Passive Amplitude and Hybrid Parasitic Beam Steerable Array Antennas

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To Jesus

In Him I live, in Him I breathe and in Him I have my being.
Abstract

Highly directive antennas with beam steering capabilities are needed to overcome high path-loss, mitigate against interference from unwanted signals and to ensure network connectivity for the next generation 5G millimetre wave communication systems. Phased arrays using phase shifters are the conventional approach of achieving beam steerable antennas. However, due to the insertion loss incurred by phase shifters and the high cost of acquiring phase shifters, alternative solutions that offer lower loss and cost are being researched. This work proposes two beam steerable antenna solutions that offer reduced insertion loss and cost. The first beam steering solution is a novel passive amplitude steering antenna that makes use of a reconfigurable ratio power divider instead of the conventional variable gain amplifiers for amplitude control. The designed antenna is a $2 \times 1$ antenna array which achieves continuous steering over the range $0^\circ$ to $21^\circ$ in one direction. The second beam steering solution combines the parasitic steering technique with the phased array approach to form a hybrid parasitic antenna. The hybrid parasitic antenna was formed from a $3 \times 1$ phased antenna array surrounded by eight parasitic elements. The parasitic elements are operated by switches and the antenna achieved a fine steering resolution of $1^\circ$. Computer simulations were performed to assess the effect of using real switches and phase shifters within the antenna. To this effect, 3-bit phase shifters were incorporated into the hybrid parasitic antenna and its performance compared with that of a conventional $3 \times 1$ phased array incorporating 5-bits phase shifters. The realised gain of the hybrid parasitic antenna was 4 dB better than that of the conventional phased array. This suggests that the novel antenna offers significant improvement in antenna performance. This work also proposed a novel approach that improves the isolation bandwidth of a power divider which will be vital for phased arrays and MIMO applications. The compensated power divider achieves a 15 dB isolation bandwidth of 114% which is more than 40% of the isolation bandwidth of conventional Wilkinson power divider. This power divider solution is suitable for ultra-wideband frequency applications.

Key words: Power Dividers, Amplitude Steering, Beam steering, Parasitic steering, Reconfigurable ratio power divider.

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Chapter 1

Introduction

Wired communication systems suffer from the limitation of mobility and high cost of installation. Hence, Radio Frequency (RF) and microwave technology play a major role in modern communication systems. RF and microwave technology are applicable in diverse aspects of wireless communication systems, wireless security systems, radar systems, environmental remote sensing and medical systems [1]. With growing interests in 5G millimetre wave technology, the need for beam steerable antennas to mitigate high path loss and interference from unwanted sources are in high demand. Phased array is the conventional means of achieving beam steering. It is simply the arrangement of multiple similar antenna elements separated by 0.5λ between successive elements. The radiation pattern of a phased array antenna is obtained as the product of the single element factor and the array factor. To steer the antenna’s radiation pattern using phased arrays, the phases of the current supplied to each element is altered by specific angles with the use of phase shifters. The beam from each array element would then add up constructively in phase to form a beam steered towards the desired direction. The phased array approach is advantageous because it offers an electronic means of achieving beam steering without the rotation of the physical antenna.

To achieve fine beam steering resolutions in phased arrays, the number of driven elements in the array and the number of bits of the phase shifter used in the phased array will be large. The large number of bits of the phase shifter will enable smaller phase difference between successive driven elements. The increase of the number of
driven elements and bits of phase shifter incurs high cost to the antenna system and a high insertion loss in the phase shifter which will be detrimental when operating in the millimetre wave frequencies. Efficient beam reconfigurable antennas with low insertion loss and fine steering resolution have received a lot of attention recently for applications such as wireless backhaul for cellular networks, millimetre wave identification, anti-collision and handheld radar.

1.1 Motivation

Beam steering is vital to enhance the signal quality between a transmit and receive antenna by reconfiguring the main lobe of the radiation pattern to improve the gain in a desired direction. It is also needed to reduce interference from unwanted signals that are in close range by steering the main lobe away from the unwanted signal. Beam steerable antennas are needed in areas such as:

1. Millimetre wave backhaul and access for cellular networks: Due to the challenge of high path-loss at millimetre wave frequencies (30-300 GHz), highly directive antennas are essential. Where mobility is required these antennas must be capable of beam steering. Beam steering may also help to mitigate for signal blockage due to obstructions by redirecting the radiation pattern of the antenna towards the strongest signal or (in the case of backhaul) to other available nodes in the network.

2. WLAN/WPAN/WiGig high data rate point-to-point links for IEEE 802.11ad: Applications such as wireless HD streaming would require beam steerable antennas for realigning the beam direction in mobile scenarios.

3. Automotive parking and anti-collision radars for blind spot detection, and collision warning.

The beam steering approach that would be suitable for these applications should be low loss, to reduce the signal degradation, and should achieve fine steering angles for mobility scenarios to maintain quality of the signal link. Conventional beam steering
techniques suffer from poor steering resolution, high cost and complexity, and high insertion losses which degrades the antenna performance.

This work seeks to improve the steering resolution for beam steerable antennas while keeping the cost, complexity and insertion loss at a minimum. Novel approaches to beam steering will be proposed and comparisons carried out to highlight the improvements made by the proposed solutions. The target performance of the antenna would be to achieve a gain of 7 dBi, a steering range of ±45° and a fine steering resolution between 1 – 2°.

1.2 Scope

Beam reconfigurable antennas will play a major role in achieving 5G millimetre wave communication system for wireless backhaul for cellular networks and in applications such as millimetre wave identification, anti-collision and hand-held radars. They would be needed to mitigate high path loss and avoid interference from moving objects, which are prevalent in the millimetre wave frequencies, to deliver high data rate and mobility in next generation communication systems. Hence, this thesis focuses on improving the steering resolution of beam steerable antennas to ensure an optimum communication system.

1.3 Contributions

The following are original contributions to knowledge that this thesis presents:

1. A design and analysis of an ultra-wide isolation bandwidth compensated power divider (CPD). The designed CPD achieves a 15 dB isolation bandwidth greater than 100% which is more than 40% greater than the conventional Wilkinson Power Divider.

2. A proof-of-concept reconfigurable ratio power divider (RRPD) with stable resonant frequency. The RRPD is operated with a potentiometer acting as an
isolating resistor. An equation relating the value of the potentiometer to the reconfigurable ratio was derived empirically.

3. A novel proof-of-concept amplitude steerable antenna based on a reconfigurable ratio power divider (RRPD). The antenna achieves a continuous steering range of 21° off boresight as the value of the potentiometer changes.

4. A hybrid parasitic array antenna that combines phased array and parasitic steering techniques. The antenna achieved fine steering resolution of 1° when parasitic steering was used. With a 3-bit phase shifter, the antenna performance can be compared with a conventional phased arraying using 5-bit phase.

1.4 Publications

The following are journal and conference publications associated with this work.


The following journal articles have been submitted and are under review.


1.5 Structure of Thesis

The rest of this thesis is organised as follows:

1. Chapter 2: A detailed literature review of beam steering techniques. A comparison of the various techniques based on the state figures of merit will be presented. Part of this chapter is published in [2].

2. Chapter 3: This chapter presents a novel approach/method of enhancing the isolation bandwidth of equal power dividers. A review of techniques used in enhancing isolation bandwidths is presented along with design and measurement results for the compensated power divider (CPD). The CPD will be analysed to show the trade-off between the wide isolation bandwidth and the return loss at the operating frequency. Measurement results for a prototype of the CPD designed to operate at 6 GHz will be presented and compared with other techniques in literature.

3. Chapter 4: This chapter presents a novel passive amplitude steerable antenna based on a reconfigurable ratio power divider (RRPD). An overview of the techniques used to achieve reconfiguration in power dividers will be presented. A novel RRPD designed employing the use of a potentiometer to control the dividing ratios will be presented and analysed. Measurement and simulation results will be presented to validate the design. The RRPD is then integrated into a $2 \times 1$ antenna array for a proof of concept passive amplitude steerable antenna. The designed antenna is analysed to show the changes in surface current density as the ratio to the driven elements are altered. Measurement results for the fabricated antenna will be compared with results obtained from simulation. The effect of inter-element spacing on the steering range will be studied. The effect of increasing the number of driven elements on the beam steering range was then investigated. In this way, it was possible to evaluate the limitations of the concept.

4. Chapter 5: This chapter presents a novel hybrid parasitic array antenna design that combines parasitic and phased array steering techniques. It shows the results
from simulation for the proof-of-concept design at 11 GHz. The antenna was analysed to find the minimum steering angle that can be achieved using parasitic steering technique alone. The chapter also compares the resolution of the designed antenna with that of a conventional phased array.

5. Chapter 6: This chapter presents the fabricated hybrid parasitic array antenna, discussed in chapter 5. Considerations on the type of feed network and phase shifter used in implementing the design are presented. Graphs describing the radiation and return loss performance of the antenna are shown. To examine the actual power losses in the hybrid parasitic array antenna, s2p files of real phase shifters and switches were used in simulation and the gain of the parasitic antenna is compared to that of a conventional phased array. A table, comparing the hybrid parasitic array antenna with other state-of-the-art beam steerable antenna designs is presented.

6. Chapter 7: This chapter concludes the thesis and proposed recommendations for future work.
Chapter 2

State-of-the-Art in Beam Steering Techniques

Antennas with beam steering capability are proposed to be the solution to the challenge of high path loss and signal attenuation faced at millimetre wave frequencies. Beam steerable antennas also help to improve the signal to interference ratio (SINR) in antenna system by redirecting the main lobe of the antenna away from sources of interference. Hence, beam steerable antennas are vital for current and future communication systems for applications such as anti-collision and radar systems, wireless backhaul point-to-point links for cellular communications, 5G millimetre wave communication systems for access backhaul, and WLAN/WPAN high data rate links.

Phased arrays have been the conventional approach of achieving electronic beam steering without the need of physically rotating the antenna. Beam steering in phased arrays is achieved by altering the phase difference in the current supplied to successive array elements. Fig. 2.1 shows an illustration of phase array steering for an eight element phased array. With no phase shift applied to the array elements, the radiation pattern of the phased array is boresight above the phased array. To achieve beam steering, a phase shift of $\beta$ is applied between successive array elements to achieve the radiation pattern $\phi_{\text{max}}$ away from boresight. The equation that estimates the maximum steered
angle based on the phase shift between consecutive elements is given as [3];

$$\phi_{\text{max}} = \sin^{-1} \left[ \frac{\lambda \beta}{2\pi d} \right]$$  \hspace{1cm} (2.1)

where $\lambda$ is the wavelength, $\beta$ is the phase difference between successive elements, $d$ is the distance between the elements, normally $0.5\lambda$, and $\phi_{\text{max}}$ is the steered angle.

Beam steerable antennas have been in use since the 1930s [4]. Research in beam steerable antennas is generating lots of interest as efforts are being made to develop an optimum beam steering solution that incurs low loss and low complexity for both point-to-point and point-to-multipoint applications. For point-to-multipoint applications, the capability of forming highly directive multiple beams is highly desired and currently being investigate. This chapter presents a detailed survey of the various techniques and approaches used in realising beam steerable antennas.

This chapter is structured as follows: Section 2.1 will present the figures of merit used in analysing various beam steering techniques. This will provide a platform to enable easy comparison of the various techniques. Section 2.2 presents the various beam steering techniques in details. Section 2.3 presents a critique of each technique with the aim of identifying a gap in literature. A summary of the chapter will be presented in Section 2.4.
2.1 Figures of Merit

To analyse the performance and effects of the various beam steering technique, it is necessary to define appropriate figures of merit. These figures of merit for beam steerable antennas are:

1. Insertion loss (IL): This is the loss incurred when the beam steering technique is incorporated in the antenna system. It is mathematically the difference between the power that goes in and the power that comes out of the device expressed in dB. This is an important figure of merit as the higher insertion loss incurred by the system, the worse the efficiency of the system.

2. Steering range: This is the maximum angle away from the bore sight or end-fire direction to which the beam can be steered. The steering range will depend on the type of radiating element used in the design. Consider, for example, a phased array antenna made up of microstrip patch antennas cannot be steered over 360° in the E-plane or H-plane due to the limitations imposed by the directionality of the element pattern. While a monopole can be steered over 360° in the H-plane. Hence, in the discussion that follows, the steering range of techniques will be classified as either full range or a constrained to a specific angular range.

3. Steering Resolution: This is the granularity in angle that can be achieved while covering the steering range. It could be continuous (i.e. achieving any steering angle), predefined (i.e. fixed large angle steps ranging from 15-20°) or fine (i.e. stepwise 2° or less).

4. Steering speed: This refers to the time taken for the beam of the antenna to be steered to a specific angle. It will be specified as fast (for instantaneous steering) or slow (for time delayed steering). This will determine the suitability of the technique either for rapid or slow changing environments. The steering speed for
a mobile user driving along the road will be different from that of a backhaul antenna that needs to switch its beam to a different node in the network.

5. Complexity: This focuses on the ease of implementing a technique. This will focus on the type of components used in achieving the technique such as switches, varactor diodes, and the ease of operating them to achieve beam steering. It has a direct effect on the cost of implementing the technique. It will be rated as either high, moderate or low.

6. Size: The size of the manufactured device will determine the application where it will be used. If it becomes very bulky, applications such as smartphones and tablets will not be able to accommodate it.

7. Cost: This will refer to how expensive it will be to implement the beam steering technique. On a sales perspective, the cost of implementing a technique will influence the cost of the device.

### 2.2 Beam Steering Techniques

The following techniques have been used to steer the radiation pattern of an antenna:

1. Mechanical steering
2. Beamforming
3. Reflectarray
4. Parasitic steering
5. Integrated lens antennas (ILAs)
6. Switched beam antennas
7. Travelling wave antennas
8. Retrodirective antennas
9. Metamaterial antennas
2.2. Beam Steering Techniques

10. Amplitude Steering

2.2.1 Mechanical Steering

This involves manually turning the antenna to focus the radiation pattern to the direction of interest. Mechanical steering becomes undesirable and difficult when the antenna is large, and heavy, and the external weather conditions are unfavourable. Mechanical steering can be implemented by means of MEMS actuators or electric motors. Recently, MEMS actuators have been used to implement mechanical steering [5] in a vee antenna system. The arms of the vee antenna were linked to the actuators by a pull/push bar. Operating the actuators steered the beam. This approach offered improved speed of scanning compared to manually steered arrays as well as low losses to the system. Mechanical steering is highly effective since it maintains the gain of the antenna and offers flexibility in the steering range of the antenna [6].

The authors in [7] designed a rotating system that enabled 360° steering range with 4° steering resolution. In the design, a parabolic reflector was fabricated on a rotating circular waveguide and a horn antenna was used to illuminate the parabolic reflector. The structure had a reduced moment of inertia due to its light weight which gave it better agility when compared to conventional mechanical rotating systems for antennas. This improved the speed of steering but it is not as fast as electronic methods of steering. Its use is limited to very slow changing environments due to the limitation in steering speed. Also, rotating mechanisms are prone to mechanical failure due to fatigue and wearing of moving parts [8]. The solutions for these problems led to electronic ways of steering beams.

2.2.2 Beamforming

The term beamforming refers to the process of combining signals from an array of elements to form a highly directional beam of radiation. The term is also used to refer to the process of precisely aligning the phases of an incoming signal from different elements of an array to form a well-defined beam in a specific direction. This is achieved by implementing a time delay on each elements signal [9]. It originated from spatial filters
that were designed to form pencil beams (i.e., highly directional radiation patterns) to receive signals from a specific location and attenuate interference from other locations [10]. Beamforming techniques can be sub classified as: RF/analogue, digital, or hybrid beamforming.

**RF/Analogue Beamforming**

Fig. 2.2 shows the architecture of an RF/analogue beamforming system. From Fig. 2.2, the signal from an element of the phased array ($A_1$ or $A_2$ or $A_N$) is fed through a low noise amplifier after which the time delay is implemented by means of a phase shifter. The time delayed signals from each element are summed to produce the resultant beam by the beamformer. This method of beamforming is relatively cheap and low power when compared to digital beamforming [11]. However, it poses a challenge for applications having a large bandwidth requirement because the phase shifters are frequency dependent and produces variable phase shift across the bandwidth with accurate phase shift at the centre frequency [9]. This will be a limitation when used at higher frequencies where large bandwidth is fundamental. The combination of phase shifters connected to array elements is called a phased array which is the means used to steer the antenna beam.

![Figure 2.2: Layout of RF/Analogue Beamforming](reproduced courtesy of the Electromagnetic Academy)
2.2. Beam Steering Techniques

A phased array is the conventional way of steering the beam of radiation from an antenna electronically towards different directions within the angular range illuminated by the radiating element. Phased arrays are attractive because they enable versatile beam steering. A diverse range of phase shifters are commercially available and they are easy to integrate into the array. The combined beam of the array can be steered by changing the phases of the signals fed to each element in the array using phase shifters. RF/analogue beamforming has the advantage of high directivity, multiple beamforming (one at a time in different directions), fast scanning when compared to mechanical steering due to its electronic circuitry, and spatial filtering [12].

According to [13], a phase shifter is a control device that has a flat group delay versus frequency within its defined bandwidth of operation. Phase shifters could be digital or analogue. Most analogue phase shifters are built with varactor diodes and offer continuous phase shift by controlling the voltage applied to the varactor diodes. This control voltage is highly influenced by noise and the noise adds significant losses to the phase shifter. For this reason, digital phase shifters have received much attention due to their immunity to the noise present on voltage control lines. There are several techniques used to implement digital phase shifters, some of which are switched-line, loaded-line, reflection-type, and vector modulator technique phase shifters [13],[14].

In the switched-line technique, the input RF signal is routed through an appropriate length of matched transmission line. A PIN diode or switching MESFET is used to switch between two or more different lengths of transmission line [13]. The length of each transmission line is calculated in such a way to achieve the desired amount of phase shift as the signal travels through the line [15]. Recent research into the switched-line technique centre on reducing the insertion losses and lengths of the transmission lines that implement phase shifts and semiconductor devices used for the switching. The following types of switch are employed within switched line phase shifters: MEMS [14], CMOS transistors [16], PIN diodes and FETs. Research into these devices is on-going and seeks to reduce the insertion losses along with the level of phase deviation across the operating frequency band. In [17], some improvement in the performance of the phase shifter were obtained by employing a metamaterial transmission line which has negative permittivity and permeability. This enabled the phase shifter to have a uni-
form phase response across the operating bandwidth. The design had an insertion loss of less than 2 dB at 2 GHz. In [18], phase shifters were developed using integrally-gated graphene transmission lines. The authors used the graphene transmission lines to create: switched-line, loaded-line, and reflective type phase shifters. In this context, integrally-gated means that double gated electrodes were integrated for ease of connecting switches, tuneable loads and key components needed to build phase shifters. The authors stated that graphene transmission lines are low loss in the Terahertz frequency band. However, the insertion losses were not measured and so it is impossible to ascertain how much loss these phase shifters will introduce into a system.

In the loaded-line technique, the transmission line is loaded with a capacitive reactance as shown in Fig. 2.3. This reactance is connected in a shunt configuration. The reflection and transmission coefficients depend on the value of this impedance. Varying the impedance alters the reflection and transmission coefficients which determines the time delay, hence adjusting the load alters the phase shift experienced by the signal [13],[1]. The technique is used to achieve phase shifts of 45° or less and has the advantage of being compact [13]. The loaded line technique has been used at X-band frequencies and showed a maximum insertion loss of 1 dB and a phase error of ±2° at C-band frequencies (5-6 GHz) [13]. This phase shifting technique is used in conjunction with others to improve the range and resolution of phase angle adjustment that can be achieved.

![Figure 2.3](image)

Figure 2.3: Schematic of loaded-line phase shifter [reproduced courtesy of the Electromagnetic Academy]

A reflection-type phase shifter typically consists of two impedance elements ($Z_a$ and $Z_b$), a 3-dB hybrid coupler and switches, as shown in Fig. 2.4. The impedance elements are connected across two ports of the 3-dB hybrid coupler by means of a single-pole,
2.2. Beam Steering Techniques

A single-throw (SPST) switch.

![Hybrid coupler diagram](image)

**Figure 2.4:** Illustration of reflection-type phase shifter [reproduced courtesy of the Electromagnetic Academy]

The hybrid coupler splits the input signal into two separate signals with equal amplitudes but a 90° phase difference. The impedances reflect the divided input signal while also introducing a phase shift. The value of the phase shift depends on the choice of reactance used. If the impedances are matched, the reflected waves from both arms cancel out. Varying the impedance alters the level of reflection and hence the phase shift in the output signal [19]. The hybrid coupler combines the reflected waves constructively to produce an output [20]. Research into reflective-type phase shifters is currently concerned with improving the load to achieve low phase variation during phase shifting and reduce insertion losses.

In the vector modulator technique, the input signal is fed into a quadrature power splitter which produces four output signals. The first output signal can be regarded as the reference signal with a phase of 0°. The remaining three signals differ in phase from the reference by 90°, 180°, and 270°. Pairs of signals (0° and 180° as well as 90° and 270°) are fed through a differential amplifier [21]. The amplitude of each of the signal is weighted by variable attenuators and then the weighted signals are combined by an input modulator to obtain a phase-dependent vector sum [22]. In [21], a gain and phase variation of 0.1 dB and ±0.8°, respectively was achieved. [23] reported a gain and phase variation of 0.5 dB and ±7° while achieving a 3-bit digital phase shift with a gain of 11.5 dB.
Digital Beamforming

Fig. 2.5 shows the architectural layout used in digital beamforming. The signals from each antenna element are sampled by an A/D. The sampled signals are then down converted to a lower frequency by mixing the output of the A/D with a complex sinusoid signal to yield a baseband signal. The baseband signals are then summed up by a channeliser and fed into a beamformer for processing. For the transmit mode, the beamformer applies steering and correction coefficients to the baseband signal and passes the signal to the channeliser. The channeliser splits the signal into different channels which is then upconverted to a higher frequency. The A/D is capable of converting the high frequency signal from digital to analogue before delivering to each antenna element [9]. The process is shown in Fig. 2.5. The beamformer is a signal processor used with an array of sensors/elements to provide spatial filtering.

Digital beamforming can be performed over wide bandwidths due to its ability to split the signals into various channels. Two main approaches for wideband beamforming are based on time-domain processing and frequency domain processing [24]. Time domain processing is performed using tapped delay line filters, the length of the lines depends on the bandwidth of signals. Frequency-domain processing is performed by using a
Fast Fourier transform to convert the wideband signal into frequency domain. Each converted signal is then processed by a narrowband processor [25]. This approach of narrowband processing can produce beamforming without bandwidth phase deviation which enables a beam to steer uniformly over its operating bandwidth. However, it has high power requirement and large cost.

**Hybrid Beamforming**

![Hybrid Beamforming Architecture](Image)

Recently, interest is growing in a hybrid of analogue and digital beamforming that is intended to reduce the complexity of digital beamforming and improve the performance of analogue beamforming. Based on Fig. 2.6, hybrid beamforming simply involves attaching digital beamforming architecture to the end of the RF/analogue beamformer. The RF/analogue beamforming section controls the phase of the signal at each element while the digital beamforming section applies baseband signal processing to enhance the performance by applying the appropriate weights and phases to the baseband signals [26]. Recent works in RF beamforming have concentrated on reducing the losses associated with phase shifters. Currently, the major research interest in digital beamforming
2.2. Beam Steering Techniques

concentrates on the development of algorithms which yield improved the computational time and signal tracking capabilities.

2.2.3 Reflectarray Antenna

A reflectarray is formed from the combination of a reflector and an array antenna. The aim, in combining the two technologies, is to utilise the strengths of each to maximum advantage. In a conventional parabolic reflector antenna, a field source is placed at the focus of a reflector, as shown in Fig. 2.7. The field generated is then directed towards the point of interest by the reflector. In reflectarray antennas, the reflector is replaced with numerous reflector elements. Each element is separated by an identical spacing. The array is designed to redirect the incident waves. This is achieved by applying predefined phases to different sections of the array [27]. The predefined phases are either set actively, by using phase shifters or passively, by the shape and size of each element of the array. The phase shifters are connected to all the elements of the array [28].

![Figure 2.7: Illustration of reflectarray steering technique](Image)
When active elements are employed within the reflectarray it can dynamically steer its radiation pattern. In [29], the authors achieved continuous beam steering using a reflectarray integrated with Barium Strontium Titanate (BST) technology. With the application of a bias voltage to the BST cells, a tuneable capacitance is realised. A continuous steering range from $0^\circ$ (boresight) to $25^\circ$ was achieved by altering the bias voltage applied to the BST cells. The design used 45 BST cells which made it complex in design and implementation.

A reflectarray does not incur transmission line losses because the elements are fed by quasi-optical (free-space) means and not via a matched transmission line [27]. It also has the advantage of generating multiple beams which can be used for point-to-multipoint applications by applying different phase shifts to sections of the array. However, the presence of phase shifters within the reflectarray introduces insertion losses. It also has the limitation of predefined beams as with phased arrays since it has phase shifters and does not offer continuous beam steering. With increasing number of elements, the complexity and cost of reflectarray increases.

2.2.4 Parasitic Steering

Fig. 2.8 illustrates the construction of a Yagi-Uda antenna. The Yagi-Uda antenna consists of a single driven element surrounded by several passive parasitics. The term parasitic is used here to describe an element which is not supplied with energy but receives energy by means of electromagnetic coupling.

The parasitic element behind the driven element, in Fig. 2.8, is called a reflector while the parasitic elements in front of the driven element are known as directors. Radiation propagates in the direction of the directors and by adding more directors it is possible to increase the directivity of the antenna. However, as the number of director increases, the electromagnetic coupling reduces and gets to a limit where adding further directors has no appreciable effect on the directivity. Over time, the Yagi-Uda concept has developed from achieving highly directive pattern in one direction to directive patterns towards several angles of interest.

Parasitic steering has been used to improve the directivity of omnidirectional elements
such as monopoles and dipoles. Various configurations of antenna have been proposed in literature, these include Electrically Steerable Passive Array Radiators (ESPAR) [30] (shown in Fig. 2.9), Circular Switched Parasitic Array (CSPA) [31], and disk-loaded monopole array antennas [32]. [33] explains that the parasitics need to be close (within
2.2. Beam Steering Techniques

0.1-0.5 wavelengths) to the active element in order to have good steering effects on the antennas radiation pattern. This separation affects the mutual impedances between the parasitic and driven element and controls the steering angles that can be obtained. The parasitics can be connected to the ground via a switch or a variable reactance [30]. The parasitic acts as a reflector when it is short circuited to ground. When open circuited, the parasitic behaves as a director. The beam is steered towards the director as in the case of Yagi-Uda [90]. In the ESPAR and CSPA configurations the beam is steered in one plane. The orientation of the parasitics determine the plane in which the beam steering can occur.

Parasitic array steering has also been achieved using patch antennas, and a steering range of ±30° and 4 dBi main lobe level has been reported for one driven element [34]. The closer the parasitic is to the driven element, the larger the steering angle. However, the mutual impedance, developed between the parasitic and driven element, will affect the reflection coefficient of the antenna. For the configuration of parasitics around a patch antenna shown in Fig. 2.10, steering could be achieved in one or two dimensions as demonstrated in [34] and [35].

![Figure 2.10: Illustration of 2-D beam steering using parasitics [reproduced courtesy of the Electromagnetic Academy]](image)

The parasitic elements along the x-axis can either act as director or reflector to steer
the beam within the xz-plane. The same applies for the parasitic element along the yz-plane, hence steering in two planes is achieved.

It is worth mentioning that most of the designs for parasitic steering involve only one driven element surrounded by a series of parasitics. In [36], tiny rectangular parasitic elements (pixels) were integrated into a linear array operating at a frequency of 5.6 GHz, which was an extension of [37] based on a single element. By setting up the parasitic pixels in a layer above the plane of the array, they achieved steering in two planes. The parasitic pixels were linked to each other by means of switches, this resulted in different switching configuration. This design increased the complexity and the number of switches needed to implement the design. The authors in [38] also made use of parasitic pixels to steer the beam by $\pm 20^\circ$ and $\pm 40^\circ$. The parasitic pixels in this case were on the same plane as the antenna elements. Switches were used to inter-connect the parasitic pixels and in turn, steer the beam. The design was complex to implement as it had a specific combination needed to achieve a particular steered angle of the beam.

The authors in [39] achieved a wide steering angle of $\pm 50^\circ$ by reducing the size of the ground plane and optimising the location of the switches within the parasitics. The design involved the use of a driven element and four parasitics, two on each side. The ground plane was truncated so that it was partially removed below the parasitics at the two extreme ends of the array. Six switches were used to operate the four parasitics (two switches for each of the parasitics close to the driven element and one switch for the extreme parasitics) and an average main lobe gain of 8 dBi was reported for all the steering directions. In [40], the authors integrated parasitics into a $3 \times 1$ phased array to improve the steering range and resolution of the phased array. This steered the array pattern by a maximum angle of $\pm 15^\circ$ without the use of phase shifters. Smaller steering angles were achieved depending on the combination of the parasitics. The authors in [41] compared the effect of two switches, BAR and HPND Pin Diode, on the radiation pattern of the parasitic array and noted some slight changes in beam direction which is attributed to the difference in capacitance in the pin diodes.

The parasitic steering approach can also be used to realise continuous steering when
tuneable components such as varactor diodes are used to operate the parasitic elements. A few authors have made use of varactor diodes instead of switches to realise continuous parasitic beam steering. In [42], parasitic elements were placed on both sides of a driven rectangular patch antenna and loaded with tuneable capacitance. By changing the bias voltage, the beam of the antenna was steered between $\pm 20^\circ$. This concept was also implemented in [43] and achieved a continuous steering range of $\pm 15^\circ$. The design had tuneable stubs which were used to maintain the resonant frequency of the antenna. With the tuneable capacitance, the range of steering is reduced when compared to the parasitic designs using switches.

### 2.2.5 Integrated Lens Antennas (ILAs)

The concept of integrating a lens over a planar radiating element was developed by Rutledge [44] and has seen much improvement since then. Fig. 2.11 is a schematic of an ILA. The antenna consists of a dielectric lens, of an elliptical or quasi elliptical shape, integrated with a set of driven elements. The elements of the array are offset by a specific distance, $d$, from each other and mounted at the back focal plane of the lens [27], [31], [45][46]. The lens is shaped and positioned in such a way that the radiation from each element appears to emanate from the same focal point. The elements are activated one at a time to implement a range of predefined angles of steering [45], [47]. The ILA achieves a highly directional beam using fewer elements when compared with a phased array antenna. ILAs also employ RF switches instead of phase shifters which have lower losses, offer simplicity and lower cost when compared to phase shifters [48].

The authors in [49] achieved a steering range of $\pm 30^\circ$ and a steering resolution of $7.5^\circ$ with the use of a double lens focusing system. The first lens, an immersion lens (a type of lens that has a high magnification power), was placed above the driven elements while a second lens, an objective lens, was placed over the immersion lens to give finer steering resolution. The design had high cost implications and had a large size. In [46], the recorded steering ranges using ILAs were $\pm 35^\circ$ and $\pm 22^\circ$ and boresight. The measurement was carried out on antennas that operated between 52 GHz and 68 GHz. Gains of up to 18.4 dBi were recorded for a 7.5mm radius lens. The insertion losses
2.2. Beam Steering Techniques

Figure 2.11: Layout of Integrated Lens Antennas (ILAs) [reproduced courtesy of the Electromagnetic Academy]

associated with each switch were 2.5 dB. In [47], the ILA was extended to accomplish 2D steering by using a 2-dimensional eight element array operating at 77 GHz. Scan angles of 17° and 25° were reported with gain of 16 dBi for 15mm lens radius. In [50], 2-dimensional beam steering was achieved using a lens placed over a 4x16 antenna array. The antenna could steer ±4° along the E-plane with four array elements and ±17° in the plane with 16 array elements. However, to achieve large steering angles the offset between each radiating element would need to be increased. Increasing the offset of the radiating elements deteriorates the focusing properties of the lens [45] and increases the gain scan loss. The gain scan loss is defined as the difference in gain measured between the boresight beam direction and the steered beam direction. The gain scan loss depends on the ratio of offset, of the radiating elements, to the radius of the lens [51]. Energy will also be lost (or dissipated) within the materials used to fabricate the lens. The magnitude of these losses will depend on the type of materials used in manufacturing the lens [52].

The ILA does not maximise the use of all the radiating elements to improve gain since each radiating element is turned on individually. The gain achieved is the combined gain of a single element and the lens. This ILA concept was improved in [47] by adding elements in two dimensions to accommodate steering in two planes but the two
dimension improvement still had the limitation of one active element at a time. Also, it only offers predefined steering angles which implies that the connection will be lost if the receiving antenna is displaced from its predefined position. Recent research in improving ILAs used materials with reconfigurable properties, such as metamaterials, to replace conventional lens. This will be discussed in subsection 2.2.9.

2.2.6 Switched Beam Antenna (SBA)

This is a technique where antenna elements are arranged to cover an angle range of interest [6]. Each element covers a section of the total range, as shown in Fig.2.12, and is turned on when there is a need to radiate in that direction. This arrangement is a bit similar to that of ILAs without the lens, with the main difference being that the elements of ILAs are located on a plane.

![Illustration of Switched Beam Antennas](reproduced_courtesy_of_the_Electromagnetic_Academy)

The arrangement, shown in Fig. 2.12 enables beam switching in eight different directions. The antennas are placed at the centre, with each antenna radiating in a different direction as represented by the arrows. However, this solution is inefficient due to high
cost and redundancy. The reason for the redundancy is that eight different feeding networks will be required for the antenna setup, shown in Fig. 2.12 although only one of these elements is turned on at any given time. This redundancy is present for all operating cases except point to multipoint transmission, where more than one antenna element can be turned on simultaneously. Also, this scenario may lead to problems due to high mutual coupling between the antenna elements. This can become a challenge to resolve. Also, if the desired beam direction falls within the region between the sectors (the boundary between consecutive antennas), connectivity cannot be guaranteed, hence a need to have continuous beam steering.

In [53], the authors made use of an active frequency selective surface (FSS) to implement a switched beam antenna. FSS are artificial materials made up of periodic patterns etched into a metallic screen which is often supported by a dielectric material. In the active FSS reported in [53], a pin diode is used to control the reflection coefficient associated with each unit cell of the FSS. By turning on the pin diodes on a section of the FSS, that section of the FSS would appear transparent at a specific frequency while the rest of the FSS would appear opaque at that frequency. A single omnidirectional antenna was placed inside a cylindrical shaped FSS. To steer the beam towards a particular direction the pin diodes in that section of the FSS were turned ON while the rest of the pin diodes were turned OFF. This solution avoids the use of multiple driven elements and redundant feed networks; however, it had increased complexity in design and operation due to the number of switches needed to operate the FSS. In [54], a multiport switched beam antenna based on two concentric disc patches both operating at the same central frequency but with different modes was reported. Four switched beams were achieved based on different combinations of the antennas. The angles achieved were $0^\circ$, $30^\circ$, $45^\circ$ and $-30^\circ$. The main drawback of this approach was the redundancy as only one port was used at a time. When both are used simultaneously, the high coupling of power between the ports will lead to low efficiency for some of the operating modes.

These solutions attempt to reduce the redundancy associated with conventional switched beam antennas, but have reduced performance in realised gain when compared to the conventional.
2.2.7 Traveling Wave Antennas (TWA)

Antennas can be broadly classified as standing wave (resonant) antennas or traveling wave antennas. For resonant antennas, a voltage standing wave pattern is formed by reflections from the open end of the wire. This implies that the current along the line goes to zero at the open end of the line [15].

In the case of travelling wave antennas, the radiating element is terminated by a matched load and a travelling wave is established on the structure. The travelling wave propagates in one direction and there are no reflections. The radiating element could be multiple elements connected together, the resulting radiation pattern will then depend on the shape of each individual element. There are two main categories of travelling wave antennas [15]:

1. Surface (slow) wave antennas: This category is regarded as slow because the guided wave propagates with a phase velocity less than the speed of light in free space. Due to its speed, it does not radiate except at points of discontinuities such as feed point and terminations. With surface wave antennas, it is difficult to produce highly directive radiation patterns and the patterns suffer from significant sidelobes. The desired location of the main beam is determined by the length of the antenna [55]. Examples of surface wave antennas are: helixes, dielectric slabs or rods, and corrugated conductors.

2. Leaky (fast) wave antennas: This is the opposite of surface wave antennas. Its guided wave propagates with a phase velocity faster than the speed of light in free space, hence referred to as fast waves. Due to its speed, it radiates continuously along its length. Unlike the surface wave antenna, it forms highly directed beams at desired angles with a low sidelobe level. The angle of the beam is controlled by the phase constant while the beamwidth is controlled by the attenuation constant of the wave. The phase constant changes with frequency, hence the beam angle for leaky wave antennas is frequency dependent [55].

Travelling wave antennas can be used for wideband frequency scanning as they radiate at all the frequencies within the operating bandwidth [56], meaning that the scan
angle is determined by the signal frequencies. However, beam steering can also be implemented at a fixed frequency in several other ways.

1. Dual feeding points: By feeding the antenna from two different points, one at a time, the beam direction is mirrored along the vertical plane that cuts through the centre of the antenna. This has been demonstrated in [57] and [58]. If both ports are fed at the same time, a boresight beam direction is obtained [57].

2. Parasitic effect/slot loading: In [59], a leaky wave antenna was surrounded by an array of patches separated by a small distance. Switches were used to short the patches to ground thereby changing the distribution of reactance around the antenna. Beam steering was achieved by altering the state of the switches. The authors reported a difference of 16° in the main beam direction between the situation where all of the patches were turned OFF and the situation when all where all of the patches were turned ON. Similar work was carried out in [60] and reported a phase difference of 37° in the main beam direction between the situation where all of the patches were turned OFF and the situation when all where all of the patches were turned ON. In [61], the authors used a moveable dielectric plunger in a slot TWA to achieve beam steering. A continuous steering range of ±25° was reported.

3. Using materials with adjustable properties: Materials, such as graphene and metamaterials have been used to steer the beam of leaky wave antennas at a fixed frequency. In [62] graphene patches loaded with varactor diodes were used as a High Impedance Surface that acts as a ground plane for a leaky wave antenna to steer the beam. Beam steering was achieved by changing the bias voltage of the varactor diode which changed the conductivity of the graphene patches. Metamaterial was used in [63] to steer the main beam by changing the reflection phase of the metamaterial by varying bias voltage of the LC resonant partially reflective surface.

There are challenges associated with the different types of travelling wave antennas. For the surface (slow) wave antenna, the difficulty of achieving highly directive beam
2.2. Beam Steering Techniques

patterns would need to be resolved. For leaky (fast) wave antennas, the main lobe
direction is frequency dependent which will be a huge concern when operating over a
large bandwidth. This frequency dependence causes the antenna to radiate in different
directions at different frequencies within the operating band.

2.2.8 Retrodirective Arrays

The term retro-directive is used in reference to anything that is directed backwards
towards the origin. Any antenna array setup that has this ability of redirecting a
signal backwards towards the source of the signal without any prior information of the
location of the source is a retrodirective array. The idea of retrodirectivity is based on a
corner reflector principle used in radar applications where a ray of light is redirected by
using reflectors perpendicular to each other [64]. Retrodirective arrays have generated
a lot of interest in the research community due to this unique ability and their relative
design simplicity when compared to beamforming techniques. There are two main
architecture topologies for retrodirective arrays: Van Atta array and antennas based
on phase conjugating mixers architectures [65].

Fig. 2.13a shows a schematic of the Van Atta array. It makes use of antenna pairs
connected to each other in such a way as to redirect the signals. The antenna pairs
are equidistant from the centre of the array. While one set of antenna elements act as
the receiver, the other set behave as a transmitter which reradiates the signal back to
the source. The phasing of the signal is achieved using the transmission line sections
shown in the figure. The length of the transmission lines is selected to ensure a reversal
in the phase of the wave [65]. The length of the transmission lines connecting each pair
of antennas are equal.

In the phase conjugation architecture, the phase reversal is achieved in each element
instead of relying on antenna pairs [64], as shown in Fig. 2.13b. The phase reversal
can be implemented using a heterodyne technique. The heterodyne technique involves
using a mixer pumped by a local oscillator with either double the frequency of the
received signal or at an intermediate frequency. This approach offers a flexible struc-
ture which can be implemented easily. Other forms of implementing phase reversal in
2.2. Beam Steering Techniques

2.2.9 Metamaterial Antenna

The concept of Metamaterials was proposed by Veselago in 1968 [67]. They are man-made structures that are designed to exhibit electromagnetic properties that cannot be achieved from naturally occurring structures [68]. Metamaterial have become attractive as their effective permittivity and permeability can assume positive or negative values [69]. With this property, the refractive index of the material can be dynamically tuned using components such as diodes and transistors. The result is an active metamaterial, such materials have been used to produce beam steerable antennas without the use of complex feeding network and phase shifters.

When used in antennas for beam steering, metamaterials are realised by printing a
2.2. Beam Steering Techniques

periodic arrangement of sub-wavelength metallic cells on a substrate [68]. The substrate is located on a layer above the radiating element and used as a frequency selective surface (FSS). To achieve beam steering, the cells are loaded with varactors. By tuning the capacitance of each cell it is possible to alter the permeability and permittivity of the material and hence the refractive index. Several steering ranges have been reported in literature. In [68], a maximum steering angle of 40° was obtained for a metamaterial based Leaky Wave Antenna (LWA). The design reported in [69] achieved a steering range of ±30° with discrete steps of 7.5°. This was done by placing an eight layer metamaterial radome above a 4×16 patch array designed to operate at 4.7 GHz. The insertion loss from each layer of metamaterial is 1 dB which is quite high. This also presents a very bulky steering solution. Also [70] reported a metamaterial antenna having a maximum steering angle of 30° together with a beamwidth of 12.5°. In [71], five metamaterial layers were placed above a horn antenna in order to achieve a maximum steering range of ±30° with intermediate angles of 0°, 10° and 20°. The design was implemented at 5.3GHz. In [72], in order to achieve wider steering angles, the authors made use of multiple radiating elements having different orientations, each with its own metamaterial section. They reported a steering range of ±56°.

Metamaterials have also been used to create lenses for beam steering. In an approach similar to the ILA layers of metamaterial were placed above a radiating element to create a lens. ILAs achieve beam steering by offsetting the radiating elements along the rear focal plane of the lens. With the use of metamaterial lenses, the refractive index of the lens can be tuned. This was carried out in [73] and achieved ±30° of steering angle.

Using metamaterials to steer the antenna beam poses some challenges. Firstly, due to the proximity of the metamaterial layers to the radiation element, the reflection coefficient (S11) of the antenna is affected and the resonance frequency of the antenna is altered for each tuning state of the metamaterial surface. This can be seen from the reflection coefficient plots in [68][72]. This would imply that it would not be convenient to use metamaterial layers as a means of steering for applications that operate over a specific narrow frequency band. Secondly, as mentioned earlier, although the steering range is increased by in proportion to the number of metamaterial layers. Increasing
the number of layers also has the effect of increasing the steering range. Lastly, most authors claim that using metamaterials for steering is less complex when compared to phased array antennas, however integrating the active devices and biasing circuits into each cell of each layer and the multiple layers seems to be quite complex to achieve and a lot more difficult to fabricate.

2.2.10 Amplitude Steering

The signal fed to an antenna element is an alternating signal that has amplitude and phase. A lot of work has been done to steer the main lobe of the radiation pattern of an antenna array by altering the phase of the signal using phase shifters. This is the approach employed in conventional phased arrays. On the other hand, very few authors have attempted to control the amplitude of the signal fed to the array element to achieve beam steering.

In [74], a variable gain amplifier was used to achieve beam steering by altering the amplitude of the signal fed to the array elements. The authors designed a printed discrete lens array where two radiating elements were placed at the focal point on the feed side of the lens. Changing the amplitude of the signals at the feed side of the lens corresponded to a phase shift at the non-feed side which resulted in the steering of the radiation pattern of the antenna. With the use of the variable gain amplifier, continuous beam steering range of $\pm 30^\circ$ was achieved. The main drawback of the design was the loss in power when the antenna was used in the receive mode. The total power received in the lens setup was well below the power received without the lens. This drop in received power was attributed to non-uniform illumination loss, spill-over loss, mismatch loss, and ohmic loss in the design.

Other work that implemented amplitude control using variable gain amplifiers targeted achieving beam shaping and side lobe reduction [75][76][77]. In [76], a variable gain amplifier was integrated into a transmission line phase shifter with the aim of shaping the main beam of the antenna and eliminating high side lobes. The reasons why using variable gain amplifiers in the RF frontend is not common are: 1) behavioural instability in actively loaded antenna noticed in experiments [78], 2) the noise performance of the
amplifier contributes to the total signal to noise ratio (SNR) of the system [79], 3) reduction in efficiency and gain during beam steering, and 4) the biasing circuitry of the amplifier will affect the radiation pattern of the antenna negatively.

### 2.3 Comparison of Beam Steering Techniques and Gaps

Table 2.1 summarises the literature on each steering technique based on some of the figures of merit presented earlier. Most papers reported antenna efficiency and not the insertion loss of the technique.

<table>
<thead>
<tr>
<th>Work</th>
<th>Technique</th>
<th>Steered Range</th>
<th>Efficiency</th>
<th>Steering Resolution</th>
<th>Complexity</th>
</tr>
</thead>
<tbody>
<tr>
<td>[7]</td>
<td>Mechanical</td>
<td>360°</td>
<td>90%</td>
<td>Discrete (4°)</td>
<td>Medium</td>
</tr>
<tr>
<td>[29]</td>
<td>Reflectarray</td>
<td>0-25°</td>
<td>6.76%</td>
<td>Continuous</td>
<td>High</td>
</tr>
<tr>
<td>[38]</td>
<td>Parasitic</td>
<td>±40°</td>
<td>80%</td>
<td>Discrete (20°)</td>
<td>Medium</td>
</tr>
<tr>
<td>[42]</td>
<td>Parasitic</td>
<td>±20°</td>
<td>-</td>
<td>Continuous</td>
<td>Medium</td>
</tr>
<tr>
<td>[49]</td>
<td>ILA</td>
<td>±30°</td>
<td>30%</td>
<td>Discrete (7.5°)</td>
<td>Medium</td>
</tr>
<tr>
<td>[53]</td>
<td>Switched Beam Antenna</td>
<td>360°</td>
<td>65%</td>
<td>Continuous</td>
<td>High</td>
</tr>
<tr>
<td>[61]</td>
<td>Traveling Wave Antenna</td>
<td>±25°</td>
<td>84%</td>
<td>Continuous</td>
<td>Medium</td>
</tr>
<tr>
<td>[71]</td>
<td>Metamaterials</td>
<td>±30°</td>
<td>36%</td>
<td>Discrete (10°)</td>
<td>High</td>
</tr>
<tr>
<td>[74]</td>
<td>Phase shifters</td>
<td>360°</td>
<td>-</td>
<td>Bit dependent</td>
<td>Medium</td>
</tr>
<tr>
<td>[74]</td>
<td>Amplitude steering</td>
<td>±30°</td>
<td>40%</td>
<td>Continuous</td>
<td>Medium</td>
</tr>
</tbody>
</table>

In several aspects, mechanical steering is an ideal means of beam steering since it does not suffer from insertion losses, and the antennas gain and efficiency is not degraded
at any point. However, the slow speed of steering compared to the electronic steering methods is its major disadvantage. RF/analogue beamforming with phased arrays offer fast steering speeds, but due to high cost and insertion loss of phase shift shifter which degrades the system performance, other low loss alternatives are needed for low cost beam steering applications. Reflectarrays have the limitation of large size, complexity, and high cost. Parasitic steering has the limitation that only one driven element employed which will not be suitable for high gain applications.

Integrated lens antennas suffer from large size due to use of a lens, and cause pre-defined steering resolution. Switched beam antennas have the limitation of having a high degree of redundancy as well as high cost which also makes them unsuitable for low cost applications. Traveling wave antennas also suffer from large size and radiate energy to unwanted directions. Retrodirective arrays have the limitation of retransmitting any signal that they receive towards the originating signal; hence it would not be suitable for applications where there is a need to attenuate interference. Metamaterial antennas suffer from large size and complexity as well as high insertion losses. Amplitude steerable antennas using variable gain amplifiers suffer from instability in antenna behaviour, high SNR, and reduced efficiency and gain while steering.

Hence, a beam steering technique that offers high gain, low loss, low cost, small size and low complexity is needed for applications such as WLAN/WPAN/WiGig high data rate point-to-point link, 5G millimetre wave communication, anti-collision and hand held radars.

2.4 Summary

In this chapter, a detailed survey of beam steering techniques have been presented. The high insertion loss of the phase shifters used in conventional phased arrays which deteriorates the efficiency of the antenna, opens up a huge gap for research into beam steerable antennas techniques. The lack of fine steering resolution in other low loss approaches also gives room for further research into low loss, low complexity, and fine steering angle beam steering solutions which is the focus of this work.
Chapter 3

Isolation Bandwidth Enhancement for Power Dividers

Phased array steering, discussed in Chapter 2, is the conventional approach for beam steering for applications such as massive Multiple Input Multiple Output (MIMO) systems. Such systems would require a mutual coupling better than 15 dB between the antenna elements in the MIMO system. The main approach to feeding the array elements in any multiple array system is power dividers.

Power dividers (PDs) can be classified into one of two categories, namely: equal and unequal power dividers. For equal split power dividers, the Wilkinson Power divider (WPD) has been well researched. The limitations of the WPD is its narrow isolation bandwidth between the output ports. Typically, the 15 dB fractional bandwidth of the isolation response for the Wilkinsons divider is between 30-50% [80] which will be insufficient for wideband applications. This chapter will investigate the techniques and approaches used to enhance the isolation bandwidth of power divider and aslo present a novel approach which has a reduced complexity in relation to other approaches in literature.

The chapter is structured as follows: Section 3.1 presents an overview of traditional (equal) power dividers. Section 3.2 presents a review of techniques used in isolation bandwidth improvement. Section 3.3 presents the geometry of the proposed compen-
sated power divider which aims to reduce the complexity in achieving wide isolation bandwidth. Section 3.4 presents the odd and even mode analyses of the compensated power divider. Section 3.5 presents the analyses of the return loss and isolation bandwidth between the output ports. Section 3.6 presents results and discussion for a fabricated compensated power divider. Section 3.7 presents guidelines for designing the compensated power divider and a summary of the chapter is presented in Section 3.8.

3.1 Overview of Traditional Power Dividers

A lot of research has been carried out on power dividers which makes it challenging to give a comprehensive overview of power dividers. For clarity and to relate the work done in this thesis on power dividers to the existing literature, a classification system for power dividers will be needed. One way of classifying power dividers is according to the nature of the input and output ports [81]. Based on this criterion, there are; single-ended-to-single-ended power dividers (traditional power dividers), balanced-to-balanced power dividers, balanced-to-single-ended power dividers and single-ended-to-balanced power dividers. The focus in this review is on the traditional type of power dividers.

Traditional power dividers can be further sub-divided into the following types: T-junction power divider (TJPD), Fig. 3.1a, Resistive power divider (RPD), Fig. 3.1b, Wilkinson power divider (WPD), Fig. 3.1c, Gysel power divider (GPD), Fig. 3.1d, and Bayley Polygon power dividers (BPD), Fig. 3.1e. These power dividers are shown in Fig. 3.1. The T-junction power divider, Fig. 3.1a, is a 3-port network that can be implemented in any type of transmission line technology such as microstrip. The TJPD can be described as lossless, however it cannot be matched at all the ports and the outputs are not isolated [1]. The RPD, Fig. 3.1b, makes use of lumped element resistors and achieves matching at the ports of the device. However, the RPD is lossy as it makes use of lossy components, and the power at the output ports are 6 dB below the input power [1]. This makes the RPD a bad choice at microwave frequencies.

The Wilkinson Power Divider (WPD), Fig. 3.1c, was developed by E. J. Wilkinson
3.1. Overview of Traditional Power Dividers

(a) T-Junction PD

(b) Resistive PD

(c) Wilkinson PD

(d) Gysel PD

(e) Bagley PD

Figure 3.1: Traditional power dividers
in 1960 [82]. It is an example of the wider class of T-junction power dividers. In comparison with a conventional TJPD the WPD features transmission lines which differ in width along with an isolation resistor that is located between the output arms. This power divider has the advantage of being matched at all the ports, low loss and the output ports are well isolated from one another [1]. The WPD can have any number of output ports and any desired dividing ratio. The main limitation of the WPD is that good isolation between the output ports is only provided over a narrow operating frequency bandwidth which makes it unsuitable for wideband applications. Additionally the WPD has low power handling capability which is limited to 100W [83]. The power handling capacity is limited by the efficiency with which the lumped element resistor can dissipate heat. This can be improved by adding a heat sink to the lumped element. Various techniques have been used to improve the performance of the WPD which will be discussed in the next section.

The Gysel power divider (GPD), Fig. 3.1d, [83][84], developed by U. H. Gysel in 1975, was designed to address the need for dividers with higher power handling capabilities. In the design of the GPD, there is a provision for heat sink, which makes use of two grounding resistors linked to the external ground. Like the WPD, the GPD can be designed for any number of output ports and power dividing ratios. However, the GPD suffers from limited isolation bandwidth, large size of device, high cost and complexity of fabrication [85].

The Bagley Polygon Divider (BPD), Fig. 3.1e, is a type of power divider that does not make use of any lumped element component. The BPD was developed by G. Bagley of the Royal Aircraft Establishment, in UK [86]. The BPD can be designed for any odd number of outputs that are matched to 50Ω at the operating frequency [87]. The BPD has a large size which could be a huge disadvantage for applications with space constraints. It also suffers from poor isolation between the output ports [88].

A common thread in the traditional power dividers is the need for better isolation between the output ports. The WPD has the best isolation among the traditional power dividers, however this isolation still needs to be improved for wideband applications. Several techniques have been used to improve the isolation bandwidth of the WPD and
it will be important to investigate these techniques based on the following figures of merit:

1. **15 dB isolation bandwidth**: For multiple antenna systems, the choice of 15 dB isolation between each antenna element is essential to limit mutual coupling between the antenna elements within the system. This is the bases for the choice of 15 dB isolation bandwidth level between the output ports of the power dividers. To quantify the isolation bandwidth, either the fractional bandwidth, or the bandwidth ratio, or the frequency range of the 15 dB bandwidth will be specified. The isolation bandwidth ratio is the ratio of the upper frequency limit to the lower frequency limit using the reference 15 dB value. Some of the papers in the literature fail to specify the designed frequency of the power divider; hence it would be inaccurate in such case to estimate the fractional bandwidth.

2. **Insertion loss**: The loss incurred by the technique.

3. **Complexity of designing, fabricating, and operating the device using the technique**. This would take into account the number of lumped elements, transmission line sections, and transitions segments needed to implement the technique. The conventional WPD will be used as a reference to compare the complexities of other techniques.

4. **Size of the device**.

### 3.2 Techniques used in Isolation Bandwidth Improvement

The major techniques used to improve the return loss and isolation bandwidth of traditional power dividers are;

1. **Multistage matching isolation sections and Tapered lines**

2. **Isolation networks**

3. **Coupled Lines**
3.2. Techniques used in Isolation Bandwidth Improvement

4. Transition

5. Combination of two or more techniques

These techniques are discussed in detail in the following subsections.

3.2.1 Multistage Matching Isolation Sections and Tapered Lines

Fig. 3.2 shows the layout of a multistage matching isolation sections. In 1968, Cohn [89] presented detailed analyses on cascaded multiple sections of transmission line designed to improve the operating bandwidth and isolation of the conventional WPD. He showed that when using a seven-stage isolation section, it was possible to achieve a 20 dB isolating bandwidth of 10:1. Where the first number (10, in this case) represents the upper 20 dB cutoff frequency (denoted $f_2$) and the second number (1, in this case) represents the lower 20 dB cutoff frequency (denoted $f_1$). For this particular example $10/1 = f_2/f_1$. The number of lumped element resistor required within the design is equal to the number of sections needed. For this reason, a seven-stage matching isolation section is physically quite large. It also suffers from the insertion losses of the seven lumped elements and transmission line sections.

Several authors have implemented the multistage matching isolation sections in recent times. The authors in [90] used the method of least squares (MLS) to optimise the values of the transmission line impedances for each multistage section to achieve an optimal performance in isolation bandwidth with fewer stages of isolation sections. For
3.2. Techniques used in Isolation Bandwidth Improvement

a four-stage isolation section, they achieved an isolating bandwidth ratio of 5:1. The optimisation improved the isolation between the ports at the operating frequency but reduced the 15 dB bandwidth when compared with the bandwidth before optimisation. The design has the limitation of large device size and increased insertion loss from the four lumped elements resistors needed to implement the design and the additional transmission line sections. It also suffers from high complexity in design due to the choice of optimisation algorithm.

In [91], the authors achieved a bandwidth ratio 11:1 using three step impedance sections and five multistage matching isolation sections. The power divider was fabricated on silicon with PolyStrata technology. The PolyStrata technology is achieved using a sequential building process that deposits layers of different metals over a substrate to form a coaxial waveguide microstructure [92]. The divider operated from 2-22 GHz, however the wide isolation bandwidth was reported using 11 dB as reference which is below the industry standard of 15 dB. Due to rigor in the design and fabrication technique, the cost and complexity of implementing this power divider will be very high. In [80], the authors designed a three-stage matching isolation section and achieved an isolation bandwidth ratio of 4:1. The power divider based on coplanar waveguide (CPW) technology and operated from 800 MHz to 3.2 GHz. Although the authors reported an insertion loss of 0.8 dB, the size of the power divider was large due to the use of multiple sections of matching isolation. In [93], the authors combined a single matching isolation section with quarter wavelength short circuited stubs and coupled lines to achieve bandwidth improvement. The power divider was designed to operate from 3.1-10.6 GHz and had four short circuited stubs. However, the 15 dB isolation bandwidth ratio was 2:1, from 5-10 GHz and did not cover the specified operating frequencies. The design was compact, but had an insertion loss of 3.3 dB and has a highly complex structure.

In [94], the authors achieved wide isolation bandwidth by making use of tapered transmission lines and lumped element resistors. Depending on the number of resistors added to the tapered transmission line, the power divider achieved a 15 dB isolation bandwidth ratio of about 9:1, from 2-17 GHz when five resistors were used. The main limitation of the designed power divider was that the amplitude imbalance, towards the
upper end of the operating frequency band, was worse than 0.5 dB. This limits the use of the power divider over the entire operating bandwidth. In addition to the amplitude imbalance, the complexity of the design is high due to the challenge in positioning of the lumped element resistors needed to improve the isolation bandwidth.

The challenge with using cascaded multistage matching isolation network is in reducing the size of the device along with the insertion loss associated with the additional transmission line segments. The technique has a high degree of complexity due to the process of obtaining the values of the transmission line impedance for each stage of the matching isolation as well as the value of the lumped element resistor needed for each section. Ripples have also been noticed in the transmission coefficient (S21), which is attributed to the discontinuities at the step impedance [94].

3.2.2 Isolation Network

![Diagram of isolation network](image)

Figure 3.3: Schematic layout of isolation network

Fig. 3.3 shows the schematic of the approach used in isolation network. A network of lumped element used to replace the isolation resistor in conventional WPD. The approach was first used to achieve dual frequency band power divider operation [95].
In recent years, this has found application in improving the return loss and isolation bandwidth of power dividers. In [96], the authors used a series RLC isolation network to achieve both dual frequency operation and improved isolation bandwidth. The design achieved a 15 dB isolation bandwidth ratio of 2.6:1, operating from 0.75-2 GHz with a compact size. However, the insertion loss was not recorded. This is expected to be relatively high as the design makes use of three lumped element components. The design also presents some complexity in the obtaining the values of the resistor, inductor and capacitor to be used. In [97], the authors used a parallel LC isolation circuit shorted to ground to improve the isolation bandwidth. A wide isolation bandwidth was achieved from 0-2.5GHz. However, the design was for a filtering power divider that had a narrow return loss bandwidth.

In [98], the authors designed an isolation network comprising of two series RLC circuits which were used to replace the isolating resistor. The isolation network was shorted to ground via a transmission line section at the junction between the two RLC circuits. The design made use of six lumped elements and it had a 15 dB isolation bandwidth ratio of 4.8:1, operating from 1.5-7.3 GHz. The design had an insertion loss of 1 dB and was compact. However, the return loss bandwidth (1.8-6.8 GHz) of the power divider was relatively narrow as the isolation network did not improve the operating return loss. The complexity of the design was high as was the insertion loss. In [99], the authors designed an isolation network having five lumped elements and two stubs shorted to ground for isolation bandwidth improvement. The design had a 15 dB isolation bandwidth ratio of 3:1, from 0.5-1.5 GHz, for a 1 GHz design. The reported insertion loss of the design was 0.2 dB. The design improved both the 10 dB return loss bandwidth and the 15 dB isolation bandwidth. The main drawbacks of the design are its complexity and difficulty in fabrication.

A common thread in the use of isolation networks to improve the isolation bandwidth, of a power divider, is the complexity involved with the design. The use of lumped element components also presents some challenges in terms of obtaining the precise values and insertion loss. The isolation bandwidths achieved, using this technique, are lower than those associated with multistage matching isolation sections but the use of the isolation network presents a more compact design and less insertion loss.
3.2.3 Coupled Lines

When unshielded transmission lines are close to each other, the power they carry can be coupled from one line to the other due to the interaction of their electromagnetic fields. For coupling to take place, the transmission lines must be symmetric, having the same width and relative position to the ground [1]. The closer the transmission lines are to each other, the stronger the coupling between them. The coupled lines improve isolation by suppressing and weakening the surface current propagating in the transmission line with a physical separation between the coupled lines. With the use of coupled lines, compact power dividers can be designed. Fig. 3.4 shows the schematic of the coupled line approach. The first work to use coupled lines to improve the isolation of a power divider was reported in 1971 [100]. The author combined two-stage matching isolation network with coupled lines to achieve an isolation bandwidth better than 20 dB from 2.9-12 GHz (4:1) having a much-reduced physical size. In [18], in addition to shorted stubs, the authors made use of coupled lines. This improved the isolation bandwidth of a single stage Wilkinson power divider.
3.2. Techniques used in Isolation Bandwidth Improvement

In [101], the authors use the coupled line technique to reduce the size of the conventional Wilkinson power divider. The design showed little improvement in return loss and isolation bandwidth when compared to the conventional Wilkinson power divider. The design had one section of coupled lines and a lumped element isolation resistor. A major drawback of the design was that the isolation bandwidth was shifted 300 MHz from the designed centre frequency of 2.4 GHz. In [102], the authors designed a dual band Wilkinson power divider using coupled lines. By making the power divider dual band, they improved the isolation bandwidth of the power divider to have a 15 dB isolation from 3-11 GHz (3.6:1). The design had two sections of coupled lines and two lumped element resistors. The design was compact but had some degree of complexity in obtaining the widths of the coupled line sections.

The authors in [103] designed Wilkinson power divider with a wideband isolation bandwidth using coupled lines. The Wilkinson power divider structure was connected to the input and output ports by means of coupled lines. The designed had three coupled-lines section and retained the isolation resistor in conventional Wilkinson power divider. The power divider was designed at 2 GHz and achieved a 15 dB fractional isolation bandwidth of 65%, from 1.2-2.6 GHz (2:1). The design was compact but had a high degree of complexity in design due to the irregular shape of the transmission lines. In [104], a planar dual band coupled-line power divider was designed. The power divider had four coupled-line sections and had a lumped element resistor. The device operated at two distinct narrowband frequencies, namely 1 GHz and 2.7 GHz. The isolation only covered the frequency of operation and hence would not be regarded as wideband. The design was very complex and the insertion loss incurred by the design was not stated.

To summarise, the use of coupled lines improves the isolation bandwidth of the power divider when compared with the conventional Wilkinson power divider. However, when compared to the two previously discussed methods, the improvement in the bandwidth is poor. The coupled-line technique achieves a compact structure and offers flexibility in design parameters [102], but it has a high degree of complexity, especially in designing the width of coupled-lines and the spacing between them [1]. Most of the authors did not highlight the insertion loss incurred in their designs.
3.2.4 Transitions

In microwave engineering, transitions can be described as points where two transmission line technologies intersect, such as microstrip line and slotline, as shown in Fig. 3.5. The use of transitions has been encouraged as it can be used for bandwidth improvement. For the microstrip-to-slotline transition, the characteristic impedance of the microstrip and slotline controls the bandwidth that can be achieved. To achieve maximum bandwidth, the characteristic impedance of the microstrip line will need to be 2.618 times the characteristic impedance of the slotline [105]. This approach makes use of both sides of the substrate, i.e. the slotline is etched on the ground plane.

Several authors have implemented transitions to improve the return loss and isolation bandwidth of the conventional Wilkinson power divider.

In [106], the authors used a microstrip-to-slotline transition to design a wideband Wilkinson power divider. The design used two transitions, with the slotline etched on the ground plane. They achieved a 10 dB isolation between 4-12 GHz. However, the threshold of 10 dB is lower than the industry accepted level of 15 dB. The design had a large size and high complexity due to the reliance on optimising the shape of the microstrip line. In [107], the authors used a three-layer approach involving two substrates to design a wideband power divider. The two substrates were placed back-to-back with both ground planes touching each other. The slotline was etched into the
3.2. Techniques used in Isolation Bandwidth Improvement

ground plane while the input and output ports were placed on the bottom and top faces of the substrate stack-up, respectively. The design retained the isolation resistor used within the conventional Wilkinson power divider and achieved a 10 dB return loss bandwidth of 90%, operating from 3.5-8 GHz. However, the 15 dB isolation bandwidth was narrowband when compared to the 10 dB return loss (from 4.5-6.5 GHz). An insertion loss of about 1 dB at 8 GHz was reported.

In [108], two microstrip-to-slotline transitions were used to design a wideband 180° hybrid. The design had no lumped element resistor and had a 10 dB return loss bandwidth between 4-9.5 GHz. The authors reported a 10 dB isolation bandwidth between 3-10 GHz, however the 15 dB industry accepted isolation bandwidth was between 5-8.5 GHz, a ratio of 1.7:1, which is regarded as narrowband. The design was compact, with dimensions $3.8 \times 2.6 \text{cm}^2$. The design had a high complexity as the shape of the slots were irregular and need to be properly calculated to achieve the desired performance.

In [109], the author designed a multilayer power divider for C-band applications. The design incorporates two substrates placed back-to-back with a slot etched in the ground plane, located in the centre of the stack-up. The design had two microstrip-to-slotline transitions and achieved a wideband return loss bandwidth from 4-8 GHz. However, the 15 dB insertion loss was narrowband (3.5-4.5 GHz), not covering the required C-band frequencies. The author reported an insertion loss of 0.6 dB. In [110], the author designed a three layer power divider using the microstrip-to-slotline transition. He reported a wide 10 dB isolation bandwidth behaviour from 5-11 GHz. However, the isolation was worse than 15 dB over the operating range of the power divider. The use of transitions improved the return loss and isolation bandwidth of the power divider to values better than 10 dB. However, for the 15 dB industry accepted level, the technique did not have as much effect. The use of multilayers in this technique increases the cost and difficulty of fabrication which might be unwanted and discouraged for low cost applications. The technique could be made to be compact, but the complexity of this technique is high.

This technique of bandwidth improvement for power dividers does not perform as well as the other techniques previously discussed.
3.2.5 Combination of Two or more Techniques

Several authors have combined two or more techniques to improve the return loss and 15 dB isolation bandwidth of power dividers. It was mentioned that the authors in [93] combined short circuited stub lines and coupled lines to achieve a wide bandwidth. The authors in [111] combined microstrip-to-slotline transition with coupled line to improve the 15 dB isolation bandwidth of a power divider operating at 2.5 GHz centre frequency. Four slot shapes were etched on the ground plane and the separation between the coupled lines was 0.2 mm. A 15 dB isolation bandwidth from 1.2-4.2 GHz (3.5:1 bandwidth ratio) was reported along with a 10 dB return loss fractional bandwidth of 150%. The power divider had a high design complexity and would be challenging to fabricate. The same authors incorporated shunt open ended stubs in [112] to further improve the performance of the power divider. They achieved 100% 16 dB fractional isolation bandwidth and recorded an insertion loss of 0.5 dB. The number of slots used in the design was reduced compared to their previous work but the power divider occupied a large area. The power divider had two lumped element resistors and it had a high complexity.

In [113], the authors achieved a wideband performance by combining coupled lines and short circuit stubs. The design had a lumped element resistor and the short ended stub was hardwired to the ground plane. The 10 dB return loss bandwidth of the power divider was between 3-10.5 GHz. The power divider had a wide 10 dB isolation bandwidth which matched the return loss bandwidth, but the 15 dB isolation bandwidth was between 5-10 GHz (2:1 bandwidth ratio). The designed power divider was compact in size and had an insertion loss of 1.2 dB.

In [114], the authors combine a multistage matching isolation with parallel branch lines to reduce the size and maintain performance of the power divider. The parallel branch lines replaced the resistors in the conventional multistage matching. The power divider had five stages, made use of one chip resistor, and was designed to operate from 18.8-40 GHz. The reported 15 dB isolation bandwidth was 78% (2:2:1 bandwidth ratio). The insertion loss for the power divider was about 2 dB at the upper operating frequency band which was the main drawback of the design.
3.2. Techniques used in Isolation Bandwidth Improvement

In [115], the authors combined the coupled line method with the use of composite right-left-handed (CRLH) transmission lines to achieve improved isolation bandwidth. The power divider had three coupled line section and two lumped element resistors. The CRLH transmission line was used to link one of the ports to the power dividing section. The power divider was designed to operate at 2.7 GHz and had a 15 dB return loss bandwidth of 82.9%. The isolation between the ports was better than 17.2 dB over the operating range of the power divider and an insertion loss of 0.1 dB was reported. The drawbacks of the designed power divider are the complexity in design and fabrication associated with the coupled line technique. Additionally, the power divider is large and would be quite expensive to produce due to the use of the CRLH transmission line.

The use of a combined approach to improve the return loss and isolation bandwidth of the power divider increases the complexity of the design when compared with the conventional Wilkinson power divider. It also increases the cost and, in some cases, increases the size of the power divider.

3.2.6 Comparison of Techniques used for Isolation Bandwidth Improvement

Based on the stated figures of merit, Table 3.1 shows a comparison of recent works in literature. All the techniques have high complexity due to the number of components needed in design and fabrication. For large isolation bandwidth using multistage impedance matching and tapered lines, the size of the designed power divider will be large which will incur high insertion losses. These techniques will not be suitable for ultra-wideband applications that have space limitations. Hence there is a need to have an ultra-wideband return loss and isolation bandwidth power divider that has a similar complexity and size as the conventional Wilkinson power divider.

This work presents a novel approach to realising ultra-wide bandwidth isolation in power dividers using a compensated impedance. Design and analyses of the novel compensated power divider are presented in the following sections.
3.3 Geometry of the Compensated Power Divider

Table 3.1: Comparison of recent works for bandwidth improvement

<table>
<thead>
<tr>
<th>Work</th>
<th>Technique</th>
<th>15 dB Isolation {ratio, (BW GHz)}</th>
<th>IL (dB)</th>
<th>Complexity</th>
<th>Size</th>
</tr>
</thead>
<tbody>
<tr>
<td>[80]</td>
<td>Conventional WPD</td>
<td>1.6:1 (0.75-1.25)</td>
<td>-</td>
<td>1 lumped element</td>
<td>0.23\lam \times 0.33\lam</td>
</tr>
<tr>
<td>[80]</td>
<td>Multistage IM</td>
<td>4:1 (0.8-3.2)</td>
<td>0.8</td>
<td>3 TL sections, 3 lumped elements</td>
<td>0.18\lam \times 0.41\lam</td>
</tr>
<tr>
<td>[94]</td>
<td>Tapered TL</td>
<td>8.5:1 (2-17)</td>
<td>-</td>
<td>1 TL section, 5 lumped elements</td>
<td>2.4\lam \times 2.7\lam</td>
</tr>
<tr>
<td>[99]</td>
<td>Isolation Network</td>
<td>3:1 (0.5-1.5)</td>
<td>0.2</td>
<td>2 TL sections, 5 lumped elements</td>
<td>0.16\lam \times 0.13\lam</td>
</tr>
<tr>
<td>[103]</td>
<td>Coupled Lines</td>
<td>2:1 (1.2-2.6)</td>
<td>-</td>
<td>3 TL sections, 1 lumped element</td>
<td>0.37\lam \times 0.44\lam</td>
</tr>
<tr>
<td>[108]</td>
<td>Transition</td>
<td>1.7:1 (5-8.5)</td>
<td>-</td>
<td>2 Transitions</td>
<td>0.51\lam \times 0.82\lam</td>
</tr>
<tr>
<td>[112]</td>
<td>Trans + Coupled</td>
<td>3.5:1 (1.2-4.2)</td>
<td>0.5</td>
<td>2 Transition, 3 TL sections, 2 lumped elements</td>
<td>0.27\lam \times 0.47\lam</td>
</tr>
</tbody>
</table>

3.3 Geometry of the Compensated Power Divider

Fig. 3.6 shows the structures of the conventional Wilkinson Power Divider (WPD) and the compensated power divider (CPD). The fundamental difference between the CPD and WPD is the presence of the compensating impedance ($Z_1$) before the dividing arms ($Z_2$). The role of the compensating impedance ($Z_1$) is to either increase or decrease the return loss bandwidth of the divider depending on the requirements of the system. The dividing arms of the conventional WPD always have an impedance of $Z_0\sqrt{2}$, while, in the CPD, $Z_1$ and $Z_2$ can take any value.

The focus of research into CPDs has been to improve the 10 dB return loss bandwidth of the device. To achieve good performance of power dividers for MIMO and other multiple antenna applications, the isolation between the output ports should better than 15 dB over the operating frequencies. This thesis will show how the 15 dB isolation bandwidth of the device can be improved to achieve ultra-wideband performance based on optimising the compensated impedance ($Z_1$).
3.4 Odd and even mode analyses of Compensated Power Divider

Odd and even mode analyses make use of superposition and circuit symmetry to analyse circuits with high complexity. The structure of the CPD, shown in Fig. 3.6b, is symmetrical about the horizontal centre line after the compensated impedance $Z_1$. Hence to achieve a symmetric structure, the CPD is redrawn, as shown in Fig. 3.7, with the ports replaced by their characteristic impedances ($Z_0$). To achieve a symmetrical circuit, an additional input port resistance and compensated impedance were added in parallel to the existing structure.

To analyse the circuit, additional voltage sources, $V_{g2}$ and $V_{g3}$ are defined for the circuit. For the odd mode [1]

$$V_{g2} = -V_{g3};$$

$$V_{g2} + V_{g3} = 0V_0$$

(3.1)

and for the even mode,

$$V_{g2} = V_{g3} = 2V_0;$$

$$V_{g2} + V_{g3} = 4V_0$$

(3.2)

Superimposing the two modes would produce excitation voltages of 0 and $4V_0$ for the odd and even mode analyses respectively. For the odd mode excitation, the circuit...
3.4. Odd and even mode analyses of Compensated Power Divider

appears to be shorted along the line of symmetry in Fig. 3.7. The top portion of the odd mode circuit is shown in Fig. 3.8.

To evaluate the odd mode voltage \( V^o \), the impedance seen looking into \( V^o \) in Fig. 3.8 is obtained. The port resistance, \( Z_0 \) and the compensated impedance are shorted and so do not contribute to the odd mode voltage. The resultant circuit is shown in Fig. 3.9.

The short circuit before the impedance \( Z_2 \) is translated into an open circuit by the quarter wave section \( Z_2 \). Hence, the voltage \( V^o \) can be obtained using KirchhoFFS current law, as shown below;

\[
V_{g2} - iZ_0 - ir/2 = 0
\]

\[
i = V_{g2}/(Z_0 + r/2)
\]

\[
V^o = ir/2 = \frac{rV_{g2}}{Z_0 + r/2}
\]
3.4. Odd and even mode analyses of Compensated Power Divider

The odd mode voltage for the section below the line of symmetry in Fig. 3.2 is the inverse of the voltage for the top section as $V_{g2} = -V_{g3}$.

For the even mode excitation, there exists an open circuit along the line of symmetry shown in Fig. 3.7 due to the equal excitation. The top portion of the even mode circuit is shown in Fig. 3.10. To derive the even mode voltage, $V_e$, it will be necessary to compute the impedance ($Z_e$) seen looking into $V_e$.

The characteristic impedance ($2Z_0$) of port 1 is matched to the characteristic impedance ($Z_0$) of port 2, in two stages, involving the compensated impedance ($Z_1$) and the impedance of the dividing arm ($Z_2$). For the first stage,

$$2Z_1 = \sqrt{2Z_0 \times Z_{T1}};$$

$$Z_{T1} = \frac{2Z_1^2}{Z_0}$$ (3.4)

Where $Z_{T1}$ is the port impedance matched to input of the dividing arm impedance $Z_2$. $Z_{T1}$ is then matched to the $Z_e$ by the impedance $Z_2$. $Z_e$ is the impedance seen across $V_e$.

$$Z_2 = \sqrt{Z_{T1} \times Z_e};$$

$$Z_e = \frac{Z_2^2}{Z_{T1}}$$ (3.5)

Figure 3.9: Resultant Circuit

Figure 3.10: Even mode circuit
Substituting the value of $Z_{T1}$ in eqn. (3.4) into eqn. (3.5) gives the value of $Z_e$:

$$Z_e = \frac{Z_2^2 Z_0}{2 Z_1^2}$$  \hfil (3.6)$$

Applying the voltage dividing rule around the loop between $V^e$ and the even mode excitation,

$$V^e = V_{g2} \frac{Z_e}{Z_0 + Z_e}$$  \hfil (3.7)$$

Substituting the value of $Z_e$ in eqn. (3.6) into eqn. (3.7)

$$V^e = \frac{V_{g2}}{2 Z_1^2} \left( \frac{Z_2^2 Z_0}{Z_0 + \frac{Z_2^2 Z_0}{2 Z_1^2}} \right)$$  \hfil (3.8)$$

A relationship between the compensated impedance, dividing arm impedance, isolation resistance, and port impedance can be derived, using the expressions for the odd and even mode voltages, derived above. From voltage division, for a constant value of $V_{g2}$, $V^o$ and $V^e$ will be equal (for the section of the circuit above the line of symmetry as the only source of voltage drop for both even and odd mode circuits is the port impedance, $Z_0$, closest to the source of excitation. Hence, $V^o = V^e$ for top portion of symmetry and $V^o = -V^e$ for bottom section of the circuit.

For the condition $V^o = V^e$;

$$\frac{r V_{g2}/2}{Z_0 + r/2} = \frac{V_{g2}}{2 Z_1^2} \left( \frac{Z_2^2 Z_0}{Z_0 + \frac{Z_2^2 Z_0}{2 Z_1^2}} \right)$$  \hfil (3.9)$$

Eqn. (3.9) is further reduced to

$$\frac{Z_2^2}{2} \left( Z_0 + \frac{r}{2} \right) = \frac{r}{2} \left( Z_1^2 + \frac{Z_2^2}{2} \right)$$  \hfil (3.10)$$

$$\frac{Z_2}{Z_1} = \sqrt{\frac{r}{Z_0}}$$  \hfil (3.11)$$

$$Z_2 = \sqrt{k} Z_1$$  \hfil (3.12)$$
3.5 Return loss and output port isolation Analyses

Where \( k = r/Z_0 \) and \( r \) is the isolating resistance. Eqn. (3.12) shows the relationship between \( Z_1, Z_2, Z_0, \) and \( r \). When \( r = 100\Omega \), and \( Z_1 = Z_0 = 50\Omega \) i.e. \( k = 2 \), the CPD becomes a conventional WPD. To improve the performance of the CPD, the value of \( k \) is altered, depending on the design requirement.

Considering the case when \( V_o = -V_e \), the relationship between the ratio of the dividing arm impedance \( (Z_2) \) to the compensated impedance \( (Z_1) \) is given by eqn. (3.13).

\[
\frac{Z_2}{Z_1} = j\sqrt{\frac{r}{Z_0 + r}} \quad (3.13)
\]

It should be noted that Eqn. (3.13) results in an imaginary number. It is impractical to realise an imaginary impedance in microstrip technology, hence only Eqn. (3.12) is used for analysis. In the next section, analyses will be carried out to show how changing the value of \( k \), in eqn. (3.12), affects the 10 dB bandwidth of the return loss curve and the 15 dB isolation bandwidth of the CPD.

### 3.5 Return loss and output port isolation Analyses

Two properties that are important for the CPD are the 10 dB return loss bandwidth and the 15 dB isolation bandwidth. In order to analyse the effect of changing the value of \( k \) on the performance of the CPD, a new CPD was designed and simulated using the Advanced Design System (ADS) circuit simulator from Keysight. The new CPD was designed to operate at 6 GHz, this frequency was chosen arbitrarily. The model in CPD model in ADS is shown in Fig. 3.11

For a given impedance ratio \( (k) \), the compensated impedance \( (Z_1) \) was varied from 40Ω to 80Ω. The 10 dB, 15 dB and 20 dB fractional return loss bandwidth, for each value of \( Z_1 \), was acquired through simulation and plotted against the compensated impedance. This exercise was carried for different values of \( k \) and the results are plotted in Fig. 3.12. On inspection of Fig. 3.12 it is clear that for ultra-wideband applications at the 10 dB level, a low value of \( Z_1(< 55\Omega) \) is required. Increasing the value of the compensated impedance causes the device to become narrowband. This behaviour is consistent irrespective of the values of \( k \).
3.5. Return loss and output port isolation Analyses

The CPD becomes a conventional WPD when $k = 2$ and $Z_1 = 50\Omega$. The properties of the WPD can be improved to achieve wider operating bandwidth by employing lower values of compensating impedance together with a different impedance ratio ($k$). Altering the value of the compensated impedance, while maintaining a constant impedance ratio ($k$) has the effect of causing a slight variation in the return loss bandwidth. Fig. 3.13 shows the change in return loss at the operating frequency as the ratio ($k$) is altered.

The optimum value of return loss (better than 50 dB) is achieved when $k = 2$. This is the ratio used in the conventional WPD. Based on the priorities of a particular design, the return loss requirement can be relaxed in order to optimise other properties of the
power divider. For $k > 3$ a wider 10 dB fractional bandwidth can be achieved at the cost of a poorer return loss, compared to the WPD, at the operating frequency. For values of $k$ beyond 3.8, the return loss is worse than 10 dB and the device appears to be operating in two distinct frequency bands (i.e. dual band). The output ports (port 2 and 3) remain matched as the impedance ratio $k$ is altered.

The value of compensated impedance ($Z_1$) that produces the widest 15 dB return loss bandwidth was determined for a constant dividing ratio ($k$). 10 dB fractional bandwidth is conventionally used as reference. However, the 10 dB return loss bandwidths corresponding to the various compensated impedance values, considered within the parametric study, were too close together to enable clear demarcation between them. This motivated the choice of 15 dB reference for the return loss bandwidth. At the 15 dB level the return loss bandwidth, for the various values of the compensated impedance, could easily be identified and the best impedance could be determined. Fig. 3.14 is a graph of the 15 dB return loss bandwidth against the compensated impedance ($Z_1$). The value of compensated impedance ($Z_1$) corresponding to the widest 15 dB fractional bandwidth is 41.42Ω.

To study the effect of varying the impedance ratio ($k$) on the isolation bandwidth of the CPD, the 15 dB isolation bandwidth, obtained from simulation, was plotted for
3.6 Results and Discussion

To validate the analyses carried out on the CPD and show the improved 15 dB isolation bandwidth, a prototype was fabricated. The fabricated CPD had $k = 3.4$, and was
3.6. Results and Discussion

Figure 3.15: Isolation bandwidth of CPD vs ratio

The CPD was fabricated on an FR4 substrate with permittivity of 4.55, with thickness of 1 mm. According to the manufacturers datasheet the loss tangent of the substrate is 0.0175 at 1MHz. The dimensions of the CPD are: $l_1 = 6.57$ mm, $l_2 = 6.96$ mm, $l_3 = 6.67$ mm, $W_1 = 2.52$ mm, $W_2 = 0.71$ mm, and $W_3 = 1.86$ mm. The isolating resistor used was an 0805 package 100Ω resistor.

Fig. 3.17 shows return loss curves for the CPD at the input and output ports. The simulated results are plotted using solid lines while the results obtained from measurement are shown with dashed lines.
3.6. Results and Discussion

Figure 3.16: Layout and structure of fabricated CPD

Figure 3.17: Measured and simulated return loss of CPD

ment are plotted using dashed lines. At the operating frequency (6 GHz), the return loss is at 13 dB in simulation and 11.5 dB in measurement. The CPD has a 10 dB fractional return loss bandwidth greater than 150%. Both input and output ports are matched over this entire bandwidth. The transmission coefficients, S21 and S31, are at -3.5 dB instead of -3 dB, indicating that the CPD results in a 0.5 dB of insertion loss.
3.6. Results and Discussion

Figure 3.18: Measured and simulated port isolation of CPD compared with the port isolation of a conventional WPD

Fig. 3.18 shows the isolation between the output ports (S32) of the CPD. The 15 dB bandwidth is 114%, which is more than twice that for the conventional Wilkinson divider. The result confirms the inverse relationship between the return loss at the designed frequency and the isolation bandwidth. If the return loss is relaxed even more, a wider isolation bandwidth can be achieved. There is close agreement between the simulated and measured values.

Table 3.1 compares the proposed CPD with other state-of-the-art designs from the literature. The technique reported in [91] achieves a wide isolation bandwidth of over 17 GHz using multiple matched transmission line sections. However, the technique has the disadvantage of large size and high complexity in design. This disadvantage also applies to the approach reported in [112] where transitions and coupled line sections are employed to improve the isolation bandwidth. The isolation network technique, reported in [99], involves the use of several lumped elements. Unfortunately, it would be challenging to implement this technique at frequencies higher than 3 GHz due to the parasitic effects associated with the lumped elements. Coupled line segments will also be challenging to implement at higher frequencies due to difficulty in realising the optimum spacing between the coupled lines due to the reduced quarter wavelength and manufacturing tolerance. The CPD proposed in this paper provides improved isolation
bandwidth compared to the isolation network and coupled line techniques and it is suitable for UWB applications. Furthermore, the complexity of proposed design is lower than that of all of the other techniques used for bandwidth improvement.

### 3.7 Design Guide for Compensated Power Dividers

In order to redesign the compensated power divider for any application, the following design steps should be taken into consideration.

1. Specifying the operating frequency: For any RF application there is an intended frequency of operation. This will determine the length and width of the transmission lines used in the design.

2. Selection of substrate parameters: The permittivity and thickness of the substrate can be chosen arbitrarily.

3. Minimum return loss of the application: To design an efficient CPD, the minimum return loss acceptable for the application must be known. The target return loss influences the choice of the ratio between \( Z_1 \) and \( Z_2 \). For return loss levels better
than 20 dB, the ratio, $k \leq 2.5$. For return loss better than 15 dB, the ratio, $k \leq 3$. For return loss better than 10 dB, $k \leq 3.5$. For low sensitivity applications such as mobile phone antennas, $k$ can be greater than 3.5.

4. Optimising the isolation bandwidth: To achieve a wider 15 dB isolation bandwidth, the value of the isolating resistor is reduced from the optimum value calculated using equation (3.11). This exact value is obtained by parametric study in simulation using the value of the isolating resistor as the variable.

Following the steps listed above, the CPD can be redesigned at any frequency and for any application of interest.

3.8 Summary

A wide isolation bandwidth between the outputs of a power divider is necessary for wideband antenna applications to reduce the effect of mutual coupling between antenna elements. This chapter surveyed the various techniques used in achieving wide isolation bandwidth in power dividers and proposed a compensated power divider which can be used to improve the isolation bandwidth without increasing the complexity of the design. Odd and even mode analyses were used to derive a relationship between the compensated impedance ($Z_1$) and the arm impedance ($Z_2$). It was shown that this relationship controls the isolation bandwidth between the output ports. A wide isolation bandwidth CPD with 100% 15 dB isolation bandwidth was designed, fabricated and measured to validate the analyses. However, the wide isolation bandwidth is achieved at the cost of a poorer return loss at the operating frequency. The proposed power divider will be useful for ultra-wide bandwidth application and massive MIMO to achieve 15 dB isolation between the antenna elements.
Chapter 4

Passive Amplitude Steering Antenna

Several beam steering techniques were reviewed in Section 2.2. These techniques achieve beam steering by altering some properties of the antenna and/or by using external components. The amplitude steering technique was discussed in Subsection 2.2.10. This approach made use of variable gain amplifiers (VGAs) to actively alter the amplitude of the signal fed to the elements in an array. The use of VGAs for amplitude steering is not common due to: 1) the instability in behaviour of actively loaded antennas, 2) degradation in the SNR performance of the system, 3) reduction in radiated efficiency and gain during beam steering, and 4) effect of the biasing circuitry on the radiation pattern of the antenna.

This chapter presents a novel concept of implementing amplitude steering in an antenna system by using a passive control device. The amplitude of the signals fed to the driven elements in the array is altered by means of a Reconfigurable Ratio Power Divider (RRPD). Section 4.1 presents the concept of amplitude steering using vector diagrams. The main approaches to achieving amplitude control is also presented in this section. Section 4.2 presents an overview of RRPDs and highlights the approaches used in achieving reconfigurable ratios in RRPDs. Section 4.3 presents design and analyses of the RRPD showing the return loss performance of the ports as the ratio is reconfigured. Section 4.4 presents measurement results and discusses the limitations in
the proposed RRPD. Section 4.5 presents the design of a 2×1 proof-of-concept antenna array to validate the concept of passive amplitude steering. Section 4.6 highlights how the change in signal amplitude, supplied to each element, affects the surface current distribution on each antenna element. Section 4.7 presents simulation and measurement results to validate the analyses. Section 4.8 discusses the behaviour of the amplitude steerable antenna when the number of array elements is increased. Emphases will be placed on the maximum steering angle that can be achieved as the size of the array is increased. Section 4.9 presents a brief comparison with the state-of-the-art to highlight the advantages of the novel technique. A summary of the chapter will be presented in section 4.10.

4.1 Concept of amplitude steering

The radiation pattern of an antenna can be considered as a vector, having a phase which can be pictured as the main beam direction and magnitude which can be pictured as the peak gain. When multiple vectors act on an object, a resultant vector can be computed using the properties of all individual vectors. Fig. 4.1 gives a simple illustration of resolving vectors.

![Vector illustration](image)

Figure 4.1: Vector illustration

In this example \( \vec{w}_1 \) and \( \vec{w}_2 \) are the individual vectors and \( \vec{W} \) is the resultant vector. Note that, in this figure, the magnitude of a vector is signified by the length of the line.
whilst the direction of the vector is signified by the orientation of the line. Hence, the magnitude and phase of the resultant vector depends on the magnitude and phase of the individual vectors. This implies that if the phase of the individual vectors is maintained constant, while altering their magnitude, the direction of the resultant vector can be controlled. However, it comes at the cost of a reduction in magnitude of the resultant. This is illustrated in Fig. 4.2

In Fig. 4.2, the phase of the individual vectors is maintained constant; however, the magnitude of vector $\hat{\mathbf{w}}_2$ is varied. On inspection of the figures it is clear that this changes the phase and magnitude of the resultant $\hat{\mathbf{W}}$. In an array antenna this would have the effect or steering the main beam. The proximity of the individual vectors to each other determines the extent to which the change in the magnitude and phase affects the resultant. The closer the vectors, the more effective the change in phase and magnitude of each individual vector has on the resultant.

The conventional approach for steering the main beam of an array antenna involves using phase shifters to change the phase of the currents supplied to each element of the array. A lot is known about steering the main beam of an array antenna by altering the phases of the signals supplied to the elements using phase shifters. But relatively little has been reported about steering the main beam of an antenna by changing the amplitudes of the signals supplied to the elements (so called amplitude steering).

There are two main approaches for varying the amplitude of the signals supplied to individual antenna element: 1) active control using a Variable Gain Amplifier (VGA),
4.2 Overview of Reconfigurable Ratio Power Divider (RRPD)

Reconfigurable ratio power dividers are a type of unequal power divider with the ability to alter the power dividing ratios on their output ports. They have applications in: power amplifiers (PA), and electronically steerable subsystems including reconfigurable antennas for radar and satellite use [118][119][120][121]. The RRPD has been around since the mid-20th century and a detailed review of the state-of-the-art and operating principles of RRPD designs available was presented in [122]. The main approaches that have been used to reconfigure the power dividing ratio of a RRPD are:

1. Phase shifter-based approach
2. Switch-based approach and
3. Varactor-based approach

These devices (phase shifter, switch, and varactor) can alter their states (or values) and hence they form the main components used in reconfiguration.

4.2.1 Phase shifter-Based Approach

The first RRPD design made use of two hybrid couplers and a phase shifter [122]. Fig. 4.3 presents an illustration of the phase shifter based approach. A hybrid coupler is a four-port device with one input port (port 1) and two output ports. There is a 90° phase difference between the two output ports (i.e. ports 2 and 3), and an isolated port (i.e. port 4). The coupler can act as either a power divider or combiner. To implement the RRPD, one of the output ports of the coupler were connected to the phase shifter.
The output of the phase shifter is then connected to the isolated port of the second coupler. The second output of the first coupler is connected to the input of the second coupler. By altering the phase using the phase shifter, the amplitude ratio between the outputs at a specific frequency can be controlled. The output from one arm of the second coupler will always remain the same while the other output is controlled by the phase shifter. The ratios that can be achieved is dependent on the type of phase shifter used. Due to limitations in the design of phase shifters at that time, the reconfigurable ratios that could be achieved were 1:1 and 1:0 [122].

In [120] the authors designed an RRPD using two hybrid couplers and a programmable reflection type phase shifter with five states. The five phase states of phase shifter were $0^\circ$, $53^\circ$, $90^\circ$, $127^\circ$ and $180^\circ$ which produced the following discrete dividing ratios; 1:0, 4:1, 1:1, 1:4 and 0:1 respectively. The RRPD had a good return loss value better than 30 dB for all the states of the phase shifter but the frequency of minimum return loss shifted by over 100 MHz, from 11.7-11.8 GHz as the state of the phase shifter was altered. The authors did not report the insertion loss incurred which is expected to be high due to the use of phase shifters. In [123] the same authors designed a similar RRPD having five discrete dividing ratios. This was achieved by making use of a reflection-line MEMS phase shifter and eight MEMS switches. The dividing ratios achieved were 1:0, 6:1, 1:1, 1:6 and 0:1. The return loss of this design was better than 17 dB which is worse than their design.

In [124], the authors designed a dual band RRPD. The design incorporated two diplexers, two hybrid couplers, and two tuneable phase shifters. The phase shifters provided
phase shifts of 0-90° at each operating band and combined to give seven phase shift angles. The phase shift values applied were 0°, 35°, 53°, 90°, 127°, 145°, and 180° which produced dividing ratios of 0:1, 1:10, 1:4, 1:1, 4:1, 10:1, and 1:0 respectively. An insertion loss of 2 dB was reported along with a worst-case return loss of 15 dB. The 1:10 ratio reported in this work is the highest ratio reported for RRPDs when using a phase shifter and dual hybrid coupler method.

The phase shifter and hybrid coupler method achieves discrete values of dividing ratios. The main drawback of this technique is the high insertion losses incurred within the phase shifter. The design also has high complexity due to the integration of the biasing lines for the phase shifter.

### 4.2.2 Switch-Based Approach

![Illustration of Switch-based technique with coupled lines](image)

Another approach used in achieving reconfigurable ratios in power dividers employs the use of switches to alter the impedance of a line segment such as coupled lines and impedance circuits. The general idea is to design a circuit within the power divider
circuit that could be operated by changing the state of the switches (Open or Short) and thereby controlling the ratio of the power supplied to the output. A schematic of this technique is shown in Fig. 4.4

A number of authors have achieved different RRPDs using this technique. In [125], the authors achieved two power dividing ratios, 1:1 (equal split) and 1:0 (shutting off an output port) making use of two coupled line sections and five switches. When all the switches are turned off, the device behaves like a conventional Wilkinson power divider with equal split of power between the output ports. In order to change the dividing ratio, four out of the five switches should be turned ON at the same time. The designed showed a good return loss performance better than 40 dB for both states. The insertion loss was 0.21 dB at the worst-case. The main drawback of the design is the range of dividing ratios it could achieve. Additionally, the complexity of the design is high due to the number of switches needed to achieve reconfiguration.

In [126], the authors designed and implemented an RRPD with compact size for use in high efficiency power amplifiers. The design had two coupled line sections and six switches. The switches are used to control the characteristics of the coupled line sections to improve the return and insertion loss performance. The divider was designed with an unequal ratio of 1:2 and by means of the switched coupled lines, it could also achieve 1:0 ratio. The RRPD had a worst-case return loss of 21 dB, however, the frequency at which the return loss was reported was altered while reconfiguring the ratios. The reported insertion loss was 0.9 dB for the 1:0 operation ratio. The design had a high degree of complexity in operation due to the several switch states required to reconfigure the ratio of the power divider. The design could only achieve two reconfigurable ratios; this is a huge drawback of the design.

An RRPD operating with two ratios, 1:1 and 1:0, was reported in [119]. The designed RRPD had one coupled line section and four switches. The return loss was better than 17 dB and the reported insertion loss was 0.3 dB. The complexity of this design was lower than that of previously mentioned designs based on coupled lines. However, the limited range of power division ratios is a major drawback of the design. In [127], the authors designed a 1:3 RRPD using two coupled line sections and four switches.
The RRPD had two operating ratios, 1:3 and 1:0. The worst-case return loss of the RRPD was 16 dB when operating with ratio 1:3 and the reported insertion loss was 0.8 dB. The complexity of the design was high due to the use of slotlines to achieve high dividing ratios and the several operating states of the switches.

In [118], the authors designed an RRPD using two coupled line sections and two switches. The reconfigurable ratios achieved were 1:0 and 0:1. The RRPD redirected the input power supplied between the two output ports, one port at a time. The return loss was better than 16 dB at the designed frequency and the insertion loss incurred was not reported. The design was relatively simple compared to the previously reviewed designs but it still had the limitation of having only two reconfigurable ratios.

In [128], the authors designed an RRPD with three operating ratios, 1:0, 0:1, and 1:1. The design had a two meandered coupled line sections and six switches. The measured return loss was better than 15 dB for the worst-case and an insertion loss of 1 dB was reported. The design had more operating ratios when compared with other designs using coupled lines and switches. The design has a high degree of complexity due to the number of switches required and the biasing circuitry needed for each switch.

In [129], the authors achieved reconfiguration between four output ports by switching between impedance circuits incorporated into the device. The RRPD had the ability to alter the number of output ports that it could support. It achieved 1:1 ratio for one output port, 1:2 ratio for two output ports, 1:3 ratio for three output ports, and 1:4 ratio for four output ports. Reconfiguration is only done when the number of ports is either increased or decreased. The design had seven switches and had a return loss better than 20 dB for all the dividing ratios irrespective of the number of output ports used. The worst-case insertion loss of the design was 1.9 dB when operating in 1:1 ratio. The RRPD was complex both in design and operation as a particular combination of switch states are needed to achieve the various reconfigurable ratios.

In [130], the authors used composite left/right handed transmission lines (CLRH TL) to design a RRPD incorporating MEMs switches. By operating the switches connected to the CLRH TL, they were able to control the properties of the CLRH TL and in turn reconfigure the power divider. The design yield the following discrete dividing ratios:
1:1, 1:1.5, and 1:2. It is estimated that more than fifteen switches were required to implement the design. This estimate is based on counting the switches in the figures of the paper as the authors did not give the total number used in their design. The reported return loss was better than 20 dB and the resonant frequency was maintained over the reconfigurable ratio. The RRPD occupied a large area and was very complex in design and operation.

The principal limitations of the switch-based approach based on the use of coupled lines is the limited number of operating ratios and the low dividing ratios that can be achieved. To achieve higher power dividing ratios, the width of the high-power arm of the divider would need to be very wide which makes the implementation of coupled lines difficult [105]. Also, to achieve more operating ratios, the complexity of the design will increase.

### 4.2.3 Varactor-Based Approach

![Illustration of varactor-based approach with four varactors](image)

Figure 4.5: Illustration of varactor-based approach with four varactors

Recently tuneable impedance components such as varactor diodes have been used to implement continuous ratio reconfiguration in power dividers. These devices are used in conjunction with either hybrid couplers or unequal Wilkinson power divider structures in order to create a RRPD. Fig. 4.5 illustrates the varactor based technique used in
4.2. Overview of Reconfigurable Ratio Power Divider (RRPD)

conjunction with unequal Wilkinson power dividers. In [131], the authors used varactor diodes and diplexers, serving as tuneable discontinuities, to achieve reconfigurable ratios with two hybrid couplers. By controlling the bias voltages of the two varactor diodes, any value of dividing ratios from 1:0 to 0:1 can be achieved. This gives a very wide range of tuneable ratios. The return loss of the RRPD was better than 15 dB across the operating band and the worst-case insertion loss was 3 dB. The drawback of the design was its large size due to the use of two hybrid couplers and an unstable resonant frequency. The resonant frequency of the RRPD is altered as the varactor diodes are tuned. The design had a moderate complexity due to the need for biasing circuitry associated with the two varactor diodes.

In [132], the authors designed a RRPD having a dividing ratio that ranged from 1:3 to 1:10.4. The design was based on a Wilkinson unequal power divider with ratio of 1:6. A Defected Ground Plane (DGS) was placed behind the lower power arm and two varactor diodes were used to tune the impedance of the DGS section. The return loss was better than 20 dB for all the values of the biasing voltage but the insertion loss incurred was not reported. The resonant frequency of the RRPD was altered when the biasing voltage is changed. The designed RRPD is compact as it is simply an unequal Wilkinson power divider, however it has high complexity as it involves a DGS section etched on the ground plane and biasing circuitry needed to tune the two varactor diodes. In [133], the authors designed a RRPD having a dividing ratio ranging from 1:1 to 1:4.9. The RRPD had four varactor diodes and a return loss of better than 20 dB at the designed frequency of 1 GHz. The reported insertion loss incurred in the design was 0.6 dB across the tuning range of the varactor diodes. The large number of varactor diodes used increased the complexity of the design.

In [134], the authors designed an RRPD that achieves a power dividing range from 1:1 to 1:3.4 with the use of two varactor diodes. The design was based on a conventional Wilkinson unequal power divider and had a return loss better than 15 dB at the designed frequency. The reported insertion loss incurred was 0.6 dB. The designed RRPD had a compact size and a moderate complexity due to the integration of biasing lines for the two varactor diodes. In [135], the authors designed an RRPD capable of reconfiguring its power dividing ratio from 3.5:1 to 1:3.5. The design made use of
four varactor diodes and was based on the conventional Wilkinson power divider. The reported return loss was better than 11 dB over the tuning range of the varactors and the worst-case insertion loss was 2 dB. The design RRPD compact in size, but had a high complexity in design and operation due to the four varactor diodes used. The insertion loss incurred is also very high at 2 dB.

In [136], a 1:5 RRPD was designed. three PIN diodes were used with different biasing voltages to achieve a 2-mode operation. The first mode of operation is achieved when the power is delivered to one of the output ports while the other is isolated, signifying a ratio of 1:0. In the second mode, the divider operates with ratio 1:5. The worst-case return loss reported was better than 15 dB and the insertion loss was 1.6 dB for the worst-case. A major drawback of the reported RRPD was that it had just one ratio when the two output ports were in use and could not reconfigure the power between the two ports. The authors inserted two switched elements with matching networks (SEMN) and a reconfigurable complex impedance transformer network (RCITN). Pin diodes were used to control the SEMNs and the RCITN which in turn controlled the power dividing ratios. The use of the SEMN and RCITN blocks made the design complicated. In [137], the authors designed an RRPD base on the hybrid branched-line coupler and four varactor diodes in simulation using ADS. The varactor diodes were operated in pair, i.e. two varactor diodes had the same biasing voltage. By controlling the biasing voltage of the varactor diodes, a reconfigurable ratio ranging from 1:2 to 1:4 was achieved. The reported return loss was better than 20 dB over entire range of power division ratios, however the insertion loss of the device was not reported. A major drawback of the design was that the outputs were 90° out of phase and would not be suitable for applications that require in-phase signals. The complexity of the design was not as high when compared with designs requiring SEMN and RCITN blocks.

Some other authors have reported a 3-way reconfigurable power division using varactors [138][139]. Pin Diodes were used with nine SEMN and four RCITN blocks, each block had an associated pin diode. The RRPD had the ability to alter the number of output ports it could support. Based on the bias voltage used, a 1-way, 2-way or 3-way transmission can be achieved. For the 1-way mode, bulk of the input power is transmitted to any of the three outputs depending on the biasing voltage applied to the
4.2. Overview of Reconfigurable Ratio Power Divider (RRPD)

SEMN and RCITN blocks within the RRPD while isolating the other two ports. For the 2-way mode, the RRPD divides the power between any two outputs while isolating the third. For the 3-way mode, the RRPD divides the power equally between the three ports. Reconfiguration is only achieved when the number of output ports is altered. The design in [138] made use of 13 diodes while the design in [139] had six diodes. The insertion loss reported in [138] for the 1-way, 2-way and 3-way modes were 1.13 dB, 0.93 dB and 0.74 dB respectively. A major drawback in this approach is that the reconfigurable ratios are only accurate at the centre frequency. Beyond this frequency, a sharp drop in the power delivered to the outputs is noticed. Also, due to the number of SEMN and RCITN blocks used, the complexity of the RRPD obtained are high in both design and operation.

The varactor based approach is the most common approach use presently. Its main drawback is the complexity in design and operation based on the number of varactors used. Design biasing circuitry needed for the operation could be a major challenge when space and external power supply are limited.

4.2.4 Summary of approaches used in RRPD

Table 4.1 presents a comparison of various works on RRPD based on the three approaches.

The phase shifter-based approach could be used to achieve high dividing ratios, but due to the high insertion losses associated phase shifters, integrating this approach degrades the performance of the RRPD and increases the size of the device. The switch-based approach suffers from low range of reconfigurable ratios and few states of operation. To increase the number of states, the design becomes more complex to incorporate more switches which also implies that the insertion losses will increase. The varactor-based approach gives a continuous means of reconfiguration and it has a wide range of reconfigurable ratios. All these techniques need the used of biasing circuitry to operate the switches or varactors, this increases the complexity of the design. The instability of the resonant frequency when using the phase shifter and varactor-based approaches will pose a challenge for the use of RRPDs for beam steerable applications over narrow
bandwidths. Hence, an approach that achieves a high range of reconfigurable ratio, with reduced complexity and a stable resonant frequency over the reconfigurable range of the RRPD is needed for applications such as narrowband beam steerable antennas.

A novel RRPD which makes use of a potentiometer to reconfigure the ratio between the dividing arms was designed in this work. The design equations and analyses of the RRPD are presented in the next section.

### 4.3 Design and Analyses of RRPD

A key feature of the proposed RRPD, which is based on a conventional WPD with unequal ratios, is that the isolation resistance is replaced with a variable resistance \( R_{\text{var}} \). A compensated impedance \( Z_1 \) is introduced in order to allow more control over the impedance of the dividing arms. Fig. 4.6 shows a schematic of the RRPD.
4.3. Design and Analyses of RRPD

The impedance of the dividing arms $Z_2$ and $Z_3$ can be calculated using the equations for Wilkinson unequal divider [1].

$$Z_2 = Z_0 \sqrt{\frac{1 + k^2}{k^3}}$$  \hspace{1cm} (4.1)

$$Z_3 = k^2 Z_2$$  \hspace{1cm} (4.2)

where $Z_0$ is the characteristic impedance, $k^2$ is the dividing ratio, $Z_2$ is the impedance of the arm carrying a higher power signal, and $Z_3$ is the impedance of the arm carrying a lower power signal. The dividing ratio of a conventional microstrip unequal WPD is limited to 1:3 [140]. This is due to the use of a very thin (typically 0.2 mm), high impedance transmission line, for the lower power arm. Use of the compensated impedance ($Z_1$) allows this ratio to be improved slightly when $Z_1$ is chosen to be less than $Z_0(50\Omega)$. Eqn. (4.3) can be used to calculate the impedance of the arms incorporating the compensated impedance,

$$Z_2 = Z_1 \sqrt{\frac{2}{k^2}}$$  \hspace{1cm} (4.3)

where $k^2$ is the power dividing ratio. $Z_3$ can then be calculated using the expression in eqn. (4.2). Eqn. (4.3) gives an approximation of eqn. (4.1) for low unequal ratios.

The variable resistance ($R_{\text{var}}$) in Fig. 4.6 is vital to achieve ratio reconfiguration. With a low value of $R_{\text{var}}$ (i.e. $0\Omega$), the potential difference between the dividing arms will
be zero (short circuit), hence the power in both arms will be the same. $R_{var}$ sets the potential difference between the dividing arms thereby controlling the ratio between the dividing arms. This is the main difference between the novel design and the conventional unequal WPD. To illustrate the reconfigurable properties of the RRPD, the proposed design was modelled in ADS and shown in Fig. 4.7.

![Figure 4.7: Model of reconfigurable ratio power divider in ADS](image)

To study the performance of the RRPD with different values of $R_{var}$, three models of the RRPD with different dividing ratios ($k^2$) were constructed. The chosen ratios were 1:1.5, 1:2, and 1:2.5. For each model, $R_{var}$ was varied from $0 - 20,000 \Omega$. For each value of $R_{var}$, the ratio of $S21:S31$ ($Y$), at the operating frequency was computed and plotted as a function of $R_{var}$. The quantity $Y$ represents the reconfigurable ratio. Fig. 4.8 shows the relationship between the reconfigurable ratio and $R_{var}$.

From the analyses, it was observed that the reconfigurable ratio ($Y$) of the divider for each model has a range of ratios which starts from 1 when $R_{var} = 0 \Omega$ and approaches the square of the dividing ratio (i.e. $(k^2)^2$) when $R_{var}$ approaches infinity. Hence, for the 1:1.5 model, $Y$ had a range of 1 to 2.25, for the 1:2 model, $Y$ had a range of 1 to 4 and for the 1:2.5 model, $Y$ had a range of 1 to 6.25. Based on the relationship between $Y$ and $R_{var}$, an empirical study was undertaken to derive eqn. (4.4) that relates $Y$ to $R_{var}$.

$$Y = \frac{2(a^2 - 1)}{\pi} \tan^{-1}\left(\frac{0.45\pi}{180a} R_{var}\right) + 1$$  \hspace{1cm} (4.4)
Where \( a \) is the dividing ratio \((k^2)\), \( R_{\text{var}} \) is the variable isolation resistance, and \( Y \) is the reconfigurable ratio. The dotted lines in Fig. 4.8 were obtained using eqn. (4.4). It is clear that there is good agreement with the values recorded from ADS.

A major challenge encountered in the implementation of the RRPD is the choice of component used to represent \( R_{\text{var}} \). For a proof-of-concept solution, a standard low frequency film type (CERMET) potentiometer was used to implement \( R_{\text{var}} \). It is well known that lumped elements have unwanted parasitic reactance at higher frequencies [141]. This has discouraged the use of resistors, inductors, capacitors and potentiometers at higher frequencies. To understand the behaviour of these components at high frequencies, it is helpful to have an equivalent circuit of the component.

An equivalent circuit for a film type potentiometer was unavailable at the time of writing. In order to model the behaviour of the potentiometer in ADS, an equivalent circuit for a film type resistor [142] was used. This approach was justified since the film type resistor has a similar package to the potentiometer. In the model for the potentiometer the fixed resistance was replaced with the variable resistance of a potentiometer. Fig. 4.9 shows an approximate equivalent circuit of the potentiometer. However, this circuit has not be experimentally validated.

The following additional parasitics model certain features associated with a potentiometer: 1) the inductance of the wiper arm \((L_{\text{package}})\), 2) capacitance between the
4.3. Design and Analyses of RRPD

Figure 4.9: Approximate equivalent circuit of a film type potentiometer

packaging and the resistive element \((C_{\text{package}})\), 3) the inductance developed in the connecting leads (i.e. the terminals of the component) \((L_{\text{lead}})\), and 4) the capacitance between the body of the lumped element and the ground \((C_{\text{ground}})\). The expression for the impedance seen looking to the terminals of the potentiometer when inserted in a microwave circuit is given as [142]:

\[
Z_{\text{var}} = R_{\text{var}} \sqrt{\frac{1 + \left(\frac{\omega L_{\text{package}}}{R_{\text{var}}}\right)^2}{1 + \left(\frac{\omega R_{\text{var}} C_{\text{package}}}{\omega R_{\text{var}} C_{\text{package}}}\right)^2}}
\]  

(4.5)

Where \(\omega = 2\pi f\), \(Z_{\text{var}}\) is the variable impedance, \(R_{\text{var}}\) is the resistance of the potentiometer, \(L_{\text{package}}\) is the inductance of the wiper arm, and \(C_{\text{package}}\) is the capacitance induced by the packaging. For the purpose of studying the impedance behaviour of the potentiometer at 2.4 GHz, \(R_{\text{var}}\) was varied from 0 to 5\(k\Omega\) and plotted against \(Z_{\text{var}}\) using equation (4.5).

The values for \(L_{\text{package}}\) and \(C_{\text{package}}\) were obtained from manufacturer datasheet for a similar package film type resistor [142] and were \(C_{\text{package}} = 35\text{fF}\) and \(L_{\text{package}} = 1\text{nH}\). Fig. 4.10 shows the impedance \((Z_{\text{var}})\) of the potentiometer, at a frequency of 2.4 GHz, as a function of \(R_{\text{var}}\). The potentiometer no longer behaves in a linear manner as is expected at lower frequencies. From Fig. 4.10, the maximum impedance value \((\text{Max. Val})\) of the potentiometer is determined by the packaging capacitance \((C_{\text{package}})\).
and minimum impedance value (Min. Val) of the potentiometer is determined by the inductance of the wiper arm ($L_{package}$). A reconfigurable impedance, $Z_{var}$, is achieved over the tuning range of $R_{var}$. The shape of the curve for $Z_{var}$ in Fig. 4.10 changes as frequency increases. The parasitic impedance of $L_{package}$ ($X_L$) increases with frequency while that of $C_{package}$ ($X_C$) reduces with increasing frequency. This implies that as frequency increases, the $Z_{var}$ curve will gradually flatten and then reverses as $X_L$ becomes greater than $X_C$ with further increase in frequency. This gives an idea of the behaviour of the potentiometer at 2.4 GHz and it shows that the potentiometer can be used as a reconfigurable impedance.

4.4 Results and Discussion of the designed RRPD

The RRPD was designed with a fixed dividing ratio of 1:2.5 (i.e. $k^2 = 2.5$). The compensated impedance ($Z_1$) was set to 60Ω. Applying eqn. (4.3), $Z_2$ and $Z_3$ are 53.67Ω and 134.16Ω respectively. The substrate used was FR4 with dielectric constant of 4.55, thickness of 1.6 mm, and loss tangent of 0.0175 at 1 MHz. These values were obtained from the manufacturers datasheet. A BOURNS 3386 1kΩ potentiometer was used. The structure of the RRPD is shown in Fig. 4.11.
The dimensions of the RRPD are: $l_1 = 35.6mm$, $l_2 = 13.55mm$, $l_3 = 20mm$, $l_4 = 32.2mm$, $l_5 = 17.02mm$, $l_6 = 17.22mm$, $w_1 = 2.96mm$, $w_2 = 2.62mm$, $w_3 = 0.24mm$, and $w_4 = 2.14mm$. The potentiometer was modelled, in simulation, using two lumped elements, namely a capacitor in parallel with a series resistor and inductor.

Fig. 4.12 shows the return loss performance of the RRPD. The RRPD is resonant at 2.4 GHz with return loss of 19 dB. When the value of the potentiometer is set to the minimum (Min. $Z$), the return loss attains its best value. The resonant frequency is maintained at 2.4 GHz for all potentiometer settings. However, the return loss becomes slightly worse as the value of $Z_{var}$ increase. There is a difference of 4 dB between the return loss values obtained, through simulation and measurement, for Max. $Z$ (at 2.4 GHz). This discrepancy can be attributed to the difference in the actual and modelled values of the capacitance and inductance in the potentiometer.

The potentiometer in this design plays three major roles: 1) it reconfigures the power dividing ratio between the output arms, 2) matches the output ports at the designed frequency, and 3) it provides isolation between the output ports. Fig. 4.13 shows curves representing the transmission coefficients (S21 and S31), of the RRPD, as a function of frequency. Five pairs of curves are plotted corresponding to different $Z_{var}$ values.

The insertion loss of the RRPD can be estimated from the comparison the input power level and the sum of the output power levels at port 2 and 3. This is shown in Fig.
4.4. Results and Discussion of the designed RRPD

Figure 4.12: Measured and simulated return loss of RRPD for maximum and minimum potentiometer values.

Figure 4.13: Simulated transmission coefficients of designed RRPD

4.14 and at 2.4 GHz, the insertion loss is 0.94 dB and 1.05 dB for Min. Z and Max. Z respectively.

Five different potentiometer settings were employed, namely: Min. Z, IP 1, IP 2, IP 3, and Max. Z. This had the effect of altering the power dividing ratio. With the
4.4. Results and Discussion of the designed RRPD

Figure 4.14: Total insertion loss of RRPD (Input power level compared with output power levels)

potentiometer set to Min. Z, the ratio S21:S31 at 2.4 GHz is slightly above 1 instead of 1 when eqn. (4.4) is used. This is due to parasitic inductance of the potentiometer. The power dividing ratio is 1:6 when the value of the potentiometer is set to Max. Z. The potentiometer has the effect of reconfiguring the power dividing ratio and of altering the matching on the output ports, as mentioned earlier. Figs. 4.15 and 4.16 show the measured return loss of the output port (i.e. ports 2 and 3) of the RRPD for each of the five potentiometer settings.

Considering Fig. 4.15, the impedance matching on the port carrying the higher power (i.e. Port 2) improves as the value of the potentiometer ($Z_{var}$) increases. Port 2 is resonant at 2 GHz; however, it has a wide 10 dB bandwidth that covers the intended 2.4 GHz band.

A different effect is noticed, in Fig. 4.16, on the port carrying the lower power (i.e. port 3). The lower power port is not matched at the intended frequency for the minimum value of the potentiometer (Min. Z). But the match improves as $Z_{var}$ is adjusted to IP 1 and IP 2. Further increase in $Z_{var}$, beyond IP 2 to Max. Z, has a negative effect on the matching at port 3. At the Max. Z, port 3 is no longer matched at 2.4 GHz.

Fig. 4.17 shows the measured isolation between the output ports. For $Z_{var} = $ Min. Z,
4.4. Results and Discussion of the designed RRPD

Figure 4.15: Measured return loss for the High power port (port 2) of RRPD

Figure 4.16: Measured return loss for the lower power port (port 3) of RRPD

the isolation between the ports is better than 12 dB. Increasing $Z_{var}$ to IP 1 and IP 2 improves the isolation between the output ports. When $Z_{var}$ is further increased to Max. Z, the isolation deteriorates. In practice the value of the potentiometer will need to be set to a value that not only gives the required dividing ratio but also ensures that
4.4. Results and Discussion of the designed RRPD

Figure 4.17: Measured port isolation (S32) of RRPD

the output ports are always matched and well isolated at the designed frequency.

Table 3.2 shows a comparison of the designed RRPD with state-of-the-art designs.

Table 4.2: Comparison with state-of-the-art

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</tr>
<tr>
<td></td>
<td></td>
<td>1:1</td>
<td>0.94</td>
<td></td>
<td>9 SEMN blocks</td>
</tr>
<tr>
<td></td>
<td></td>
<td>1:1:1</td>
<td>0.74</td>
<td></td>
<td>4 RCTTN blocks</td>
</tr>
<tr>
<td>This work</td>
<td>Potentiometer</td>
<td>1:1-1.6</td>
<td>0.5-0.74</td>
<td>Continuous</td>
<td>1 potentiometer</td>
</tr>
</tbody>
</table>

The designed RRPD has a reduced complexity when compared with other approaches. However, using a potentiometer implies manual reconfiguration which is the main draw-
4.5 Design of Proof-of-concept antenna

In order to evaluate the concept a passively controlled amplitude steerable antenna, a simple $2 \times 1$ antenna array, operating at 2.4 GHz, was designed. The design utilised an FR4 substrate having a permittivity of 4.55, thickness of 1.6 mm, and loss tangent of 0.0175 at 1 MHz. Circular microstrip patches were used as driven elements. The radius of each patch is 16.8 mm and the feed point is positioned 6.25 mm above the centre of the patch along the y-axis. Fig. 4.18 shows the structure of the antenna.

Figure 4.18: Structure of amplitude steerable antenna.

The spacing between the patches is $0.53\lambda_g$ (from centre-to-centre) and $0.03\lambda_g$ (from edge-to-edge). Initially, the amplitudes of the signals supplied to each driven element were altered using the combine result tool within postprocessing environment in CST microwave studio.

The RRPD was designed to fit back-to-back with the antenna. The length of the
substrate used for the RRPD was longer than that of the antenna. Hence, in Fig. 4.18, the ground plane is visible between the edge of the antenna substrate and port 1. The RRPD was built on the same substrate material as the antenna. Fig. 4.19 shows the layout of the RRPD.

![Figure 4.19: Structure of the RRPD (back)](image)

The power dividing ratio (1:2.5) was chosen such that a reconfigurable ratio from 1:1 (equal power division) to about 1:6 (about 90% of the power to one port) could be achieved with an acceptable level of input matching.

### 4.6 Analyses of change in amplitude and surface current

In order to better understand the operation of the antenna the surface current distributions were plotted in CST for different applied amplitude excitations (see Fig. 4.20).

Fig. 4.20a shows the surface current distribution for an equal power dividing ratio between the two-antenna element. The resulting radiation pattern features a main
4.6. Analyses of change in amplitude and surface current

(a) Ratio 1:1
(b) Ratio 2:1
(c) Ratio 3:1
(d) Ratio 6:1

Figure 4.20: Surface current distribution at 2.4GHz

beam directed towards boresight, as shown in Fig. 4.21a. When the amplitude ratio, between the dividing arms changes, the surface current distribution on the antenna also changes. This can be seen in Fig. 4.20b-d. This change in amplitude results in the main beam of the antenna steering towards the element with lower surface current density. This is shown in Fig. 4.21b-d.

The gain of the antenna reduces as it steers away from boresight. This is called scan loss and it is similar with the reduction in magnitude of a resultant vector as described earlier using the vector analogy. The scan loss of the antenna is 0.05 dB/deg. For conventional phased arrays with ideal phase shifters, the estimated scan loss is given as a factor of $\cos(\theta)$ [143], where $\theta$ is the steered angle. For a conventional phased array (using ideal phase shifters) with similar boresight gain, the scan loss will be 0.02 dB/deg. The use of real phase shifters will significantly degrade the scan loss.

The total efficiency of the antenna system also reduces. Fig. 4.22 shows the drop in efficiency in relation to the steered angle.
4.7 Simulation and Measurement results

Fig. 4.23 shows photographs of a hardware prototype for the antenna connected to the reconfigurable ratio power divider. A 1kΩ BOURNS 3386 potentiometer was employed within the hardware prototype.
4.7. Simulation and Measurement results

The potentiometer has a maximum value of $1k\Omega$. When incorporated into the circuit, it becomes a complex impedance due to parasitic effects at high frequencies. The maximum value of the resistance is referred to as Max. $Z$ while the minimum value is Min. $Z$. The return loss ($S_{11}$) of the combined structure is shown in Fig. 4.24. The return loss nulls for the antenna lie at 2.46 GHz (Max. $Z$) and 2.48 GHz (Min. $Z$). The measured return loss shows a wider bandwidth when compared with simulation.
4.7. Simulation and Measurement results

The return loss for Min. Z is better than 15 dB. But as the power division ratio is increased, the return loss degrades slightly. The worst-case return loss, i.e. Max. Z, is better than 13 dB.

The relationship between the potentiometer setting and the power dividing ratio was presented in Section 4.3. Fig. 4.25 shows the x-z plane radiation pattern of the fabricated prototype with the potentiometer turned to the minimum value. This situation corresponds to the case where the surface current density is equally distributed between the two driven elements. The main beam of the antenna, in this situation, is directed towards boresight and the surface current density on the two driven elements is equal. The H-field pattern is 20 dB lower than that of the E-field.

![Radiation pattern for equal power split between the antenna elements.](image)

By increasing resistance of the potentiometer, it is possible to steer the main beam of the array antenna towards the direction of the antenna element on which the surface current density is lower. Fig. 4.26 shows the x-z plane radiation pattern corresponding to Max. Z. The antenna has a maximum steering angle of 21°. The antenna steers continuously as the resistance of the potentiometer is turned from 0Ω to 1kΩ. The H-field is more than 20 dB lower than the E-field. The antenna has a continuous steering range of 0° to 21°. Fig. 4.27 shows the intermediate steering angles achieved. The
4.7. Simulation and Measurement results

Figure 4.26: Radiation pattern for Max. Z value of potentiometer

potentiometer was measured at some points before installation, these points had the following values: $0.26k\Omega$, $0.69k\Omega$ and $0.87k\Omega$. Fig. 4.27 shows the radiation pattern for these points.

Figure 4.27: Various steering angles
The total efficiency of the antenna varies from 55% (for Min. Z) to about 45% (for Max. Z). The poor efficiency is partly due to the lossy nature of the FR4 substrate used at high frequencies and the insertion loss of the reconfigurable ratio power divider (RRPD) which increases for high values of reconfigurable ratios. If the simulation is re-run with a lossless dielectric, the worst-case efficiency for the antenna (i.e. for Max. Z) is 81%. This shows that the poor efficiency is mainly attributable to the lossy nature of the FR4 substrate. To study the effect of the inter-element spacing on the overall radiation pattern of the antenna a number of simulations were performed in which the centre-to-centre spacing between the antenna elements was varied. Fig. 4.28 shows a graph of the maximum steered angle corresponding to each value of centre-to-centre spacing.

![Figure 4.28: Max. steered angle vs. centre-to-centre spacing](image)

From inspection of the graph in Fig. 4.28, it is clear that the maximum steered angle of the antenna reduces as the spacing between the elements increase. This behaviour is similar for a conventional phased array antenna. For small values of $d (< 0.36\lambda_0)$, the main beam steering angle is large due to the strong surface current interaction between the driven elements. With $d (> 0.36\lambda_0)$, the surface current interaction between the elements are much weaker, such that changing the dividing ratio will have negligible
effect on the overall radiation pattern of the antenna. This is similar with vectors acting on an object. The closer they are to the object, the more the effect they have on the object.

The effect of the spacing on the amplitude steerable antenna is similar to that for the Electrically Steerable Passive Array Radiators (ESPAR) and conventional parasitic steerable antennas. The parasitic elements need to be close to the driven element for it to have any effect on the radiation pattern of that element. The key difference between the amplitude steerable antenna and the conventional parasitic steerable antenna is the direction of steering when operated. When a parasitic element is switched ON it acts as a reflector. For this reason, the main beam of a parasitic steerable antenna steers away from the parasitic element that is turned ON, while the amplitude steerable antenna steers towards the element with lower amplitude. This is illustrated in Fig. 4.29.

![Conventional Parasitic Steerable Antenna and Amplitude Steerable Antenna](image)

(a) Conventional Parasitic Steerable Antenna  (b) Amplitude Steerable Antenna

Figure 4.29: Conventional parasitic and amplitude steerable antennas

The results of the measurements agree with simulations; this validates the concept of amplitude steering using RRPDs. The limitation of this design is that it steers towards one direction only (i.e. $0^\circ$ to $+21^\circ$ rather than $\pm 21^\circ$). With additional circuitry, the design could be made to steer towards both directions off boresight and cover $\pm 21^\circ$. In order to achieve this objective, an additional RRPD would be needed. The second RRPD would be mirrored, with respect to the first RRPD, about the feed point of the antennas. Consequently the dividing arms, of the two RRPDs, would connect to different driven elements and switches would be used to isolate one RRPD from the
4.8 Analyses with more driven elements

Extension of the amplitude steering technique to include more driven elements was carried out with the aim of studying the maximum steerable angle as the number of driven elements increases. Fig. 4.30 shows the structure of the expanded array.

![Figure 4.30: Larger amplitude steerable arrays](image)

To increase the number of driven elements, additional elements are integrated along the y-axis. With this setup, the array would steer along the x-z plane of the antenna, as before. It is important to maintain the symmetry of the feed point along the y-axis. If the symmetry is lost, the pattern of the individual elements would not add up constructively when amplitude steering is applied. The amplitudes of the signals, supplied to the driven elements, were varied. This was achieved using combine result tool under postprocessing in CST microwave studio.

Fig. 4.31 shows a plot of the maximum steering that can be achieved when the antenna size is increased. For a $2 \times 2$ antenna array, the maximum steering angle that can be achieved is $23^\circ$. This was the highest recorded value. The maximum steering angle reduces as the array size increases beyond $2 \times 2$ elements. For $2 \times 3$ and $2 \times 4$ antenna arrays, the maximum steering angle is $16^\circ$ and $12^\circ$ respectively.

For larger arrays, such as a $4 \times 4$ array, the amplitude steering has negligible effect on the array radiation pattern. It can therefore be concluded that the amplitude steering
4.9 Comparison with State-Of-The-Art

The amplitude steering technique based on the use of a RRPD has the advantage of enabling continuous beam steering without the use of parasitic elements and phase shifters. Table 4.3 shows the comparison of this approach to the use of VGA for amplitude steering.

<table>
<thead>
<tr>
<th>Work</th>
<th>Steering angles achieved</th>
<th>Steering resolution</th>
<th>Control mechanism</th>
<th>IL (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>[74]</td>
<td>±30°</td>
<td>Continuous</td>
<td>VGA</td>
<td>-</td>
</tr>
<tr>
<td>This work</td>
<td>0-21°</td>
<td>Continuous</td>
<td>RRPD</td>
<td>0.5-0.74</td>
</tr>
</tbody>
</table>

The use of VGAs is not common due to the drawbacks that have been highlighted earlier.
in the chapter. The approach compare favourable with parasitically steered antennas. Most parasitically steered antennas make use of tuneable reactances or switches to change the states of the parasitics. This increases the complexity and size of the antenna. The insertion loss of the RRPD is significantly lower than that of phase shifters used in conventional phased arrays as can be seen in Fig.4.14.

4.10 Summary

A novel approach of passive amplitude steering has been presented in this chapter. A reconfigurable ratio power divider which is controlled by a potentiometer was used to achieve passive amplitude steering. A $2 \times 1$ array was designed and fabricated to validate the proposed concept. The designed antenna had the capability of steering continuously from boresight to $21^\circ$. This approach provides a small scale and cheap alternative for beam steering. The main limitation with the proposed RRPD is the poor matching in the output port of the low power arm when a high reconfigurable ratio is achieved. A possible improvement to this design would be the use of digital potentiometers which would offer a digital control of beam steering rather than the analogue control achieved in this work.
Chapter 5

Parasitic and Phased array steering

Pattern reconfigurable antennas are essential in next generation (5G) communication systems. In Chapter two, a number of beam steering techniques that have been reported in the literature were discussed. In Section 2.3, a need for improvement in the steering resolution of the various reported techniques was identified. Most of the steering techniques such as lens, parasitic, and switched beam yield coarse resolution steering. Phased arrays can achieve fine steering resolution; however, this would require the use of phase shifters having a large number of bits of resolution. The insertion loss of the phase shifter is proportional to the number of bits, the insertion loss of the phase shifter becomes very high and this becomes a major concern with the use of phased arrays.

In this chapter, a novel concept for improving the steering resolution of an antenna that involves combining two beam steering techniques, namely parasitics together with a phased array is introduced. Section 5.1 will illustrate an approach for improving beam steering resolution by combining two known beam steering techniques. Section 5.2 describes the design of the hybrid parasitic antenna. Section 5.3 presents simulation results for the novel antenna. Section 5.4, analyses the effect of altering the number of driven elements within the antenna. The effect of increasing the number of driven elements on the steering resolution will also be discussed. In Section 5.5, the steering
resolution of the hybrid parasitic antenna will be compared with that of a conventional phased array incorporating digital phase shifters having 2, 3, 4, 5 and 6-bits. Section 5.6 provides design guidelines which have been derived for the hybrid parasitic antenna. A summary of the chapter is presented in Section 5.7.

Some of the contents in this chapter have been published in [40].

5.1 Approach

There are two approaches to improve the steering resolution. The first approach is to have a very fine resolution over the entire steering range. This has been the focus of researchers reported to date. Most of the reported work tends to extend the range of steering but does not improve on the steering resolution. The second approach is to utilise a combination of coarse and fine angle steering techniques which will yield fine resolution steering over the entire steering range. Fig. 5.1 illustrates this novel approach to beam steering.

![Figure 5.1: Dual Beam Steering Technique](image)

To achieve coarse steering, phased array steering with a low number of bits (2-3 bits)
of phase shifting resolution is implemented. This would avoid the high insertion losses associated with the use of phase shifters having a large number of bits of phase shifting resolution. For fine steered angles, parasitic elements operated with switches are used. This gives a combined parasitic and phased array method of realising fine beam steering resolution over a large steering range which is controlled by the phase array.

5.2 Structure and Design of Antenna

![Concept of Antenna](image)

Figure 5.2: Concept of antenna

Fig. 5.2 shows the concept of the proposed novel antenna. Conventional parasitic antennas have a single driven element surrounded by parasitic elements. The parasitic elements are connected to the ground plane by means of switches. Switching the parasitic element ON (shorting to ground) and OFF (isolating from ground) results in steering the radiation pattern of the antenna. A conventional phased array simply consists of driven elements separated by a particular spacing, typically $0.5\lambda$. Each of the driven elements is controlled by a phase shifter. Altering the phase of each driven element, in relation to the other driven elements, has the effect of steering the beam.

A major challenge that hinders the use of parasitic elements in conventional arrays is the difficulty of getting the parasitic element to the array centre. This is necessary if they are to have a meaningful effect on the radiation pattern. The parasitic elements have to be close to centre of the array and also close to each of the driven elements.
Hence instead of surrounding the radiating elements by parasitic elements, like the conventional parasitic antenna, the parasitics are inserted between each element of the array. The spacing between the phased array elements is increased slightly in order to accommodate the parasitic elements and prevent them from touching the driven elements.

To design the antenna, the intended frequency was chosen to be 11 GHz, the substrate permittivity and thickness were chosen to be 2.2 and 1.58 mm, respectively. The thickness of the copper cladding was 18 microns. The radiating element within the array is a circular microstrip patch operating in the $TM_{110}$ mode. The radius of the patch was calculated using the formula for the effective area of circular microstrip patch antenna [1]:

$$a_e = \frac{F}{\left\{1 + \frac{2h}{\pi\epsilon_r F} \left[\ln\left(\frac{\pi F}{2h}\right) + 1.7726\right]\right\}^{1/2}}$$  \hspace{1cm} (5.1)

And F is given as:

$$F = \frac{8.791 \times 10^9}{f_r\sqrt{\epsilon_r}}$$  \hspace{1cm} (5.2)

Where $a_e$ is the effective area of the patch, $h$ is the thickness of the substrate in cm, $\epsilon_r$ is the dielectric constant of the substrate, and $f_r$ is the intended resonant frequency of the antenna.

Fig. 5.3 shows the structure of hybrid parasitic antenna. The radius of the patch was calculated to be 4.74 mm. Each radiating element was fed using coaxial port with its feed point offset, along the y-axis, by 2.7 mm from the centre of the patch, as shown in Fig. 5.3. The spacing between the driven elements is $0.53\lambda$ which was obtained by parametric studies carried out to achieve acceptable antenna performance parameters. The co-ordinates for the centre of the parasitic elements, taking the centre of each driven element as the origin, are $(-0.265\lambda, 0.265\lambda)$, $(-0.265\lambda, -0.265\lambda)$, $(0.265\lambda, 0.265\lambda)$, and $(0.265\lambda, -0.265\lambda)$. The radius of each parasitic element is 4.5 mm. The radius of the vias within the parasitic element is 1 mm. A total of eight parasitic elements are used in the design. The minimum spacing between the parasitic and driven elements is 0.71 mm, this is within the fabrication tolerance at the workshop.
From Fig. 5.3, the parasitic elements are arranged in rows above and below the driven elements. The parasitics are operated in pairs, parasitics having the same number are operated simultaneously. For this reason, there are $2^N$ possible combinations of parasitic switch states, where $N$ is the number of parasitic pairs. So in this case there are $2^4 = 16$ states. The structure is symmetrical and hence, some of these combinations are the mirror image of one other along the plane of symmetry (i.e. the Y-axis).

For an antenna with this topology there are three different scenarios for steering the radiation pattern of the antenna:

1. Switching the parasitic elements (so called, parasitic steering)
2. Applying different phase shift to driven elements (Phased array steering)
3. Parasitic and phase array steering

The results for operating scenarios 1 and 3 are presented in the following sections. Operating scenario 2 is not discussed in this work as it is well known and reported in text books on array antennas.

**5.3 Antenna Performance under parasitic steering**

In order to simplify the understanding of the parasitic states, the parasitic combinations are represented, in a 4-bit notation with a number above each pair of parasitics, in Fig.
5.3. Antenna Performance under parasitic steering

5.3 to represent each bit position. 0 is used to denote a parasitic in the OFF (open) state and 1 is used to denote when that parasitic is ON (short). For instance, 1100 would represent parasitic 1 and 2 are ON and parasitic 3 and 4 are OFF while 0011 would represent parasitic 1 and 2 are OFF, parasitic 3 and 4 are ON.

![Figure 5.4: Simulated return loss of designed antenna](image)

Fig. 5.4 shows the return loss of the designed antenna when all the parasitics are either switched ON or OFF. With the parasitics all switched off (0000), the antenna resonates at 11.7 GHz with a 10 dB operating bandwidth of 4.7%. With all parasitics turned ON (1111), the resonant frequency shifts to 12.12 GHz which is due to the mutual coupling that develops between the radiating elements and the parasitics. This can be seen as a disadvantage of using parasitics. However, the 10 dB operating bandwidth of the antenna improves to 13.76% which is about three times the operating bandwidth of switch state 0000 and covers our intended operating frequency of 11.7 GHz.

Fig. 5.5 shows the radiation pattern for the antenna. The antenna shows better realised gain and side lobe levels with the switches turned ON, in comparison to the case when the switches are turned OFF. However, the efficiency of the antenna drops when the switches are turned ON. This is attributed to the losses incurred within the switches. A summary of the gain and side lobe levels is presented in Table 5.1.
5.3. Antenna Performance under parasitic steering

Figure 5.5: Radiation pattern of the Designed Antenna at 11.7 GHz.

Table 5.1: Summary of Antenna performance

<table>
<thead>
<tr>
<th>Switch State</th>
<th>Realised Gain (dBi)</th>
<th>Side lobe level (dB)</th>
<th>Efficiency</th>
</tr>
</thead>
<tbody>
<tr>
<td>All ON</td>
<td>11.3</td>
<td>-16.5</td>
<td>77.8%</td>
</tr>
<tr>
<td>All OFF</td>
<td>5.65</td>
<td>-9.2</td>
<td>82.6%</td>
</tr>
</tbody>
</table>

When other combinations of switches are use, the return loss of each radiating element is altered. The fractional bandwidth of each driven element increases or decreases around the designed frequency of 11.7 GHz depending on the state of the switches around them. When the driven elements are surrounded by parasitics that are turned ON the antenna exhibits a wider frequency operating bandwidth. Conversely the operating bandwidth is narrower when the parasitics around them are turned OFF. Hence, the parasitics can be used to improve the bandwidth of microstrip antennas [144]. Fig. 5.6 shows the reflection coefficients of each driven element for different switch combinations with the minimum achievable instantaneous operating bandwidth highlighted by the shaded box. Table 5.2 shows a summary of the switch combination and the beam steering angles that are achieved for each switch combination.
### Table 5.2: Parasitic combination and steered angles

<table>
<thead>
<tr>
<th>Combination</th>
<th>1000</th>
<th>1100</th>
<th>1110</th>
<th>1010</th>
<th>0100</th>
<th>1101</th>
<th>1001</th>
<th>0110</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>0001</td>
<td>0011</td>
<td>0111</td>
<td>0101</td>
<td>0010</td>
<td>1011</td>
<td>0101</td>
<td>1110</td>
</tr>
<tr>
<td>Achieved beam angle</td>
<td>$\pm 5^\circ$</td>
<td>$\pm 10^\circ$</td>
<td>0$^\circ$</td>
<td>$\pm 2^\circ$</td>
<td>$\pm 15^\circ$</td>
<td>$\pm 3^\circ$</td>
<td>0$^\circ$</td>
<td>0$^\circ$</td>
</tr>
</tbody>
</table>

![Figure 5.6: Return loss of various switch combinations](image-url)
Fig. 5.7 shows the radiation patterns for the maximum steered angle achieve through parasitic steering.

Figure 5.7: Maximum steered angle of HPAA using parasitic steering.

5.4 Antenna performance under parasitic and phased array steering

The main purpose of integrating parasitic elements into a conventional array is to achieve a dual means of steering. The phased array is used to steer the beam in large angle steps while the parasitics are used to steer the beam in smaller angle steps. To implement phased array steering in simulation, progressive phase shifts are applied to the signal at each port of the driven element, i.e. 0 for the first element, $X^\circ$ for the second element, and $2X^\circ$ for the third element, where $X$ is the desired phase shift. In Fig. 5.8, the main beam of the antenna has been steered coarsely to $15^\circ$ away from boresight by applying a progressive phase shift of $45^\circ$, between the driven elements.

Combining the parasitic switch state 0011 with phased array steering achieved using the progressive phase shifts between the driven elements, the beam is steered to $22^\circ$. 
5.4. Antenna performance under parasitic and phased array steering

Figure 5.8: Parasitic and phased array steering

This is an addition of 7° steering compared to the use of phased array steering alone. Using switch state 1100 to the phased array steering, the beam is steered to 7°. The combinations (1100 and 0011) gives the maximum steered angles away from the direction that would be achieved using phased array steering alone. Finer steps of steering are achieved when other switch combinations are used. The steering range of parasitic technique for each phased array steered angle was studied along with the steering range of the phased array. The results are presented in Fig. 5.9.

The maximum and minimum steering angles for each combination were recorded and plotted against the case of full phased array steering obtained with all the parasitics turned on (i.e. state 1111). The shaded region, around the array steering curve, represents the range of fine steering that can be achieved. Consider the situation where the antennas main beam is steered towards the positive direction by means of the phased array. Under this situation, the parasitic combination that yielded the highest steering angle before phased array steering was applied (i.e. 0100) no longer produces as much steering. However, the reverse combination (i.e. 0010) which yields the minimum steering away from boresight tends to produce more steering towards boresight. The efficacy of the parasitic steering approach reduces as the steered angle, achieved using the phased array approach, nears the upper limit on the maximum steering angle. This
5.5. Effect of the number of driven elements on antenna performance

The performance of the novel antenna incorporating three driven elements have been examined. This antenna topology can be employed with any integer number of driven elements. To analyse the effect of increasing or decreasing the number of driven elements, the following will be considered: 1) change in antenna structure, 2) main beam gain, 3) side lobe level, and 4) maximum and minimum steering angles.

5.5.1 Effect of Varying the Number of Radiating Elements

The number of driven elements, in the novel antenna, is related to the number of parasitic elements by the expression:

$$N_p = 2(N_d + 1)$$  \hfill (5.3)
Where \( N_p \) is the total number of parasitic elements, and \( N_d \) is the number of driven elements. Hence, an antenna incorporating three driven elements would require 8 parasitic elements. Fig. 5.10 shows the structure of the hybrid parasitic antenna for one, two, three and four driven elements. Conceptually, the antenna can be regarded as being comprised from a pair of parasitics along with one or more repeated blocks. The repeated block comprises of a driven element and a pair of parasitics. The repeated block is simply translated using the chosen spacing between the driven elements.

![Diagram of antenna structures](image)

Figure 5.10: Structure of antenna for different numbers of driven elements (DE=Driven element, PE=Parasitic Element)

### 5.5.2 Gain and Side lobe levels

In a conventional array, increasing the number of driven elements will increase the main beam gain. The same rule applies to this antenna. Specifically it is possible to improve the gain and side lobe level by adding more parasitic elements to the array. To highlight the performance improvements that can be achieved by increasing the number of array elements, the gain and side lobe levels for the hybrid parasitic antenna is compared with those of a conventional array having the same number of driven elements. The comparison is shown in Fig. 5.11.

The parasitic switch states used to achieve the gain was 1111, which corresponds to the
situation where all the parasitics are turned ON. The gain achieved using the hybrid parasitic antenna in this work is better than the gain of the conventional arrays by 0.5-0.6 dBi.

From Fig. 5.11, the side lobe level of the hybrid parasitic antenna with 0.53λ spacing between the driven element is comparable with that of a conventional array having 0.5λ spacing between driven the elements. For a greater inter-element spacing, the side lobe level of the array gets worse when compared to the conventional phase array. However, the side lobe level of the hybrid parasitic antenna is similar to the conventional phased array which is due to the use of parasitic elements in the design.

5.5.3 Maximum and Minimum steering angles

Applying all the possible switch combinations, the maximum and minimum achievable steering angles for the hybrid parasitic antenna as a function of the number of driven element was recorded and shown in Fig. 5.12.

The maximum possible achieved steering angle for the hybrid parasitic antenna reduces as the number of driven elements increase. The angle approaches 5° for more than six driven elements. With the knowledge of the maximum steerable angle, the antenna de-
5.5. Effect of the number of driven elements on antenna performance

Figure 5.12: Maximum steered angle vs No. of driven elements

The minimum achievable steering angle reduces as the number of driven elements increases. This implies that a finer steering resolution can be achieved by increasing the number of driven elements within the hybrid parasitic antenna. The fine steering resolution can be attributed to an increase in the number of parasitic combinations that results from using a larger number of driven elements. For one driven element, the maximum steerable angle is equal to the minimum steerable angle. This is expected as there is only one pair of parasitic elements either side of the driven element. Progressively increasing the number of driven elements has the effect of increasing the resolution of parasitic steering while reducing the steering range.

From Fig. 5.12, the difference between the maximum and minimum steering angles is largest when there are two driven elements. However, for this situation, there are only four parasitic switch combinations yielding only two fine steering angles. Consequently, the steering resolution is insufficiently fine. With three driven elements, a finer steering resolution together with a suitable steering range can be achieved, hence the choice to build and implement an antenna array incorporating $1 \times 3$ driven elements.
5.6 Comparison for minimum steering resolution

In order to evaluate the improvement that this design offers, the minimum steering resolution of the hybrid parasitic antenna is compared with that of a conventional phased array employing digital phase shifters. A phase shifter having a larger number of bits offers a finer angular steering resolution. However, phase shifters having a larger number of bits are more expensive and have higher insertion losses which degrades the performance of the system.

The 2, 3, 4, 5, and 6-bit phase shifters were configured to yield the minimum available steering angle. The exercise was repeated for different numbers of driven elements. The results are presented in Fig. 5.13.

![Figure 5.13: Minimum steered angle achieved for different number of driven elements](image)

From inspection of the Fig. 5.13, it is clear that the steering resolution generally becomes finer as the number of driven elements in the array is increased. When 3-bit phase shifters are used with the hybrid antenna, incorporating 3 driven elements, the steering resolution obtained is equivalent to that for a conventional array incorporating a 5-bit phase shifter. Hence, this hybrid antenna topology supports the use of phase shifters having fewer bits thus reducing the insertion loss.

In Chapter 6 the steering range of the hybrid antenna will be compared with that of a
conventional phased array incorporating 5-bit phase shifters.

5.7 Design guidelines for the hybrid parasitic antenna

This section provides a design methodology which can be used to redesign the antenna for a different application e.g. different operation frequency, different choice of substrate material, etc. The steps in the procedure are listed below:

1. Design the radiator for use at the intended operating frequency.

2. Select the array spacing: The spacing between the array elements should be slightly larger than $0.5 \lambda$. This is needed to create adequate room for the parasitic elements to be close to the driven elements of the array.

3. Size of Parasitic element: The parasitic elements should have a radius comparable to that of the driven element. If the parasitic elements are too small, they will have very little or no effect on the radiation pattern which is undesirable.

4. Position of via holes: For optimum performance in terms of beam steering angle, the vias holes through the parasitic element should be as close to the driven element as possible. If they are placed further away from the driven element, the range of parasitic steering is reduced.

5. Feed Point: The resonant frequency of the driven element is altered by: 1) the presence of other driven elements, and 2) the presence of parasitic elements. This occurs due to the mutual coupling between the elements. A parametric sweep to optimise the feed point of each driven element for good return loss performance at the centre of the operating frequency is carried out.

Following these steps will ensure an easy design and replication of the above mentioned antenna.
5.8 Summary

This chapter presented a hybrid parasitic antenna that utilises two beam steering techniques in order to achieve finer resolution beam steering of $2^\circ$ (achieved using only parasitic steering) over a wide steering range of $\pm 33^\circ$. The approach involved combining the phased array approach with the parasitic steering. Coarse resolution steering was achieved using phased array steering while fine resolution steering was achieved using the parasitic elements. It was shown that increasing the number of driven elements reduces the achievable steering range under parasitic steering. A comparison was made between the minimum steering resolution achievable using the parasitic steering approach, and the minimum resolution achievable using 6-bit phase. Finally, design guidelines were also presented to enable the antenna to be redesigned for a different application.
Chapter 6

Hardwired Prototype of Hybrid Parasitic Antenna

The previous chapter introduced a hybrid parasitic antenna that combines parasitic and phased array steering technique to achieve beam steering. This chapter will present, a hardwired prototype of the hybrid parasitic antenna. Several versions of the antenna with different hardwired parasitic combinations were fabricated and measured to mimic different switch settings in the antenna. Section 6.1 discusses feed network considerations with emphasis on achieving equal phases between the outputs of the feed network. The section also discusses the choice and design of phase shifter used in the antenna along with the optimisation carried out to achieve suitable antenna performance. Section 6.2 presents the design and fabricated hybrid parasitic antenna. Section 6.3 compares results obtained from measurement with those obtained via computer simulation. An in-depth analysis highlighting the differences between the measurement and simulation results is presented along with suggestions on methods for reducing the sources of errors. Section 6.4 presents simulation results for the hybrid parasitic antenna incorporating S-parameter files for real switches and phase shifters. This will help to show the performance of the antenna and the effect of incorporating real switches and phase shifters within the antenna in a practical scenario. Section 6.5 compares the beam steering resolution of the hybrid parasitic antenna with that of other State-of-the-Art (SOTA) techniques in literature. This will show the improvement offered by the hybrid
parasitic antenna proposed in this work. A summary of the chapter will be presented in Section 6.6.

6.1 Fabrication of hardwired parasitic array antenna

A number of challenges had to be resolved in order to fabricate the hybrid parasitic antenna that combines two beam steering techniques. The first challenge was to design a feed network for the driven elements of the array. Power dividers are required in feed networks, as discussed earlier. These power dividers are challenging to design at higher frequencies (e.g. millimetre wave frequencies) because it is easy to excite a transverse resonance across the width of the transmission line. The second challenge was in the choice of phase shifter needed to implement phased array steering in the novel antenna. With each challenge resolved, the feed point of the antenna would need to be optimised to achieve good performance.

6.1.1 Feed network

There are two main approaches for feeding the driven elements within a phased array, namely: in series and in parallel [145]. The series feed has the drawback of adding phase shift and attenuating the signal to the next element and it also introduces some complexity to phased array steering using phase shifters [146]. Hence, for this design, the parallel feeding network is used.

Power dividers are one of the main components within a conventional phased array antenna incorporating radiators which are connected in parallel. The power divider is expected to deliver signals with the same amplitude and phase to each driven element. Power dividers with $2^n$ (where $n$ is an integer $> 0$) number of outputs having equal amplitude and phase are relatively easy to design using multiple cascaded two-way power dividers. For other power divider with output arms not in the order of $2^n$, it is challenging to match the phases of the signals at the outputs of the device. Hence, this challenge is encountered in the design of a three way power divider for the hybrid parasitic antenna with three elements. There are two approaches to tackle this
1. Design a four-way power divider and load the fourth output with a 50Ω resistor. The four-way divider would be constructed by attaching two one-to-two way power dividers to the outputs of a one-to-two way divider. A disadvantage of this approach is that 25% of the input power will be dissipated needlessly in the 50Ω load resistor. However, the necessary dividers are relatively easy to design. Additionally, this approach would ensure equal signal amplitude and phase at each element. This was the first approach implemented in this research, however there was a major challenge observed during measurement that was underestimated in the design phase. At 11 GHz, components such as resistors suffer parasitic reactance and are not pure resistance as compared with lower frequencies. The behaviour of the resistor at higher frequencies is analysed in Section 4.3. For this reason, the use of resistors to terminate microstrip transmission lines to ground is not viable.

2. Design of a three-way power divider. This is a more complex design challenge compared to the previous option. The main challenge in the design of the three-way power divider is to match the phases of the three outputs. Fig. 6.1 illustrates the challenge. Each power dividing arm is a quarter wavelength long at the centre of the operating frequency band. Consider a source port, \(x\) mm away from the centre of the driven elements. The top and bottom dividing arms are offset by an angle of \(\theta\) from the central arm of the power divider. Unfortunately, the distance between the upper arm of the power divider and the uppermost driven element (\(x/\cos\theta\lambda/4\)) is greater than the distance between the central arm of the power divider and the central driven element (\(x\lambda/4\)). This difference in length will introduce an undesirable phase difference in signals arriving at the central driven element and the driven elements at the edges.

In order to compensate for this difference, the length of the central transmission line is extended. This is achieved by further meandering the central feedline as shown in Fig. 6.2.
6.1. Fabrication of hardwired parasitic array antenna

Figure 6.1: Conceptual illustration of challenge in realising a 3-way power divider

Figure 6.2: Structure of designed feed network. ($w_1=2.42 \text{ mm}$, $w_2=0.94 \text{ mm}$, $l_1=4.71 \text{ mm}$, $l_2=5.08 \text{ mm}$, $l_3=3 \text{ mm}$, $l_4=4.52 \text{ mm}$, $l_5=5.18 \text{ mm}$, $l_6=7.48 \text{ mm}$, $l_7=3 \text{ mm}$, $l_8=3.92 \text{ mm}$, $l_9=6.63 \text{ mm}$)

Fig. 6.3 illustrates the result of modelling the structure in CST Microwave Studio in order to assess the phases at the outputs. At an operating frequency of 11.5 GHz (highlighted by the shaded block in Fig. 6.3) there is relatively little phase difference between S21, S31, and S41.
6.1. Fabrication of hardwired parasitic array antenna

6.1.2 Phase shifter integration

Three major factors influenced the choice of the phase shifter used in this design, namely: cost, performance, and ease of integration with the antenna design. The best performing phase shifters tend to be more costly which is undesirable for this application. In order to demonstrate the concept of this hybrid parasitic antenna, three different prototypes were fabricated. Each prototype incorporated a transmission line phase shifter realising a different combination of specific phase shifts at the output ports.

The equation relating the phase shift to the length of transmission line is given by [1]

\[ \theta = \beta l \] (6.1)

where \( \theta \) is the phase shift in radians, \( \beta \) is the propagation constant and \( l \) is the length of the transmission line in millimetres. For a microstrip design, \( \beta \) is expressed as

\[ \beta = k_0 \sqrt{\epsilon_e} \] (6.2)

where \( \epsilon_e \) is the effective relative dielectric constant of the substrate and \( k_0 \) is the wave number in free-space. The wave number, \( k_0 \), is related to the operating frequency by

\[ k_0 = 2\pi f/c \]

Where \( f \) is the frequency of operation and \( c \) is the speed of light in free-space. Hence, the equation relating the length of transmission line to the phase shift is

Figure 6.3: Phase of Signal at power divider output ports
6.1. Fabrication of hardwired parasitic array antenna

given as [1]:

\[ l = \frac{\theta \pi \times c}{2\pi f \sqrt{\epsilon_e}} \]  \hspace{1cm} (6.3)

The effective dielectric constant of the substrate, \( \epsilon_e \), is given by [1]:

\[ \epsilon_e = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \frac{1}{\sqrt{1 + 12d/W}} \]  \hspace{1cm} (6.4)

Where \( \epsilon_r \) is the dielectric constant of the substrate, \( d \) is the thickness of the substrate in millimetres, and \( W \) is the width of the transmission line. The substrate used for the feed network was a Taconic TLY-5 substrate with: \( \epsilon_r = 2.2 \), \( d=0.8 \text{ mm} \), and loss tangent of \( \delta = 0.0009 \) at 10 GHz. In order to set the characteristic impedance \( (Z_0) \) of the transmission line to 50Ω, the ratio of the thickness of the substrate to the width of the transmission was calculated to be 0.3242. Hence, inserting this ratio into eqn. (6.4), the \( \epsilon_e \) is 1.8713.

In order to achieve a main beam steering angle of 22° a progressive phase shift of 80° needs to be applied between driven elements needed. This value was obtained from simulation in CST Microwave Studio. Based on equation 6.3, the length of transmission line required to give a phase shift of 80° is 4.24 mm. Hence, with the first driven element set to 0° phase shift, an extra length of 4.24 mm will be added to the second driven element to achieve an 80° phase shift relative to the first driven element. An additional length of 8.48 mm will be added to the third driven element to achieve 160° phase shift relative to the first driven element.

![Figure 6.4: Feed network and transmission line phase shifts](image)

To accommodate the additional lengths of transmission line whilst also enabling connection with the antenna, the entire feed structure was rotated by 18° clockwise about
the feed point of the first driven element. This would ensure that the first driven element maintains $0^\circ$ phase shift with respect to the other driven elements. Fig. 6.4 shows the layout of the feed and the phase shifts applied.

6.1.3 Hardwired parasitic operation

To operate the parasitic elements, in the hybrid parasitic antenna, the choice of hardwiring the parasitics and using switches were considered. For hardwiring the parasitic elements, copper conductors are used to connect/short the parasitic elements to ground through a via hole to represent the on state for each parasitic element. The conductors are soldered to the parasitic element and the ground plane, and the parasitic elements will maintain that specific state. The parasitic elements without the hardwiring is in the off state. Multiple versions of the hybrid parasitic antenna would be fabricated with different hardwired parasitic combinations to mimic switching between different switch states.

The second approach to operating the parasitic elements is achieved using switches. In other to use the switches, via holes will be drilled through the parasitic elements to the ground plane. A portion of the ground plane around the via hole will be etched out to accommodate a switch. A copper conductor will be inserted through the via hole in a similar manner to the hardwired operation and soldered to the parasitic element. The switch will then be inserted between the ground plane and the copper conductor, one terminal soldered to the conductor and the other terminal soldered to the ground plane. By biasing the switches, the parasitic elements can be toggled off and on.

In this thesis, the hardwired option was chosen for the proof-of-concept hybrid parasitic antenna.

6.2 Design of hybrid parasitic antenna

Fig. 6.5 shows the fabricated prototypes for the hybrid parasitic antenna connected to a three-way power divider incorporating transmission line phase shifts. The antenna was fabricated on a Rogers RT/Duroid 5880, with a dielectric constant of 2.2, thickness
of 1 mm, copper cladding of 17µm, and loss tangent of 0.0009 at 10 GHz. The power divider incorporating transmission line phase shifts was fabricated on a Taconic TLY-5 substrate having a dielectric constant of 2.2, thickness of 0.8 mm, copper cladding of 17µm and loss tangent of 0.0009 at 10 GHz. The radius of the patch and parasitics were 4.74 mm and 4.5 mm respectively. The radius of the via holes were 0.75 mm, and the feed points of the driven elements was offset by 2.89 mm from the centre of each patch. These dimensions are different from those reported in Chapter 5 as the design has been optimised for fabrication. The antenna and feed network were connected back-to-back.

Figure 6.5: Fabricated prototypes for hybrid parasitic antenna and power divider incorporating transmission line phase shifts.
6.3 Results and discussion

The switches within the parasitics of the fabricated prototype antenna were hardwired, as explained in Section 6.2. Five different versions of the antenna were fabricated in order to enable the beam steering performance of two different parasitic combinations to be measured. The five different versions represent:

1. Switch state 1111 with equal phase
2. Switch state 1100 with equal phase
3. Switch state 1111 with 80° progressive phase shifts between driven elements
4. Switch state 1100 with 80° progressive phase shifts between driven elements
5. Switch state 0011 with 80° progressive phase shifts between driven elements

Fig. 6.6 shows the return loss, obtained through simulation and measurement, for switch state 1111. The antenna shows good return loss behaviour around the operating frequency of 11.5 GHz, with a 10 dB return loss bandwidth close to 500 MHz. There is a close agreement between simulation and measurement for this parasitic combination.

![Figure 6.6: Return loss of switch state 1111 with 0 phase shift.](image)
Fig. 6.7 shows the return loss of the antenna corresponding to switch state 1100/0011. These combinations are reciprocal and produce the same return loss. On inspection of Fig. 6.7 it is clear that the return loss obtained through simulation and measurement shifts to 11.6 GHz. This shift is expected when parasitics are used. The change in the states of the parasitics alters the mutual coupling between the driven elements and the parasitic, which also changes the input impedance of the antenna. Although the position of the return loss is altered, simulation results show that the antenna has a good 10 dB return loss bandwidth which covers the intended operating frequency. There is a significant difference between the 10 dB return loss bandwidths indicated through simulation and measurement. The measurement result features a noticeable resonance at 10.9 GHz which is not present in the simulation result. This resonance also appears in Fig. 6.6 for switch state 1111 but is not as pronounced as that for switch state 1100/0011. The resonance at 10.9 GHz alters the position of the lower 10 dB return loss cut-off frequency associated with the operating band of the antenna.

The resonance at 10.9 GHz could be caused by the difference between the fabricated prototype and the designed model in CST Microwave Studio. These differences include: soldering bumps in prototype, unwanted air-gaps between the substrates of power divider and hybrid parasitic antenna, and manufacturing tolerances. The fabricated
6.3. Results and discussion

The antenna has a good return loss at 11.5 GHz, but with a reduced bandwidth compared with the simulation result.

![Graph of return loss plots for various parasitic switch states.](image)

**Figure 6.8:** Return loss plots for various parasitic switch states.

Fig. 6.8 shows the simulated return loss curves associated with the other switch combinations. Generally speaking, turning off a larger number of parasitics has the effect of shifting the resonant frequency of the antenna further away from the intended operating frequency. However, the bandwidth of the antenna covers the intended operating frequency for all parasitic combination states.

Fig. 6.9 shows the radiation pattern for the E-plane of the antenna with parasitic switch state 1111 and 1100 with zero phase shift. For the parasitic switch state 1111, the measured realised gain of the hybrid parasitic antenna is 9.15 dBi which is close to the simulated realised gain of 9.18 dBi. The main beam of the antenna is at 0° (boresight) which is expected since there is no steering when all the parasitic switches are turned on. The measured sidelobe level of the hybrid parasitic antenna is between 3-5 dB lower than the value found through simulation. This effect is noticeable in all the measured radiation patterns. This could be attributed to the measurement setup in the anechoic chamber. The test antenna is clamped to the curved surface of a cylindrical stand on the turn table. During measurement, the test antenna rotates about the centre...
of the cylindrical stand and not about the axis of the antenna. This results in a change in distance and alignment between the receiving antenna and the antenna under test for each position of the turn table away from boresight. This change in distance also affects the gain seen at the receiving antenna, hence the difference observed between simulation and measurement away from the main beam direction.

For the parasitic switch state 1100, the main beam of the antenna is steered $10^\circ$ off boresight. Since the structure is symmetrical, the reverse parasitic switch state 0011 steers the beam by $-10^\circ$ along the E plane. The measured realised gain of the antenna is 9.8 dBi which is better than the gain obtained for boresight. The improvement in gain when the main beam of the antenna is steered is known as negative scan loss. This is the opposite of scan loss which is the drop in gain when the beam is steered off boresight.

Applying an $80^\circ$ progressive phase shift to antenna for the parasitic switch state 1111, the main beam of the antenna is steered to $22^\circ$ along the $x-z$ plane of the antenna using phased array steering. The pattern is shown in Fig. 6.10. The realised gain in that direction is 7.9 dBi. This signifies a scan loss of 1.2 dBi from the boresight direction. Applying parasitic switch states 0011 (Fig. 6.10) and 1100 (Fig. 6.11) to the
6.3. Results and discussion

Figure 6.10: Radiation pattern for parasitic switch states 1111 and 0011 with 80° phase shift.

Phase array steered antenna produces a steering angle and gain of 30°, 8.21 dBi and 15°, 8.73 dBi respectively.

Figure 6.11: Radiation pattern for parasitic switch states 1111 and 1100 with 80° phase shift.
When steering to large angles away from boresight, the sidelobe level of the antenna increases. This is similar to conventional phased array and also other forms of beam steering. The measured sidelobe level for the hybrid parasitic antenna when the radiation pattern is steered by $30^\circ$ using phase shift and parasitic switch states 1100 is 7 dB below the main beam. The use of the parasitic steering technique improves the steering resolution of the hybrid parasitic array antenna and the scan loss of the antenna. This is shown in Fig. 6.9, 6.10, and 6.11 as improvements in gain over the gain when parasitic steering is not used.

If 3-bit phase shifters are employed, the antenna can provide a steering resolution of $1^\circ$. This steering resolution is available over the full steering range of $\pm 40^\circ$ for the hybrid parasitic antenna using a worst-case sidelobe level of -6 dB as reference. This is shown in Fig 6.12. In order to achieve the same steering resolution using a conventional phased array antenna it would be necessary to employ 5-bit phase shifters. Fig. 6.12 shows a comparison between the hybrid parasitic array antenna with 3-bit phase shifter and a conventional phased array antenna with 5-bit phase shifter. The x-axis represents the angle of the antennas main beam while the y-axis represents the realised gain and side lobe level. A line of best fit is added to the plot to observe the trend.

Figure 6.12: Hybrid parasitic antenna vs. conventional phased array based on CST Simulation
The hybrid parasitic antenna, incorporating a 2-bit phase shifter, shows better gain performance when compared to the conventional phased array incorporating 5-bit phase shifters. However, the side lobe level for the hybrid parasitic antenna incorporating 2-bit phase shifters is significantly worse than that of the conventional array. When a 3-bit phase shifter is used with the antenna, the side lobe level improves significantly and compares favourably with the conventional phased array. It should be pointed out that ideal phase shifters, having no insertion losses, were used in these simulations. The effect of using real phase shifters and switches will be presented in the next section.

6.4 Simulation with real switches and phase shifters

In practical scenarios, the effect of the power losses in the switches and phase shifters cannot be neglected. Hence, it is necessary to analyse the hybrid parasitic array antenna to see the effect of these losses on the performance of the antenna. Real switches and phase shifters were selected for use within the antenna in order to study their effects on the antenna’s performance. The switches used were TriQuint SP4T switches operating between 0.2-20 GHz with part number TGS2304-SCC [147]. The switches exhibit insertion losses of 0.6 dB each at a frequency of 11 GHz. The phase shifter used was TriQuint Qorvo 6-bit phase shifter operating between 8-12 GHz with part number of TGP2109-SM [148]. This phase shifter exhibits a nominal insertion loss of 6 dB which was the best insertion loss value available for a 5-bit or 6-bit phase shifter at the time of the analyses.

The circuit simulator in CST Microwave Studio was used to assess the effect of incorporating real switches and phase shifters into the antenna. This was achieved by making use of s2p files, for the switch and phase shifter, from the manufactures. To import the .s2p file into the circuit simulator, discrete ports were used to replace the vias in simulation. The discrete ports were set to serve as an impedance rather than an excitation port. The .s2p file was then imported for each discrete port for both the ON and OFF states of the switch. Fig. 6.13 shows the circuit simulation model describe above.

It was difficult to find a suitable s2p file for a suitable 3-bit phase shifter. A nominal
Figure 6.13: Circuit simulation model of hybrid parasitic array antenna incorporating real switches.

insertion loss value of 2 dB is reported in literature [149] for 3-bit phase shifters. This value was employed within an attenuating block in CST microwave studio. Fig. 6.14 shows the performance of the hybrid parasitic array antenna and the conventional phase array incorporating the selected switches and phase shifters. The x-axis of the figure represents the steered main beam angle. The y-axes represent the realised gain and side lobe levels. The hybrid parasitic antenna shows better performance in terms of realised gain (more than two times better) when compared to the conventional phased array incorporating 5-bit phase shifters. This indicates that although we have added 8 switches to the antenna, reducing the number of bits of phase shifting resolution yields a net improvement in antenna gain of around 4 dB. This is an interesting and somewhat surprising result. The result implies that the switch losses do not add-up as though chained together in series. It is suspected that the losses would add up differently had the switches been inserted in the driven elements or the feedlines of the driven elements. The side lobe level of the hybrid parasitic antenna is similar to that of the conventional phased array.

The hybrid parasitic antenna exhibits better realised gain and steering resolution when compared to the conventional phase array antenna. The hybrid parasitic antenna also
has the advantage of having a better scan loss when compared to the conventional phased arrays as the beam is steered away from boresight. This confirms that the approach of combining the parasitic and phased array steering techniques produces improved performance in gain when compared to conventional phased array and also a better steering resolution when compared to conventional parasitic steering.

### 6.5 Comparison with State of the Art (SOTA)

Table 6.1 compares the steering resolution achieved in this work to others found in literature using different steering techniques.

For phased array using phase shifters, the steering resolution is dependent on the number of bits associated with the digital phase shifters. However, increasing the number of bits leads to an increase in the insertion loss. The novel technique presented in this work has a better steering resolution when compared to other techniques such as Integrated Lens antennas (ILAs), parasitic steerable antenna, traveling wave antennas, reflectarray antennas and conventional phased arrays. However, it should be noted that the beam width using three driven elements is quite broad and would not require such fine steering.
### Table 6.1: Comparison with State-of-the-Art

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<th>Technique</th>
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<th>Steering resolution</th>
<th>Efficiency</th>
<th>Freq. (GHz)</th>
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<td>Discrete (7.5°)</td>
<td>30%</td>
<td>77</td>
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<tr>
<td>[38]</td>
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<td>Discrete (20°)</td>
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<td>Freq. dependent</td>
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<td>[29]</td>
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<td>±40°</td>
<td>Discrete (1°)</td>
<td>75%</td>
<td>11</td>
</tr>
</tbody>
</table>

### 6.6 Summary

This chapter presented design guidelines for fabricating the proposed combined phased array and parasitic steerable antenna. The chapter also presented a detailed discussion on the choice and design of the feed network stating the advantages of alternative solutions. Measurement and simulation results showing the performance of the novel antenna were also presented and compared. Simulation results were shown to assess the effect of using real switches and phase shifters. A comparison was drawn against the performance of a conventional phased array having the same steering resolution. The hybrid parasitic antenna yields a 4 dB improvement in the realised gain in comparison with a conventional phased array. Finally, the novel approach used in this work was compared with the state-of-the-art.
Chapter 7

Conclusion and Future Work

This chapter will present a conclusion of the work carried out and the contributions reported in this thesis. Discussions of the challenges and limitations encountered during the research will also be presented. Future works based on the research findings will be presented.

7.1 Conclusion

In chapter 2, an overview of beam steering techniques was presented. Beam steering techniques such as parasitic steering, integrated lens antennas, and switched beam antennas offer large beam steering angles with poor steering resolution which is inadequate for 5G millimetre wave cellular applications that require mobility. While the conventional phased array, reflectarray, and metamaterial beam steering techniques suffer from high insertion losses and high cost of phase shifter with large number of bits (for phased array and reflectarray) when fine steering resolution is required for millimetre wave applications. Hence, the main objective of this thesis was to develop novel beam steering solutions that will improve the steering resolution of beam steerable antennas while keeping the cost, complexity and insertion loss at a minimum. These novel beam steering solutions were presented in Chapter 4, and 5 of this thesis.
Passive Amplitude Steerable Antenna

In Chapter 4, the first of the solutions, a passive amplitude steerable antenna, was successfully designed using CST Microwave Studio, fabricated and measured in the anechoic chamber. The antenna consisted of two circular microstrip patch elements fed using a reconfigurable ratio power divider (RRPD). The RRPD was designed with a ratio of 1:2.5 and could achieve a reconfigurable ratio from about 1:1 to 1:6.25. The RRPD made use of a potentiometer to achieve reconfiguration and the worst-case insertion loss for the RRPD was 1.05 dB. The antenna, incorporated with the RRPD, achieved a continuous steering range from boresight to 21° in one direction. The measured antenna had a peak gain of about 5 dBi and an efficiency of 55% at boresight. The concept of the passive amplitude steerable antenna was tested with antenna arrays with more driven elements and it was observed that the steering range reduces as the number of driven elements increase. Hence, the proposed concept offers a small scale, low loss, and cost-effective approach of achieving beam steering which achieves the main objective of this thesis. The limitations of the passive amplitude steerable antenna are:

1. The lower power arm of the RRPD becomes poorly matched when the value of the potentiometer tends to the maximum value. This would limit the range of the reconfigurable ratio available to the RRPD

2. The reconfiguration of the RRPD by the potentiometer is carried out manually. Other forms of digital potentiometers would be needed to implement electronic reconfiguration.

3. The designed 2 × 1 antenna array could only steer towards one direction.

Hybrid Parasitic Array Antenna

In chapter 5, the second steering solution, which involved the combination of two other beam steering techniques, was presented. The two techniques combined were the phased array and parasitic beam steering techniques. The phased array technique controlled the large angle steering while the parasitic steering technique controlled the fine angle
steering within each large steered angle direction. The combination of the techniques resulted in a hybrid parasitic array antenna (HPAA). A $3 \times 1$ HPAA consisting of circular microstrip patches and eight circular parasitic patches was designed using CST Microwave Studio, fabricated and measured in the anechoic chamber and the results presented in Chapter 6. The antenna operated at 11.5 GHz and the measured gain at boresight was 9.8 dBi. With the use of switches to operate the parasitic elements, and a 3-bit phase shifter, a steering range of $\pm 45^\circ$ and a steering resolution of $1^\circ$ was achieved. For a similar steering resolution in phased array, a phase shifter with 5-bit would be needed. The HPAA showed a 4 dB improvement over a conventional phased array with 5-bit phase shifter when real switches and phase shifters were used in simulation. Hence, the objective of reducing the loss incurred by the phase shifter of this thesis was achieved in this solution. The cost of implementing this solution was also reduced as the cost of phase shifters increases with the number of bits. The HPAA approach is suitable for large number of driven elements in the array. The main limitation of this approach is that the switches used to operate the parasitic elements needed in achieving fine beam steering increases the complexity of designing and operating the HPAA.

**Compensated Power Divider**

In addition to beam steering solutions, a novel approach for improving the 15 dB isolation bandwidth of power dividers was also presented. The objective of this task was to reduce the complexity associated with other techniques used in improving the isolation bandwidth of power dividers such as multistage matching isolation sections, coupled lines, and isolation networks comprising of lumped elements. The novel power divider, a compensated power divider (CPD), was designed and analysed using ADS and CST Microwave studio. The CPD was fabricated and measured using a Vector Network Analyser. The isolation between the output ports of the CPD showed an ultra-wide 15 dB isolation bandwidth of 114% at an operating frequency of 6 GHz. The complexity of this approach is similar to the conventional Wilkinson power divider requiring just one lumped element resistor. The CPD compares favourable to other techniques as it achieve similar better isolation bandwidth with less complexity. The CPD presented in this work is suitable for ultra-wide bandwidth applications. The
limitation of the CPD is that it achieves the wider isolation bandwidth at the cost of poorer return loss at the operating frequency. With a knowledge of the return loss requirement of an application, a CPD with an optimised isolation bandwidth can be designed for optimum performance.

7.2 Future Work

Research is a never-ending process, the more you discover, the more you are yet to discover. Possible future work arising from the work in this thesis includes:

1. It would be interesting to investigate the relationship between the 15 dB isolation bandwidth and the isolating resistance in the compensated power divider, presented in Chapter 3. The equation derived in this work relates the best isolation at the operating frequency, however, the best isolation does not equate to the best isolation bandwidth. Initial studies on this topic indicate that it would be possible to derive an expression that relates these two variables.

2. Characterisation of the potentiometer for high frequency applications. Very little or even nothing is known about the behaviour potentiometers at microwave frequencies. Section 4.3 presented an equivalent circuit for a potentiometer based on that of a resistor in a similar package. However, this analyses was not validated experimentally, hence the need to characterised the potentiometer for high frequency operation. With this knowledge, the frequency limit of the potentiometer and the proposed RRPD would be well defined. This knowledge would also prove valuable for the approach of amplitude steering presented in Chapter 4.

3. Array steering equation for amplitude steerable antenna arrays. A formula to compute the beam steering direction of phased array that takes into consideration the inter-element spacing and the phase shift applied to each array element is available. However, there is no available equation for an amplitude steerable antenna. It would be worthwhile to derive such an equation that relates the amplitude of the signal fed to each array element and the inter-element spacing to the steered angle that can be achieved.
4. The amplitude steerable antenna proposed in Chapter 4 could only steer its beam towards one direction, off boresight. It would be interesting to investigate the potential for achieving beam steering in both directions. Initial insight into this would suggest that this could be achieved by employing a second RRPD with the arms reversed compared with those of the first RRPD. The first RRPD would then be used to steer the beam towards one direction and the second would be used to steer the beam in the other direction. The challenge with this approach will be in incorporating switches to isolate the sections of the RRPD that is not in use.

5. Design rules for the hybrid parasitic antenna. The hybrid parasitic antenna is a promising solution and an alternative to a phased array antenna which need to incorporate phase shifters having a larger number of bits in order to achieve the same steering resolution. Section 5.7 presented design guidelines which involved carrying out a parameter sweep on the feed point to optimise the return loss performance of the antenna. It would be interesting to investigate whether design equations could be derived to determine the design equations that yield optimum antenna performance without the need for parameter sweeps. The equations would be used to determine: the spacing of the driven elements, the height of the parasitics above and below the driven elements, the position of the vias within the parasitics, and the feed point on the driven elements. The design rules can then be extended for various number of driven elements.
Bibliography


[70] H. Li, D. Ye, F. Shen, B. Zhang, Y. Sun, W. Zhu, C. Li, and L. Ran, “Reconfigurable Diffractive Antenna Based on Switchable Electrically Induced Trans-


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