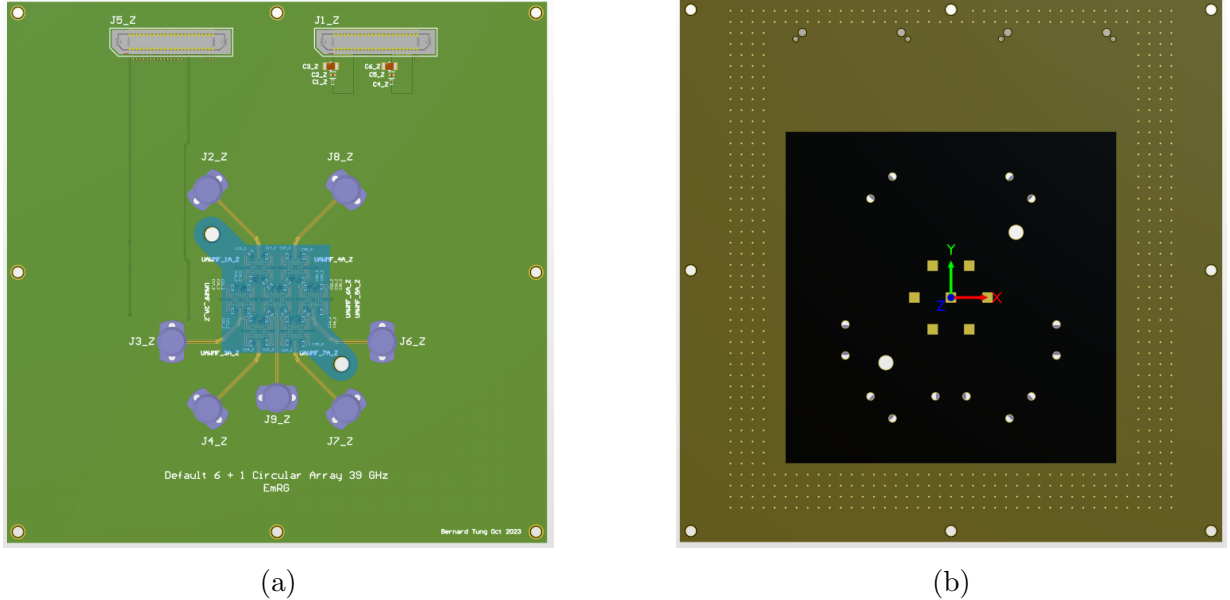


Fig. 4.9: 3D model of 6+1 circular array using dual-differential antenna elements: (a) topside of PCB, and (b) backside of PCB.

In Fig. 4.9a, the topside of the PCB design is shown. Similar to the active design proposed in Fig. 3.23, two connectors are used to route the necessary SPI signals and provide the dual power supplies to the BFIC. A blue heat sink is placed on top of the seven BFIC and fixed in place with screws. Seven 2.4mm vertical launch RF connectors are located on the topside of the PCB to route directly to each BFIC. The RF connectors were routed far away enough to provide clearance for the heat sink and enough distance between adjacent RF connectors to allow for cable connections.

Chapter 5

Conclusion and Future Works

5.1 Conclusion

This thesis proposed a fully-digital beamforming array operating at mm-wave frequency using a novel dual-differential antenna design capable of polarization configurability. The beamforming array utilizes a circular configuration to overcome the stringent antenna spacing requirements at mm-wave frequencies. Overall, the array is able to incorporate seven RF data streams and operate in two linear polarization states and two circular polarization states.

In Chapter 2, different antenna metrics were presented, and desirable antenna features were discussed. Afterward, a literature review on antenna designs with multi-polarization capabilities and digital beamforming solutions at mm-wave frequency was presented. The literature review presented the need for antenna designs operating at mm-wave frequencies that are polarization reconfigurable while exhibiting stable radiation patterns. Additionally, most beamforming architectures at mm-wave frequencies utilize RF beamforming, however, digital beamforming arrays solutions are not as commonplace. The literature review presented the need for fully digital beamforming arrays at mm-wave frequencies.

In Chapter 3, a dual-differential antenna design was proposed. Both passive and active versions of the design were presented and analyzed. Both designs were shown to have good return loss within the range of operating frequency and across all polarization states. Furthermore, the radiation patterns of the antenna designs did not vary significantly between

different polarization states.

In Chapter 4, the array factor of a circular array was presented and a short parametric study on circular arrays was conducted. Different parameters of a circular array were varied and the effects on the array factor were investigated. By changing parameters such as the number of elements and the radius of the circular array, sidelobe levels and beam width could be changed. Lastly, a six element circular array with a center element, or 6+1 array, was presented. The array elements used the same dual-differential antenna design presented in Chapter 3.

5.2 Future Works

- **Multi-ring circular array:** In Chapter 4, the viability of circular arrays was explored and proposed. However, unlike uniform rectangular arrays, scaling circular arrays is not as straightforward and requires further investigation in optimizing and implementing circular arrays with larger number of elements to improve the scanning angle of the array, as well as reducing the sidelobe levels. One way to increase the number of array elements within a circular array is to implement multiple concentric rings of array elements with different radii. However, further investigation into the viability and design methodology of multi-ring circular arrays is required.
- **Fully connected hybrid beamforming:** In Chapter 4, a fully-digital beamforming array was proposed, which incorporates seven RF data streams. However, an alternative solution to achieve multi-stream multibeam arrays is to use fully connected beamforming ICs. By using commercially available fully connected multibeam beamforming ICs, a hybrid beamforming architecture can be realized. Implementing a hybrid beamforming array while using the proposed dual-differential antenna design can offer a compromise between cost and performance; an array that is polarization reconfigurable while needing a reduced number of RF data streams.
- **Dual-polarized reconfigurable antenna design:** In Chapter 3, a polarization reconfigurable active antenna design was presented. The presented design is able to achieve four different polarization states, however, the presented design is still single polarized at any given time. A future design can provide dual-polarization while also

achieving polarization reconfigurability by using a stacked-patch design. However, the isolation between the two patches should be taken into account

- **Dual-band reconfigurable antenna design:**

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APPENDICES

Appendix A

Dual-Channel Upconverter Board

A custom solution was required to provide the fully-digital beamforming array with the necessary number of RF input data streams. Analog Devices ADMV1013 wideband microwave upconverter was selected due to multiple reasons: the output RF frequency is wideband and completely covers the frequency band of 5G FR2 n260, utilizes 4-wire SPI protocol which allows for daisy-chain, and provides up to 23 dB of conversion gain necessary to provide sufficient power levels at the input of the RF front-end of the antenna array.

A.1 System Requirements

The dual-channel upconverter board systems requirements are mainly dictated by the power and frequency requirements of the selected upconverter IC, ADMV1013. The ADMV1013 IC is a wideband, microwave upconverter from Analog Devices and uses a 40-terminal land grid array packaging. The RF output of the ADMV1013 covers the frequency range 37–40 GHz, which is the frequency range the 6+1 circular array operates at. In Table A.1, the frequency range of the LO signal, IF signal and RF signal.

An additional requirement of the dual-channel upconverter board is the ability to scale in number of channels; manufacturing multiple copies of the dual-channel upconverter board must work together, without requiring additional components or resources. For example, if eight total channels are required, then four copies of the dual-channel upconverter

Parameter description	Min	Max
RF Output	24 GHz	44 GHz
LO Input	5.4 GHz	10.25 GHz
LO Quadrupler	21.6 GHz	41 GHz
IF Input	0.8 GHz	6 GHz

Table A.1: Frequency requirements of ADMV1013

board is required. However, with four copies, four LO sources are required and four sets of SPI control is required. To circumvent this issue, the dual-channel board was designed such that multiple copies could be daisy-chained together. The LO signal was daisy-chained, as well as the SPI 4-wire signals, consisting of clock (CLK), master out slave in (MOSI), master in slave out (MISO), and chip select (CS).

The dual channel board requires only one LO input; the LO signal is split using two cascaded Wilkinson power dividers (PD-0530SMG). The first Wilkinson power divider splits the LO signal in two: one is routed to the next Wilkinson power divider, whereas the other output is routed off-board and can be daisy-chained to the LO input of the next board. The output of the second Wilkinson power divider is connected to the LO input of each ADMV1013. Careful consideration was taken into account to ensure the power levels at the LO input and the power levels at the LO output were the same. The reason for this is to allow multiple dual-channel upconverter boards to be daisy-chained without requiring external amplifiers to boost the LO signal, as the Wilkinson power divider will attenuate the signal by 3 dB for each daisy-chained board. A fixed -6 dB attenuator and wideband microwave amplifier with 12.3 dB gain were used in conjunction to overcome this issue. Variable gain attenuators and variable attenuators were considered; however, the aforementioned components that were able to meet the frequency and power requirements are not common and the few that met the requirements required differing supply voltages and separate SPI signals, which would add additional design complexity and cost to the board.

Similarly, the SPI 4-wire signals are also daisy-chained. Within each dual-channel upconverter board contains two ADMV1013 ICs, which are daisy-chained together, whereas between each dual-channel upconverter boards, the SPI signals are also daisy-chained together. Further investigation is still needed to determine how many dual-channel upcon-

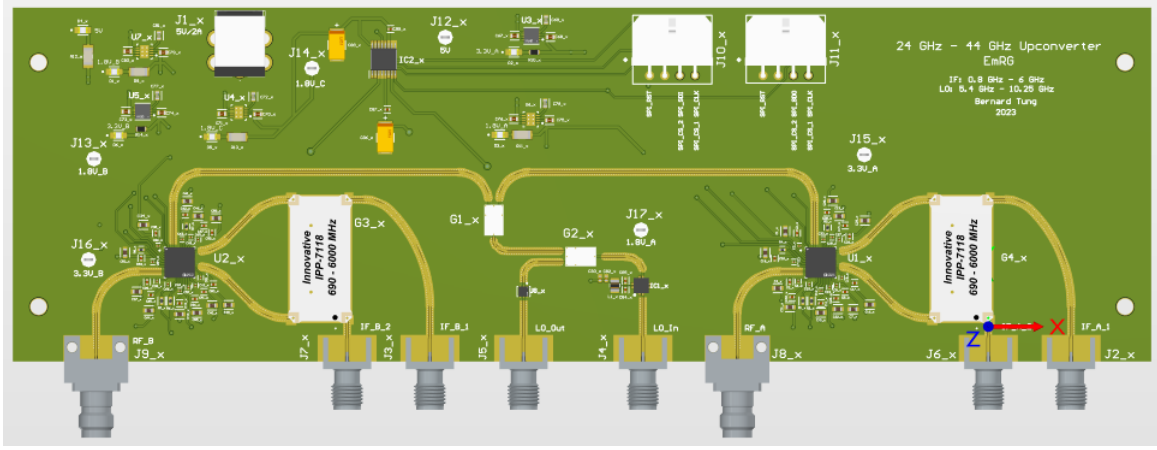


Fig. A.1: 3D drawing of upconverter board design

verter boards can be daisy-chained together; too many daisy-chained components can be limited by the SPI CLK skewing and result in inability to program the ADMV1013 IC correctly.

The 3D drawing of the dual-channel upconverter board is shown in Fig. A.1. On one edge of the board, exists all the RF connectors, LO connectors, and IF connectors. This layout was chosen because of the pin layout of the ADMV1013 and to keep the RF outputs on the same side of the board; to connect the LO signal to both ADMV1013 without using another layer for routing, this layout was considered the simplest and most symmetric in terms of routing as it did not require via transitions or stripline transmission lines, which would require additional PCB stack-up complexity and cost.

The dual-channel upconverter board has a single 5 V supply input, which is distributed to multiple voltage regulators to be converted down to 3.3V and 1.8V, to provide the necessary voltage supply for the ADMV1013 IC. Linear regulators were used instead of switching regulators to reduce the likelihood of EMI noise. Lastly, a DAC was required to control the RF conversion gain of the ADMV1013. The requirements of the DAC are: high-precision, output voltage range of 0-1.8 V, voltage supply less than or equal to 5 V, and 1.8 V logic SPI digital output signal for daisy-chaining capabilities. The AD5687R is a 12-bit high-precision DAC from Analog Devices that meets all the above requirements. Furthermore, the AD5687R DAC is dual-channel and can control two ADMV1013 concurrently, instead of requiring two single-channel DACs to control the two ADMV1013.