Improved designs of antennas, power dividers and antenna arrays using modified Gielis curves

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**IMPROVED DESIGNS OF ANTENNAS, POWER DIVIDERS AND
ANTENNA ARRAYS USING MODIFIED GIELIS CURVES**

SHANMUGAM BHASKAR VIGNESH

**SCHOOL OF ELECTRICAL & ELECTRONIC
ENGINEERING**

2020

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ANTENNA ARRAYS USING MODIFIED GIELIS CURVES**

SHANMUGAM BHASKAR VIGNESH

School of Electrical & Electronic Engineering

A thesis submitted to the Nanyang Technological University
in partial fulfillment of the requirement for the degree of
Doctor of Philosophy

2020

Statement of Originality

I hereby certify that the work embodied in this thesis is the result of original research, is free of plagiarized materials, and has not been submitted for a higher degree to any other University or Institution.

20 March 2020

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Date

SBVignesh

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Shanmugam Bhaskar Vignesh

Supervisor Declaration Statement

I have reviewed the content and presentation style of this thesis and declare it is free of plagiarism and of sufficient grammatical clarity to be examined. To the best of my knowledge, the research and writing are those of the candidate except as acknowledged in the Author Attribution Statement. I confirm that the investigations were conducted in accord with the ethics policies and integrity standards of Nanyang Technological University and that the research data are presented honestly and without prejudice.

20 March 2020

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Date

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(Assoc Prof) Tan Eng Leong

Authorship Attribution Statement

This thesis contains material from 1 published paper, 1 paper ‘under revision’ and 1 paper to be submitted in the peer-reviewed journals. It also contains material from 8 published international conference papers in which I am listed as an author.

Chapter 3 contents are from

1. Vignesh Shanmugam Bhaskar, E. L. Tan, K. H. H. Li, and M. S. Tse, “Super shaped CPW-fed Monopole Antenna With Parasitic Strips for Unidirectional Pattern,” 2018 IEEE international Symposium on Antennas and Propagation & USNC/URSI National Radio Science Meeting, pp. 1931-1932, 2018.
2. Vignesh Shanmugam Bhaskar, E. L. Tan, K. H. H. Li, and M. S. Tse, “Broadband CPW-fed Slot Antenna Using Super formula,” 2018 IEEE international Symposium on Antennas and Propagation & USNC/URSI National Radio Science Meeting, pp. 1277-1278, 2018.
3. Vignesh Shanmugam Bhaskar, E. L. Tan, K. H. H. Li, and M. S. Tse, “Design of Planar Mushroom-shaped wideband Monopole Antennas Using Gielis Curves,” 2018 IEEE Asia-Pacific Conference on Antennas and Propagation (APCAP), pp. 231-232, 2018.
4. Vignesh Shanmugam Bhaskar, E. L. Tan, and K. H. H. Li, “UWB double-sided printed bowtie antenna using super shape,” 2018 Asia-Pacific Microwave Conference (APMC), pp. 1612-1614, 2018.

5. Vignesh Shanmugam Bhaskar, E. L. Tan, and L. K. H. Holden, "Design of wide-band bowtie slot antenna using sectorially modified Gielis curves," *IEEE Antennas Wireless Propag. Lett.*, vol. 17, no. 12, pp. 2237-2240, 2018.

The contributions of the co-authors are as follows:

- I came up with the idea of applying modified curves for different antenna types and I prepared the manuscripts.
- Assoc. Prof. Tan Eng Leong gave some technical suggestions and edited the manuscripts.
- Asst. Prof. Li King Ho Holden and Assoc. Prof. Man Siu Tse assisted in EM simulation of antennas and procuring printed circuit board laminates for fabrication.

Chapter 4 contents are from

1. Vignesh Shanmugam Bhaskar, E. L. Tan and K. H. H. Li, "1 to 4 Way Wide-band Power Divider using Substrate integrated Waveguide and Modified Wilkinson Structures," 2018 Joint IEEE EMC & APEMC Symposium, pp. 554-557, 2018.
2. Vignesh Shanmugam Bhaskar, E. L. Tan, "Center-Fed Four-Way Power Divider with Wideband Harmonic Suppression " (To be submitted).

The contribution of the co-authors are as follows:

- Assoc. Prof. Tan Eng Leong gave the initial direction of improving the harmonic suppression, gave technical suggestions and revised the manuscripts.
- I came up with the idea of using modified curves in stubs and prepared the manuscripts.

- Asst. Prof. Li King Ho Holden assisted in EM simulation of antennas and procuring printed circuit board laminates for fabrication.

Chapter 5 contents are from

1. Vignesh Shanmugam Bhaskar, E. L. Tan, and K. H. H. Li, “Three-dimensional bandstop frequency-selective structures based on Gielis-shaped loop resonator,” 2019 IEEE International Symposium on Antennas and Propagation & USNC/ URSI Notional Radio Science Meeting, pp. 1191-1192, 2019.
2. Vignesh Shanmugam Bhaskar, E. L. Tan, and K. H. H. Li, “Periodic Leaky-Wave Antenna with Modified Gielis-Shaped Patch Elements,” 2019 IEEE International Symposium on Antennas and Propagation & USNC/URSI Notional Radio Science Meeting, pp. 135-136, 2019.
3. Vignesh Shanmugam Bhaskar and E. L. Tan, “Bidirectional linearly polarized grid-slotted patch antenna with Gielis-shaped patch,” *Asia-Pacific Microwave Conference*, pp. 1542-1544, 2020.
4. Vignesh Shanmugam Bhaskar, E. L. Tan, “Same-Sense Circularly Polarized Grid-Slotted Patch Antenna with Wide Axial Ratio Bandwidth” (submitted to IEEE Transactions on Antennas and Propagation - under revision).

The contribution of the co-authors are as follows:

- I came up with the idea of using flipped Z-shaped slots to excite grid-slotted patches to realize a same-sense circular polarized antenna with a wide axial ratio bandwidth and I prepared the manuscript.
- Assoc. Prof. Tan Eng Leong gave technical suggestions to improve the results and revised the manuscripts.

- Asst. Prof. Li King Ho Holden assisted in EM simulation of antennas and procuring printed circuit board laminates for fabrication.

20 March 2020
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Date

SB Vignesh
.....

Shanmugam Bhaskar Vignesh

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Summary

Planar antennas have been extensively used in many wireless applications due to their compact size, easy design and low fabrication cost. Some planar arrays need power dividers with filtering properties, especially in applications involving harmonic suppression. The electrical behaviour of any conductor patch or slot in the design of antenna or power divider can greatly influence its performance. Hence, sectorially modified Gielis curves (SMGCs) can be used to improve the performance characteristics of antennas or power dividers or antenna arrays such as increased impedance bandwidth, increased gain and reduced lower cut-off frequency. The SMGCs will provide additional degrees of freedom (DOF) in the design of most structures.

Firstly, the SMGCs are applied to enhance the performance characteristics of single element antennas. A CPW-fed monopole antenna with parasitic strips for unidirectional pattern is designed using SMGCs for the shape of the patch with planar reflector, to reduce the cut-off frequency while the parasitic strips are used to enhance the directivity. A broadband printed CPW-fed slot antenna is designed by using SMGCs for the shape of the slot and the patch, to provide additional paths for the surface current along the edges. This SMGC-based edge profile can influence the impedance matching. A planar mushroom-shaped wideband monopole antenna is designed using SMGCs as they are capable of generating complex shape profiles for the conductor patch conveniently and to improve the bandwidth. An ultra-wideband (UWB) double-sided printed bowtie antenna using SMGCs is designed to improve the surface current which in turn improves the impedance matching and the radiation gain. Moreover, a design of wideband bowtie slot antenna using SMGCs is discussed. The SMGCs used for the slot profile shows considerable improvement in the bandwidth and the gain when compared to a conventional bowtie slot antennas.

Secondly, the SMGCs are used in the feeding network, i.e., the power divider (PD) that

feeds the antenna arrays. A center-fed four-way PD with wideband harmonic suppression is proposed with stubs that are designed using SMGCs to provide more DOF. The harmonic suppression is improved due to the introduction of open stubs, short stubs and stub-loaded coupled lines using SMGCs. Of the two stages, the first stage is a center-fed two-way power divider with open and short stubs, while the second stage is a modified two-way Wilkinson power divider with stub-loaded coupled-line structures.

Thirdly, the SMGCs are used in the designs of antenna arrays. They are applied in array or periodic structures such as frequency selective surface structures, one-dimensional periodic leaky-wave antennas and grid-slotted patch antennas with linear and circular polarization. A three-dimensional bandstop frequency selective structure (FSS) based on a loop resonator is designed using SMGCs for the full and half loop resonator, to obtain a reduction in operating frequency when compared to a circular profile for a fixed unit cell size. A periodic leaky-wave antenna with SMGC for patch element is designed, to enhance the operating bandwidth and the gain of the leaky wave antenna. A bidirectional linearly polarized grid-slotted patch antenna with SMGCs for patch elements can slightly reduce the lower cut-off frequency. A same-sense circularly polarized grid-slotted patch antenna with wide axial ratio bandwidth is designed by using SMGCs to form the patch elements. The patch elements are excited through flipped Z-shaped slots pair using a stripline feed to generate same-sense circular polarization with a wide axial ratio bandwidth in the category of same-sense antennas.

Hence, SMGCs can play a significant role in improving the performance characteristics of antennas, power dividers or antenna arrays, since geometry of the conductor patches or slots can directly affect the surface currents.

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List of Abbreviations

ADS	Advanced design system
ARBW	Axial ratio bandwidth
CBP	Conventional bowtie patch
CMA	Characteristic Mode Analysis
CPSS	Circular polarized same-sense
CRLH	Composite right/left-hand
CRP	Conventional rectangular patch
CSRR	Complementary split ring resonator
CST	Computer simulation technology
DOF	Degrees of freedom
FBW	Fractional bandwidth
FCFS	Frequency-selective coupling structures
HFSS	High frequency structure simulator
LHCP	Left hand circular polarization
MGP	Modified Gielis patch
MRCP	Monopole with rectangular and circular patch
PD	Power divider
PLWA	Periodic leaky wave antenna
RHCP	Right hand circular polarization
RMP	Rectangular microstrip patch
SIW	Substrate integrated waveguide
SRR	Split ring resonator
TTL	Tapered transmission line
VSWR	Voltage standing wave ratio

Chapter 1

Introduction

Antennas have found extensive usage in modern-day commercial, industrial, medical, military applications, radars, communication, medical sciences, and security systems. Based on the application, the size, gain, power handling, polarization, feeding mechanism of the antenna are designed for the required operating frequency and bandwidth. Planar antennas have always been a better choice for most of the antenna applications because of their low profile, simple design, low fabrication cost, and the ability to integrate with other components such as the back-end circuits in an antenna radar system for signal processing. Planar antenna is a broad category of antenna type which involves any antenna whose design mostly involves a single dielectric layer with metal claddings on one or both sides. Designs having more than one dielectric layer with metal cladding are also considered as planar antennas. In short, designs that primarily depend on printed circuit board technology for fabrication comes under planar antennas. The following are some of the most commonly used planar types which involve patch antenna, slot antenna, printed monopole antenna, printed Yagi-Uda antenna, printed Vivaldi antenna, and spiral antenna. They may have further variations based on their feedings such as microstrip-feed, coplanar waveguide feed, aperture coupling, or proximity coupling in multilayer or stacked patch antennas. Choosing an appropriate planar antenna will be based on observing the tradeoffs among the several basic parameters like bandwidth, gain and size

of the antenna. There are different types of antennas that can be potentially used as a radiating element for high gain antenna arrays such as slot coupled printed antenna [1], dual loop antenna [2], tapered slot Vivaldi antenna [3–5], circular ring slot antenna [6], log periodic dipole antenna array [7], stripline-fed slot antenna [8], stripline-fed bow-tie antenna [9], meta-material mushroom antenna [10], and printed Lotus antenna [11].

Planar antennas are widely used in applications requiring ultra-wideband (UWB) performance since UWB technology has gained much popularity among researchers and the wireless industry after the FCC adopted its frequency band of 3.1 to 10.6 GHz. The merits of UWB is its capability to offer high capacity short-range wireless communication links using low-cost low-energy transceivers and can support many wireless standards, such as IEEE 802.11, IEEE 802.11a or HIPERLAN/2, IEEE 802.11b, IEEE 802.11g, IEEE 802.11n. Planar antennas are also used in multiband antennas as they have become a necessity for the state-of-the-art multifunction “smart phones” and wireless sensor modules for the mobile devices. Such internal antennas are generally required to cover the frequency bands of GSM- 850 / 900 / 1800 / 1900 and UMTS (824–894 / 890–960 / 1710–1880 / 1850–1990 / 1920–2170 MHz). In addition, the ever-increasing implementation of 4G devices further increases the bandwidth requirement to cover the LTE2300 (2305–2400 MHz), LTE2500 (2500–2690 MHz), and WiMAX bands.

The multiband application also involves broad-band or dual-band applications that are necessary for mobile handset antenna applications. The radiating element is usually based on a square, circular, or an elliptical-shaped patch for the planar monopole antenna to obtain a wide bandwidth. Also, a variety of printed slot antenna configurations have been used as planar antennas, including straight slot, wide slot, L-shaped slot, inverted T-slot, step slot, and fractal-shaped slot, etc. Apart from linear polarized applications, planar antennas also play a significant role in applications involving circular polarization. Circularly polarized (CP) antennas have received much attention for finding applications in satellite positioning, radio frequency identification (RFID), and sensor systems due

to their insensitivity to depolarization effects, resistance to bad weather conditions, and immunity to multipath propagation. CP waves can be generated when two degenerate modes of equal amplitude and phase difference of 90° are excited.

The shape or curve profile of the conductor patch in planar antennas can have greater impact on the performance characteristics of the antenna. To better understand the effect of shape profiles on basic antenna characteristics, three conventional geometries, i.e., rectangle, circle, and ellipse, are chosen to form the shapes of a patch. A comparison has been made by keeping the same resonant frequency when designing the patch with various geometrical shapes. The FR-4 substrates are the same for all cases with $\varepsilon_r=4.3$ and thickness 1.6 mm. The overall substrate dimensions are 25mm $\times$ 25mm. The patch shapes have been tuned to resonate at the same frequency, and their feed locations have been adjusted along y axis to obtain a better impedance matching. The formulae for the resonance frequency of the rectangular and circular patch antennas based on the cavity model are obtained from [12], which are quasi-static formulae and are valid over a wide band of frequency range. The resonant frequency (f_r) of rectangular patch antennas for dominant mode TM_{010} when length(L) > width(W) with fringing effect is given by [12]

$$\begin{aligned}
 (f_r)_{010} &= \frac{1}{2L_{eff}\sqrt{\varepsilon_{reff}}\sqrt{\mu_o\varepsilon_o}} \\
 L_{eff} &= L + 2\Delta L \\
 \varepsilon_{reff} &= \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left[1 + 12 \frac{h}{W} \right] \\
 \frac{\Delta L}{h} &= 0.412 \frac{(\varepsilon_{reff} + 0.3) \left[\frac{W}{h} + 0.264 \right]}{(\varepsilon_{reff} - 0.258) \left[\frac{W}{h} + 0.8 \right]}
 \end{aligned} \tag{1.1}$$

The resonant frequency f_{rc} for circular patch antenna with dominant mode TM_{110} with

fringing fields are given by [12]

$$(f_{rc})_{110} = \frac{1.8412}{2\pi a_e \sqrt{\varepsilon_r} \sqrt{\mu_o \varepsilon_o}} \quad (1.2)$$

$$a_e = a \left\{ 1 + \frac{2h}{\pi a \varepsilon_r} \left[\ln \left(\frac{\pi a}{2h} \right) + 1.7726 \right] \right\}^{1/2}$$

Similarly the resonant frequency for elliptical patch antennas using quasi-static considerations are given by [13, 14]

$$\begin{aligned} {}_e f_{r1} / {}_e f_{r0} = {}_o f_{r1} / {}_o f_{r0} &= \frac{\varepsilon_r}{\{\varepsilon_e(w_1) \varepsilon_e(w_2)\}^{1/2} (1 + \Delta)} \frac{1}{\Delta} \\ \Delta &= \frac{h}{w_1} \left[0.882 + \frac{0.164(\varepsilon_r - 1)}{\varepsilon_r^2} \right. \\ &\quad \left. + \frac{\varepsilon_r + 1}{\pi \varepsilon_r} \left\{ 0.758 + \ln \left(\frac{w_1}{h} + 1.88 \right) \right\} \right] \\ \varepsilon_e(w) &= \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left\{ 1 + \frac{10h}{w} \right\}^{-1/2} \\ {}_e f_{r0} &= \frac{c}{\pi a e} \{q_e / \varepsilon_r\}^{1/2}; \quad {}_o f_{r0} = \frac{c}{\pi a e} \{q_o / \varepsilon_r\}^{1/2} \end{aligned} \quad (1.3)$$

Fig. 1.1(a) shows the overlapped geometrical shapes of the patches based on 1.1 to 1.3 on a similar substrate for comparison. It should be noted that overlapped patches are almost similar in length along the y axis, but a noticeable difference in width can be seen along the x axis, as shown in Table 1.1. From Fig. 1.1(b), it is observed that at a given resonance frequency 4.2 GHz, the area of the elliptical patch is considerably smaller than that of the rectangular and circular patches. The elliptical patch area is smaller than the circular patch and rectangular patch by 34.4% and 33%, respectively. When considering only the geometries' width, the elliptical patch has its width reduced by 36.17% and 29.4% compared to that of the circular and rectangular patches, respectively. Hence, certain specific geometric shapes can significantly help reduce the layout size compared to other geometric shapes while having a similar operating frequency. It should also be noted that there is a small compromise in the gain of the elliptical patch when compared to the rectangular and circular patch as the area becomes smaller. Fig. 1.2

Figure 1.1: (a) Overlapped conventional geometric shapes for comparison. (b) S_{11} and realized gain of coaxial-fed rectangular, circular, and elliptical shaped patches.

Table 1.1: Dimensions of overlapped patches

Geometry	Dimensions (mm)	shape center from origin (x,y)	feed center from origin (x,y)	area (mm ²)	% in area*	% in width*
rectangle	17 × 16	(12.5, 12.5)	(12.5, 8.5)	272	98	90.4
circle	radius = 9.4	(12.5, 12.5)	(12.5, 8.5)	277.4	100	100
ellipse	x radius = 6; y radius = 9.7	(12.5, 12.5)	(12.5, 9.5)	182	65.6	63.8

* % of area or width normalized to that of the circular patch

shows the surface current distribution for the rectangular, circular, and elliptical patch shapes. The current distribution in the form of cones is shown as a graphical technique for providing a comprehensive and intuitive representation of the current distribution on planar patches using commercial full-wave simulation software like CST, HFSS. All the current distributions discussed in the thesis refer to the current distribution on the upper surface of the patch conductor. From the current distribution of the patch antennas, the rectangular, circular, and elliptical patch antennas operate on fundamental modes TM_{010} , TM_{110} , and TM_{010} , respectively. It can be seen that the direction of surface current is vertical i.e., along the y axis for the rectangular patch, while the direction of surface current is not entirely vertical along the y axis. Also, the surface currents are comparatively high along the edges of a conductor patch. Therefore, the shape of the patch or slot and the feed point location can greatly affect the input impedance and radiation gain characteristics of the antenna.

Figure 1.2: Surface current distribution of coaxial-fed patch antennas with (a) rectangle, (b) circle, and (c) ellipse shaped patches.

From Fig. 1.1 and Fig. 1.2, it can be inferred that the antennas have the same ground plane size, similar conductor patch area, and perimeter. However, the shape of the conductor patch can significantly influence the surface current distribution, which can affect the impedance matching, impedance bandwidth, input impedance, and radiation characteristics of the antenna. The effects of surface current distribution not only apply to the antennas but also in the design of power dividers. These power dividers can be used to feed the antenna arrays, especially for omnidirectional applications, where the antenna elements in the array are required to be placed in a circular arrangement from the center-feeding point. The key observation that the author intends to point out is that the conductors with different geometric shapes occupying a similar area on a substrate can exhibit different S parameter responses. This property of shapes can be used to improve the performance characteristics of antennas or power dividers without the need of additional components.

1.1 Objectives

The main objective of the research is to improve the designs of planar antennas, power dividers, and even some antenna arrays that have directional or bidirectional radiation. Here, the phrase ‘improve the design’ refers to improving at least one of the antenna or power divider characteristics such as impedance bandwidth, impedance matching, realized

gain, axial ratio bandwidth, or harmonic suppression, without adding any additional components like resistor, capacitor or inductor. The size of the antenna or power divider is maintained the same, while attempts should be made to improve the performance by taking advantage of geometrical modifications of the conductor patches or slot cavities in the design. The geometrical modifications of the conductor patches or slot cavities can be done using a special curve with a closed formula called modified Gielis curve. We shall limit our research scope to planar antennas only in this thesis, although we have also investigated several non-planar antennas earlier [15–17]. In order to attain the main objective of the research, a few goals need to be achieved, which are summarized as follows:

1. The antenna characteristics of the planar antennas, with microstrip-feed or coplanar waveguide feed having unidirectional or bidirectional radiation properties and linear polarization, should be enhanced by the usage of modified Gielis curves in the planar antenna design. A literature review on bowtie slot antennas gives information about the existing improvements in bowtie slot antennas using different techniques. The proposed modified Gielis curves need to be applied to the slot profile of a conventional bowtie slot and explore the configurations of the modified Gielis curve that can improve any of the bowtie slot antenna characteristics.
2. The proposed modified Gielis curve is aimed to be used in power divider designs. In the design of omnidirectional antenna arrays that are horizontally or vertically polarized, the antenna elements need to be placed in a circular configuration and excited from the output of the power dividers having a common feeding point at the center. To avoid harmonics in such antenna arrays, the power dividers should be designed with the required harmonic suppression. Though many techniques are available for achieving harmonic suppression in power dividers, the modified Gielis curves should be applied to the stubs or conductor patches to provide better control over the harmonic suppression characteristics in the design.

3. The proposed modified Gielis curve is then targeted to be used in periodic or array structures. The modified curves are first used to improve performances of structures without excitation such as frequency selective surfaces, and then later attempted for antenna array entities like periodic leaky-wave antennas and grid-slotted patch antennas. The modified Gielis curves should be used in grid-slotted patch antennas to support circularly polarized radiation with same-sense and to obtain a wide axial ratio bandwidth compared to other antennas of its category.

1.2 Major contributions of the thesis

This thesis gives suggestions to improve planar antenna designs by exploring new geometries based on Sectorially Modified Gielis curves (SMGCs) without adding any extra components or complexity. Studies on the shapes of stubs in the power dividers and periodic or array type structures are also presented. The following are the major contributions of this thesis:

1. SMGCs can improve the impedance bandwidth characteristics or reduce the lower cut-off frequency of antennas or power dividers or antenna arrays. The SMGCs can provide additional degrees of freedom while tuning for a structure.
2. SMGCs are used in the design of microstrip-fed or CPW-fed planar antennas such as monopole antenna with parasitic strips; a slot antenna with a tuning stub; mushroom-shaped monopole antenna, and an ultra-wideband (UWB) double-sided printed bowtie antenna. The SMGCs can offer better impedance matching or reduced lower cut-off frequency to some extent or better gain. A noticeable improvement is shown in the design of the wideband bowtie slot antenna using the SMGCs. The SMGCs are used for the slot profile in the bowtie slot to improve the bandwidth while maintaining the antenna size. Two prototypes of BSAs are built on FR4 and RO5880 substrates providing fractional bandwidths of about 90.3% (1.7 GHz to

- 4.5 GHz) and 71.4% (2.7 GHz to 5.7 GHz), respectively. The designed bowtie slot has a low cut-off frequency compared to antennas of its category with the same size.
3. A center-fed four-way power divider with wideband harmonic suppression is designed. The proposed four-way power divider is designed in the following two stages: The first stage is a center-fed two-way power divider with open and short stubs; and the second stage is a modified two-way Wilkinson power divider with stub-loaded coupled-line structures. These stubs and coupled-line structures are useful for harmonic suppression. The proposed four-way power divider is center-fed and can provide wideband harmonic suppression of up to $4f_o$, while providing a relatively wide -10 dB impedance bandwidth of 52%. An expression for the ABCD matrix of the stub-loaded coupled line structure with different even and odd mode electrical lengths ($\theta_e \neq \theta_o$) is presented.
 4. SMGCs are applied to periodic or array structures such as three-dimensional band-stop frequency-selective structure (FSS) based on loop resonator, and it provides a reduction in operating frequency by 10% when compared to a circular profile for a fixed unit cell size. SMGCs have improved the bandwidth of leaky wave antennas with one-dimensional periodic unit cells by 12% compared to conventional square patch elements in the periodic leaky-wave antenna.
 5. A bidirectional linearly polarized grid-slotted patch antenna with a Gielis-shaped patch is presented. A coplanar waveguide (CPW) fed slot is used to excite two meta-surface layers with a Gielis-shaped patch for bi-directional radiation and reduce the lower cut-off frequency by 4%. A same-sense circularly polarized grid-slotted patch antenna with wide axial ratio bandwidth is proposed. The rectangular grid-slotted patch elements from modified Gielis curves on top and bottom planes are excited through flipped Z-shaped slots pair using a stripline feed to generate same-sense circular polarization with better axial ratio bandwidth. The dispersion characteristics of the unit cell in the grid-slotted patches are discussed, and an approximate

relation between the first and second resonant frequencies and the axial ratio bandwidth variations are discussed. The antenna has a 3 dB axial ratio bandwidth of about 21% from 4.07 GHz to 5.02 GHz with realized gain ranging from 2 to 3.5 dBic.

1.3 Organization of the thesis

This thesis has been organized into six chapters. Chapter 1 provides the background of the conductor patches and the effects of conductor patches on the surface current distribution. The objectives of this study are also discussed. The motivation to use specialized formula to generate the contour or shape of the conductor or slots in the design of antennas or power dividers or the antenna arrays to obtain improved performance is discussed. Further, it explains the contributions and organization of the thesis.

Chapter 2 gives the extensive literature review of the theory and analysis to design the shape of the conductor or slot in the planar antenna designs, power dividers, and antenna arrays. This chapter also discusses most planar antennas, where the coplanar waveguide or microstrip line is used to excite the patch or the slot. In any case, the shape or curve profile of the conductor or slot can have an appreciable influence on the performance of the antennas, power dividers, or antenna arrays. Further, the literature review concludes by introducing the SMGCs that are based on closed-form formula.

Chapter 3 presents the usage of SMGCs in determining the shape of the conductors or slots in unidirectional or bidirectional antennas that are either fed by microstrip or coplanar waveguide. This chapter also explains the effects of SMGCs on the slot profile of a bowtie antenna and its influence on the impedance bandwidth of the antenna.

Chapter 4 proposes the design of power dividers that are center-fed or coaxially-fed perpendicular to the substrate to feed antenna elements placed in a circular arrangement. The design is based on antenna elements that are placed at an equal distance from the center-feeding. The influence of stubs on the transmission zeros and the influence of

modified Gielis curves to determine the shape of the stubs are also discussed.

Chapter 5 deals with the design of periodic or array structures using the modified Gielis curves. The ability of the modified Gielis to form geometries with right-angled edges is also discussed, and their usage in grid-slotted patches for linear polarization is explained. The SMGCs are also used to form circularly polarized grid-slotted patch antennas and produce wide axial ratio bandwidth.

Chapter 6 concludes by summarizing the research findings in the previous chapters. The advantages and disadvantages of modified Gielis curves are also discussed. Some ways to extend the usage of modified Gielis to other antenna structures are discussed. Few possible ways to extend the design of modified Gielis curves to 3D are also presented.

The flow chart representing the outline of this thesis is shown in Fig. 1.3.

Figure 1.3: Flow chart showing the outline of the thesis

In general, an antenna array is usually a combination of antennas fed by a power divider. Hence, the chapters were arranged such that SMGCs are first applied to individual antenna types to improve their performance. Then they are applied to the components

in a four-way power divider that can be used to feed an antenna array. Then, the SMGCs are applied to periodic structures like FSS and then in antennas with an arrangement of patches in an array form as in grid-slotted patch antenna. The advantage of using complex shapes generated by SMGC and its significant impact is shown in Table 5.3.

Chapter 2

Literature Review

Planar antennas are among the widely used category of antennas because of their simple and low-cost designs while having the ability to meet the requirements of various applications involving different polarizations, single or multi-band operations, narrow, wide or ultra-wideband characteristics, band notching and reconfigurability. The planar antenna designs can also be easily integrated into other back-end circuits for signal processing, and together, they make an antenna system. The planar designs for an antenna could be generally categorized as microstrip-feeding based antennas such as patch antennas (1 dielectric layer with copper cladding on top and bottom); co-planar feeding based antennas such as slot antennas (1 dielectric layer with copper cladding on one side); and aperture or proximity-coupled antennas such as stacked patches (many dielectric layers). In general, microstrip-fed antennas such as patch antennas are usually unidirectional since they have a ground plane at the bottom layer. On the other hand, co-planar based antenna designs will usually have bidirectional properties since there is only a single layer of copper cladding and no bottom layer. Generally, in antenna design, the surface current on the conductor patches determine the radiation behaviour and the input impedance of the antenna.

A study about an arbitrarily shaped microstrip patch [18] is revisited to understand

Figure 2.1: (a) Geometry of arbitrary shaped microstrip resonator with multiple layers and (b) Local coordinates associated with a triangular element. The element number subscripts have been omitted [18].

the relation between the surface currents and antenna behaviour. An analysis has been done on the effects of the substrate, which may contain any number of planar, uniaxially anisotropic, dielectric layers. They are incorporated in the formulation using Green's functions in space or spectral domain [18–23]. The current distribution on the conducting patch is approximated in terms of vector basis functions defined over triangular elements.

The patch resonator in this discussion is assumed to be planar with negligible thickness, as shown in Fig. 2.1(a), but it can take any arbitrary shape, which is modeled by triangular finite elements given in Fig. 2.1(b). The patch and ground plane are perfect conductors. The formulation is applied to the case, where the single patch is placed on the substrate having thickness h with a ground. It has a superstrate of thickness d .

The patch current expansion can be obtained from a triangular finite element. The nodes of each triangular element are assigned the indices i , j , and k in the counter-clockwise direction as shown. These indices adopt a local index scheme and assume the values 1, 2, or 3 in a cyclic manner. The sides of the n^{th} triangle S_n with an area $A^{(n)}$ are formed from three edge vectors $l_i^{(n)}$ oriented from node j to node k . The i^{th} node of S_n is defined with respect to the global coordinate origin O by a position vector $r_i^{(n)}$. The

patch current distribution on S_n is approximated as

$$\mathbf{J}_S^{(n)}(\rho) = \sum_{i=1}^3 I_i^{(n)} \Lambda_i^{(n)}(\rho) \quad (2.1)$$

where $I_i^{(n)}$ is the total current leaving the i^{th} edge and $\Lambda_i^{(n)}$ is a vector basis function defined as

$$\Lambda_i^{(n)}(\rho) = \frac{\rho_i^{(n)}}{2A^{(n)}}, \quad \rho_i^{(n)} = l_k^{(n)} L_j - l_j^{(n)} L_k \quad (2.2)$$

where L_j or L_k is the area coordinate associated with the j^{th} or k^{th} node, respectively.

Therefore, from the above-mentioned discussions, it can be inferred that the antenna radiation is greatly influenced by the current distribution in the conductor patch and also the feeding probe. The arbitrary shapes of the conductor patches with multiple layers also can affect the surface current distribution, which influences the radiation performance, impedance matching, and bandwidth of planar antennas. Throughout the thesis, commercial full-wave simulation software has been used to analyze electromagnetic structures based on finite element method (FEM) or finite-difference time-domain (FDTD) method [24, 25]. Although we have not investigated these methods in detail for the present thesis, some of their developments can be referred to the research works in our group [26–50].

The next section surveys different planar antennas with different types of feeding and radiation properties so that the conductor patches can later be modified based on the Gielis curves.

2.1 Planar antennas design for unidirectional and bidirectional radiation

2.1.1 Unidirectional CPW-fed monopole antenna

Figure 2.2: Unidirectional cpw-fed monopole antennas. (a) Circular patch monopole [52]. (b) Rectangular monopole with L-shaped reflector [53]. (c) Circular reflector for circular monopole [55].

Unidirectional radiation can be achieved in coplanar designs in the form of endfire radiation using planar reflectors placed adjacent in the same plane as the radiating patch. Some of the planar antennas having a monopole with reflector for unidirectional radiation is shown in Fig. 2.2 and in Table 2.1. Introducing slots in the radiating patch and the ground of a UWB antenna can provide dual-band and tri-band rejections for band notching techniques [52]. A coplanar-waveguide (CPW)-fed monopole antenna has been made compact by employing a simple rectangular radiator. The directivity has also been improved by using a planar reflector attached to a ground plane [53]. Another directional ultra-wideband antenna has been designed by feeding with a microstrip line, while the T-slots are inserted diagonally at the corners of the parabolic ground plane in order to increase the directivity [54]. An extended curve-shaped asymmetric CPW ground plane has been used for achieving a stable directional radiation pattern along with a UWB performance. The center of the radiating circular patch is slightly off-centered from the feeding line [55]. The gain of a planar UWB Vivaldi antenna has also been improved by

Table 2.1: Unidirectional CPW-fed monopole antennas with reflectors

Ref.	ε_r	f_l (GHz)	f_h (GHz)	FBW(%)	L (λ_g)	W(λ_g)	G _{max} (dB)	Type
[53]	3.5	0.68	0.98	36.14	0.17	0.1	2 to 5	reflector
[54]	4.3	3.4	10	98.5	0.92	0.92	2 to 5	reflector + T-slots
[55]	4.4	2	20	163.6	0.48	0.3	2 to 5.5	reflector

placing corrugation at the flares and adding grating elements in the area of the tapered slot [56]. Hence adding a planar reflector and parasitic strips can collectively improve the gain while adding Gielis curve-shaped radiating patch can provide additional surface currents.

2.1.2 Bidirectional CPW-fed slot antenna

Slot antennas have been predominantly used in wideband applications, and the shape of the slot influences the surface current distribution, which in turn affects the radiated fields across the slot. Some common planar slot antennas are shown in Fig. 2.3 and in Table 2.3. Circular shaped slot antenna, which is fed by a circular shaped patch

Figure 2.3: Bidirectional coplanar slot antenna designs. (a) Circular slot [57]. (b) Pentagon shaped slot [58]. (c) Elliptical slot with gear-shaped tuning stub [59].

excited using coplanar waveguide feed, has been shown to offer larger bandwidth [57]. Another compact CPW-fed UWB monopole slot antenna with pentagon-shaped ground slot has been used as a UWB radiator, in which a circular disc is connected to a feeding

Table 2.2: Bidirectional coplanar slot antenna designs

Ref.	ε_r	f_l (GHz)	f_h (GHz)	FBW(%)	L (λ_g)	W(λ_g)	G _{max} (dB)	Type
[57]	3.48	2.3	13.9	143.2	1.14	1.14	2 to 8.5	circular reflector
[58]	4.4	3.1	10.7	110.1	0.42	0.42	NA	discontinuous pentagon slots
[59]	4.4	2.5	13.6	137.8	0.54	0.49	NA	modified tuning stub + circular slot
[59]	4.4	2.7	13	131.2	0.53	0.5	NA	modified tuning stub + elliptical slot

stub and is encapsulated by the ground plane [58]. Using circular/elliptical slots with a modified circular tuning stub and a notched ground plane can help to provide UWB characteristics. The modified circular tuning stub is gear-shaped and can help in creating additional surface current paths in the antenna [59]. Hence the addition of protrusions along the edge of the circular shaped patch influences the surface current distribution.

2.1.3 Bidirectional microstrip-fed monopole antenna

Table 2.3: Bidirectional microstrip-fed monopole antenna designs

Ref.	ε_r	f_l (GHz)	f_h (GHz)	FBW(%)	L (λ_g)	W(λ_g)	G _{max} (dB)	Type
[60]	5.4	2	8	120	0.43	0.47	NA	elliptic patch
[61]	4.4	2.6	11	122	0.43	0.29	1 to 5	slots on elliptic patch
[62]	4.43	2.8	11	118.8	0.49	0.53	2 to 6.5	slot on ground
[63]	3	2.85	11.9	122.4	0.37	0.26	0 to 3	linear segments of patch
[64]	4.4	1.5	10	147.8	0.65	0.37	2 to 5.5	pixelated patch
[65]	2.2	2.8	11.9	123.8	0.42	0.37	2 to 5	rounded rectangular patch

Planar wideband monopole antennas have been widely used due to their compact size, low profile, and relatively easy fabrication. Different shapes have been adopted for the monopole to obtain wideband resonance frequencies and impedance matching. Some common microstrip-fed monopole are shown in Fig. 2.4 and in Table 2.3. In [60], circular, and elliptical monopoles have been studied, and their current distributions decide the modes corresponding to different resonant frequencies. In [61], circular slots have been etched out from the edge of the circular monopole as they help in adding extra

Figure 2.4: Omnidirectional or bidirectional microstrip-fed monopole antenna designs. (a) Overlapping subpatches [64]. (b) Elliptical patch with L-shaped slot in ground [62]. (c) Circular patch and ground with slots [61]. (d) Rounded patch [65].

resonant frequencies due to the increased perimeter. Changes have also been made in the ground plane by introducing slots along the top edge of the ground for an elliptical monopole antenna to minimize the ground-plane effects [62]. In [63], a cactus-shaped monopole antenna has been proposed with adjustable return loss for implementation of various ultra-wideband (UWB) sub-band designs, simply by adjusting the length of three linear segments. The current distributions among the three segments determine the resonance and matching. Also, a wideband monopole printed antenna has been designed based on overlapping sub-patches to form complex shapes to get the desired performance [64]. In [65], the antenna design is composed of rectangular and circular patches, with a name dubbed as ‘fat monopole’. The rectangular edge of the monopole,

close to the ground plane influences the impedance bandwidth and the rounded top patch improves the broadband performance, as the current flow is concentrated on the conductor edges. In general, various geometries that form the outline of the monopole would have an impact on the frequency of operation and impedance matching.

2.1.4 Bidirectional antipodal bowtie antenna

Figure 2.5: Bidirectional microstrip or coaxial-fed antipodal bowtie antenna. (a) Bowtie with triangular patch extended [70]. (b) Overlapped bowtie with asymmetric arms [68]. (c) Y-shaped bowtie [66]. (d) Semicircular bowtie patch with square and circular slots [67].

Table 2.4: Bidirectional microstrip-fed antipodal bowtie antenna designs

Ref.	ϵ_r	f_l (GHz)	f_h (GHz)	FBW(%)	L (λ_g)	W(λ_g)	G _{max} (dB)	Type
[66]	2.6	0.75	2.5	107.7	0.5	0.5	2 to 6	Y-shaped bowtie
[67]	4.4	3.1	10.6	109.5	0.5	0.5	0 to 3.5	semicircular bowtie
[68]	2.2	0.8 1.7	0.96 2.5	18.2 38.1	0.44	0.26	1.5 to 5	dualband, overlapped asymmetric bowtie
[69]	4.4	3.1	10.6	109.5	0.54	0.34	NA	overlapped bowtie
[70]	6.15	3.1	10.6	109.5	0.7	0.7	2 to 3.5	unbalanced to balanced feed with stepped impedance

Bowtie antennas have been used extensively for ultra-wideband (UWB) applications, as they are planar projections of the traditional bi-cone antennas. Many printed bowtie antenna types have been designed, such as single-sided printed bowtie, double-sided printed bowtie, and with different ways of feeding techniques. The feeding techniques include coaxial, microstrip, coplanar etc as shown in Fig. 2.5 and in Table 2.4. In [66], a Y-shaped bowtie configuration using coaxial feed was presented and compared with the traditional wideband bowtie antenna. Another coaxial fed antenna was created by printing a geometry comprising a combination of semicircle and square patch on both layers with two semicircles trimmed from the ground [67]. Also, a bowtie antenna with asymmetrical arms was designed whereby two arms of the bowtie were overlapped to tune the two operation bands for beam tilting [68]. For a double-sided bowtie structure, the bottom arm may serve as a ground plane for the microstrip line printed on the upper side of the substrate to feed the top arm. It has been shown that the feeding microstrip line performs an impedance transformation to match the 50 Ω feed line and the bowtie junction [69]. Another double-sided printed bowtie antenna with microstrip to stripline feed presents stable characteristics over the UWB frequency band [70]. In general, the shape of the bowtie patch would affect the surface current and impedance matching.

Figure 2.6: Coplanar bowtie slot antennas. (a) Conventional [71]. (b) Tapered tuning stubs [72]. (c) Rounded corners [73]. (d) Meandered slotlines [78]. (e) Modified slot with triangular metal stub [74]. (f) Passive triangular patch in the slot [75]. (g) Butterfly-shaped slot [76]. (h) Sinusoidally modified metal slot edge [81]. (i) Sliced-edge slot [82]. (j) Leaf-shaped slot [80]. (k) Tuned slot with metal cavity and resistor loading [83]. (l) Differential cpw-fed slot [84]. (m) Symmetric vias [77].

2.1.5 Design of bowtie slots using SMGCs

Bowtie slot antennas (BSAs), which are planar projections of biconical antennas, are commonly used for their relatively wideband operations, low profiles, and bi-directional radiation properties. Some common bowtie slot antennas are shown in Fig. 2.6 and in

Table 2.5: Comparisons of Bowtie Slot Antennas

Ref.	ε_r	f_{low}	f_{high}	FBW (%) ^a	L (λ_g) ^b	W (λ_g) ^b	Gain ^c
[71]	4.7	2.1	3	35.29	0.89*	0.54*	NA ^d
[72]	3.2	1.6	3	60.87	0.79*	0.36*	NA ^d
[74]	4.4	0.4	1.5	115.78 ^e	0.78	0.34	NA ^d (max 6)
[77]	2.33	6	16	90.9	1.19	0.7	5 to 8
[78]	2.2	3.1	10.6	109.5	1.11	0.92	4 to 6
[79]	9.8	2.1	2.9	32	0.3	0.25	NA ^d
[80]	2.17	6	12	66.67	1.07	0.53	4 to 7.5
[81]	2.2	4.75	8	50.98	1.08	0.62	6 to 7.7

^a -10 dB Bandwidth; ^b $\lambda_g \approx c/(f_{\text{low}}\sqrt{(\varepsilon_r + 1)/2})$; ^c Realized Gain (dB); ^d Not Available; ^e bandwidth noted at -8 dB; * Approx. values; f (GHz).

Table 2.5. In [71], a coplanar waveguide (CPW) feed, has been used to excite a conventional triangular bowtie slot antenna (TBSA). Tapered metal stubs have been added in the triangular slot area of the antenna [72] to improve the operating bandwidth without an increase in antenna size. Another work suggests that the performances of three types of bowtie antennas: quadrate, rounded-edge, and triangular shapes, are improved by adding round corners on their radiation surfaces [73]. Such modifications provide better return loss, flatter input impedance, and more stable radiation patterns. Different cases in the design show that placing rounded corners at the sharp vertexes of radiation surfaces can bring positive effects on the performances of planar antennas. Similar triangular metal stubs added in the modified slot aperture with rounded ends have also been discussed [74]. Some other CPW-fed designs involve adding passive patches in the slot area of the bowtie [75] and even altering one-half of the slot edge to form a butterfly shape [76], in order to influence the surface current path. In [77], BSA with two circular vias in its slot area has been shown to have increased antenna bandwidth. Adding a meander shaped slot to the TBSA favors ultra-wideband operation at the expense of increased antenna size [78]. A flower-shaped arrangement of CPW-fed bowtie has been discussed in [79] for obtaining lower profile and slightly better impedance matching. The slot profile, which is modified into the shape of a leaf and coupled-fed by a microstrip

line [80], has shown to have increased the impedance bandwidth. Even sinusoidal curves are used to modify the longer edges of the slot profile in BSA [81] to increase the upper cut-off frequency of the operating bandwidth. Other techniques such as increasing the slot edge length [82], resistor loading [83], and even differential feeding [84] has been shown to improve some characteristics of the bowtie slot antennas. Still, most BSAs, as mentioned above, do not have wide impedance bandwidth, and they may have a larger size.

2.2 Four-way power dividers for antenna arrays

Figure 2.7: Omnidirectional antenna with center-fed power divider to feed elements on horizontal plane. (a) Multiband and dual-polarized omnidirectional antenna [87]. (b) 45° slant-polarized omnidirectional antenna [88]. (c) Horizontally polarized omnidirectional antenna with mutual coupling [89]. (d) Broadband dual-polarized omnidirectional antenna [90]. (e) Circular connected vivaldi antenna array [85, 86]. (f) Horizontally polarized omnidirectional antenna with parasitic strips [91]. (g) Horizontally polarized spiral-shaped antenna [92].

Table 2.6: Power dividers for center-fed antenna arrays

Ref.	ε_r	f_l (GHz)	f_h (GHz)	FBW(%)	D (λ_g)	Ht(λ_g)	G_{max} (dB)	Type	Pol.	HS
[85, 86]	2.55	1.28	11.5	160	1	-	0.5 to 4	circularly connected vivaldi	H	No
[87]	4.4	0.8 1.88	0.96 2.7	18.18 35.8	0.87 -	- 0.518	3 to 5	cone, vivaldi	V H	No
[88]	2.2	1.75	2.18	21.8	0.573 -	- 0.738	<1	crossed dipole	45° S	No
[89]	2.2	1.7	3.5	69.2	1.07	-	< 1	arc dipole, parasitic strips	V H	No
[90]	4.4 1	1.53 0.82	2.95 6	63.3 152	1.17 -	- 0.267	1 to 6	discone, printed dipole	H V	No
[91]	4.4	1.58	3.88	84.2	1.03	-	0 to 1.5	printed dipole, parasitic and director strips	H	No
[92]	4.4	2.4	2.53	5.27	0.69	-	NA	arc shaped printed dipole	H	No

D: diameter, Ht: height, HS: harmonic suppression, H: horizontal, V: vertical, S: slant, NA: not available.

Power dividers (PD) are extensively used in applications involving power splitting based on the number of output ports, the amplitude at the output, isolation, and impedance matching. PDs such as Wilkinson have been used commonly for simple two way PD design and are less complicated. The complexity of the PD design increases with the number of output ports, frequency bandwidth requirement, isolation, power handling capacity, etc. Multi-way power dividers have found interest in applications involving antenna arrays for omnidirectional or multi-beam radiation. In the case of omnidirectional radiation, the antenna elements in the array need to be placed in a circular arrangement, each being fed using a power divider with a coax-feed from the center. Some of the omnidirectional antennas that require center-fed or perpendicular coax-fed input shown in Fig. 2.7 and in Table 2.6. The power divider is in the same plane as the radiating antenna elements on a horizontal plane. Circular connected Vivaldi antenna array [85,86] in a flower pattern with a simple radial power divider is designed. Multiband and dual-polarized omnidirectional antenna are designed using vertical monopole element and horizontal Vivaldi element [87]. A 45° slant-polarized omnidirectional antenna [88] is designed using

Figure 2.8: Cavity based four-way power dividers. (a) Rectangular waveguide divider/combiner [95]. (b) Chained quasi-planar divider [96]. (c) Circular waveguide [94]. (d) Substrate-independent power divider [97]. (e) Radial unequal divider [98]. (f) Multilayer SIW out-of-phase divider [99]. (g) Multilayer SIW with slot coupling [100]. (h) Y-junction SIW divider [101].

a crossed dipole that includes a balun in the feeding network. A horizontally polarized omnidirectional antenna with a center-fed power divider to feed the dipoles with parasitic strips is discussed in [89]. Broadband dual-polarized omnidirectional antenna made comprising of disc cone and dipoles with parasitic patches [90] are also proposed. Another horizontally polarized omnidirectional antenna with parasitic strips having a balun to feed slot-based elements [91] is also proposed. Horizontally polarized spiral-shaped antenna [92], where the spirals are loaded with conductor patches and a simple power divider, is designed which is fed by a perpendicular coax-fed input.

Some four-way PDs have been designed using waveguides or cavities that are center-fed or end-fed for applications requiring low insertion loss, wide bandwidth, high-power, and even dual-polarization operation [93–95]. A few cavity-based power divider using waveguide or SIW are shown in Fig. 2.8 and in Table 2.7. A four-way circular waveguide power divider [94] that is center-fed and involves an input scatterer, metallic post, output scatterer to support dual polarization applications has been designed. A Chained quasi-planar divider [96] consists of a microstrip line that is tapped for output in series. A Substrate-independent power divider [97] enclosing air-cavity is used to design a four-way

Table 2.7: Cavity based four-way PD designs

Ref.	ε_r	f_l (GHz)	f_h (GHz)	FBW(%)	L (λ_g)	W(λ_g)	Type	Feed	HS
[96]	1	4	5.8	36.7	2.97	0.1	quasi-planar, SIW	EF	No
[97]	1	5	5.8	14.8	3.3	3.3	air-filled SIW	CF	No
[98]	2.2	9	12	28.57	1.89	1.89	radially guided SIW	CF	No
[99]	2.33	8.9	10.8	19.29	3.2	1	multilayer, halfmode SIW	EF	No
[100]	3.5	8.75	10.8	20.5	1.85	0.85	multilayer SIW	EF	No
[101]	4.3	8.6	12.2	34.6	2.8	2.3	halfmode SIW	EF	No

CF: center-fed, EF: edge-fed, HS: harmonic suppression

power divider, where all the ports (input and outputs) are perpendicular to the substrate. Another cavity-based power divider with a combination of a radial waveguide and microstrip to SIW transition is used to achieve unequal power division [98]. A Multilayer SIW out-of-phase divider [99] has been proposed for a two-way power division where all the ports are end-fed. Also, a different multilayer SIW with slot coupling [100] with isolation resistors have been designed with an end-fed configuration. A Y-junction SIW-based [101] four-way power divider with an end-fed design has been proposed. However, waveguide-based PDs are not suitable for low-frequency operations and not compatible with planar designs.

Table 2.8: Microstrip based PD designs

Ref.	ε_r	f_l (GHz)	f_h (GHz)	FBW(%)	L (λ_g)	W(λ_g)	Type	Feed	HS
[102]	2.22	5	12.5	85.7	0.42	0.42	stripline, low-impedance line	EF	No
[103]	2.2	9.4	14.8	44.6	0.91	0.91	symmetrical disc	EF	No
[105]	4.3	2.4	3.8	45.1	1.3	1.3	radial sectorial arrangement	CF	No
[106]	3.38	5.5	6.4	15.1	1.08	1.08	sector patch + radial notch	CF	No
[107]	4.3	0.9	0.93	3.2	0.55	0.55	stellate structure + star slot	CF	No
[108]	4.5	0.75	1.25	50	0.35	0.35	radial, multiway RLC components	CF	No

CF: center-fed, EF: edge-fed, HS: harmonic suppression

To overcome the shortcomings of a cavity or waveguide-based power dividers such as

Figure 2.9: End-fed and center-fed microstrip based multiway power dividers. (a) Disk with stepped impedance matching [102]. (b) Microstrip disk [103]. (c) Circular microstrip disk divider / combiner [104]. (d) Sectorial components [105]. (e) Broadside-coupled filtering divider [106]. (h) Stellate structures [107]. (g) Radial four-way divider [108].

high-frequency and complex fabrication, microstrip planar structure-based four-way PDs on the printed circuit boards have been designed. Some end-fed and center-fed microstrip based multiway power dividers are shown in Fig. 2.9 and in Table 2.8. A simple design that involves a conventional disk-shaped patch that radially extends to the output ports with stepped impedance matching [102, 103] has been proposed. The circular microstrip disk divider/combiner [104] also operates on a simple radial power division. Some sectorial patch shaped transitions are added [105] along with some slots on the ground plane [106] for broadside-coupled filtering divider has been done. A stellate structure [107] that consists of star-shaped slots extending on the ground plane is used to obtain an eight-way power divider. Another radial four-way divider [108] that consists of air-bridges and slots in the ground plane to improve isolation is also proposed. However, such simple center-fed power divider suffers from harmonics, where some harmonic frequencies may be fed to the antenna arrays from the power divider.

Many end-fed power dividers have been proposed for harmonic suppression, but not many designs have been proposed for four-way center-fed power divider with harmonic

Figure 2.10: Harmonic suppression in two-way and four-way power dividers. (a) Traveling wave divider / combiner [109]. (b) Ring resonator [110]. (c) Using coupled lines [116]. (d) Coupled loops [118]. (e) Dual-band arbitrary power division [133]. (f) Stepped-impedance transformers [111]. (g) Mirrored coupling structures [112]. (h) Three parallel coupled lines [113]. (i) Four-way reconfigurable [114].

Table 2.9: End fed PD with harmonic suppression

Ref.	ϵ_r	$f_l(\text{GHz})$	$f_h(\text{GHz})$	FBW(%)	$L(\lambda_g)$	$W(\lambda_g)$	Type	Feed	HS
[109]	9.8	5	12	82.3	1.35	0.58	travelling wave	EF	No
[110]	4.8	1.41	2.68	62	0.32	0.24	ring resonator + stub	EF	Yes
[111]	2.2	0.5	1.5	100	0.21	0.21	stepped impedance transformers	EF	No
[112]	3.55	5	7	33.33	1	0.75	mirror coupling structures	EF	No
[113]	3.5	2.09	5.27	86.41	0.4	0.27	3 parallel coupled lines	EF	No
[114]	2.94	4.9	5.15	4.97	1.14	1.14	reconfigurable PD with p-i-n diodes	EF	No
[116]	3.5	0.83	1.39	50.45	0.24	0.12	stubs and coupled line loops	EF	No
[118]	2.2	2.3	2.45	6.3	1.16	1.16	filtering PD using hair pin resonator	EF	No

CF: center-fed, EF: edge-fed, HS: harmonic suppression

suppression. A few harmonic suppression techniques done in planar four way designs are shown in Fig. 2.10 and in Table 2.9. A simple end-fed traveling-wave based five-way power divider [109] has been designed, but many other end-fed power dividers have additional design inclusions to provide harmonic suppression. Some power dividers with ring resonators [110], stepped-impedance transformers [111], etc., could help to improve

Figure 2.11: End-fed two-way or four-way power divider. (a) Multi-conductor coupled lines [119]. (b) Isolation using resistor and capacitor bank [120]. (c) Stub-loaded coupled line [125]. (d) Slot lines and air-bridge resistors [126]. (e) Lumped elements for R , L and C [127]. (f) Frequency-dependent couplings [128]. (g) Arbitrary power ratio [129]. (h) $1/32^{nd}$ mode SIW resonator [130]. (i) Looped coupled-line structures [131]. (j) Varactor-tuned LC resonators [132].

harmonic suppression. A few designs include mirrored coupling structures [112], three parallel coupled lines [113], four-way reconfigurable [114] to enhance harmonic suppression or bandwidth characteristics of the power divider. Other characteristics such as selectivity of pass band or stop band have been further enhanced by using coupled lines [115–117] or looped coupled lines [118].

Many other techniques have also been proposed in the design of two-way or four-way PD based on using Wilkinson PD for wideband harmonic suppression. The PDs are

Table 2.10: End fed two-way or four-way PD

Ref.	ϵ_r	$f_l(\text{GHz})$	$f_h(\text{GHz})$	FBW(%)	L (λ_g)	W(λ_g)	Type	HS
[119]	4.5	1.4	3	72.7	0.19	0.15	multi conductor coupled line	No
[120]	3.55	1.45	1.55	6.66	0.69	0.51	impedance transformer	No
[125]	2.54	1.5	4.2	94.7	0.13	0.097	frequency selective coupling structure	Yes
[126]	2.95	4	9.6	82.3	0.46	0.37	slotline	No
[127]	3.66	0.8 1.8	1.25 2.2	43.9 20	0.26	0.22	decomposing single dual-band transformer	No
[128]	3.5	1.92	2.08	8	0.63	0.47	frequency dependent couplings	No
[129]	4.5	0.76	1.6	66.6	0.66	0.6	arbitrary, transmission line $< 1/4\lambda_0$	No
[130]	3.55	2.1	2.8	28.5	0.63	0.63	SIW	No
[131]	10.2	1	2	66.6	0.31	0.31	looped coupled line structures	No

HS: harmonic suppression

mostly end-fed as shown in Fig. 2.11 and in Table 2.10. A multi-conductor coupled lines arrangement [119] with air-bridge and using isolation resistor and capacitor bank [120] have been proposed for improving isolation and harmonic suppression. A stub-loaded coupled line structure [125], which acts as a frequency coupling structure, is proposed for wideband harmonic suppression. Some slot lines are included in the ground plane, and air-bridge resistors [126] are also proposed for harmonic suppression. Lumped elements such as R, L, and C components are used at the junctions where the conductors divide in the power dividing structure [127]. Frequency-dependent couplings [128] are designed using coupled-line combinations that can provide better selectivity. Other characteristics such as power division with arbitrary power ratio [129] and equal power division can be obtained from $1/32^{nd}$ mode SIW resonator [130]. Looped coupled-line structures [131] are also proposed as a design to improve the selectivity. Some lumped components, such as varactor-tuned LC resonators [132], have been proposed for reconfigurability.

The design requirements for an omnidirectional or multi-beam antenna array requires a power divider that is center-fed or perpendicular fed. The power divider should also be able to suppress harmonics at frequencies out of the band if the application requires

antenna arrays to avoid radiation at harmonic frequencies to meet some electromagnetic compatibility standards. The above-discussed power dividers mostly are end-fed, and those that are center-fed do not exhibit harmonic suppression characteristics.

2.3 Periodic or array structures

2.3.1 Three-dimensional bandstop frequency-selective structures

Figure 2.12: Frequency selective surfaces. (a) Geometric shapes for FSS unit cells [138]. (b) Capacitive patches and inductive wires [139]. (c) Single and double loops [144]. (d) Miniaturized ring loop [145]. (e) Slot loaded patch [146]. (f) Quasi-elliptic frequency response [143]. (g) Patch with meander connections [147].

Table 2.11: 2D and 3D frequency selective surface designs

Ref.	ε_r	$f_l(\text{GHz})$	$f_h(\text{GHz})$	FBW(%)	L (λ_g)	W(λ_g)	Type
[139]	10.2	5.1	6.1	17.8	0.32	0.16	sub wavelength periodic structures
[141]	2.2	9	11	20	0.45	0.45	circular loop
[142]	1	7.1	7.2	1.4	0.71	0.71	3D cubical unit cell
[143]	8	5.4	6.2	13.8	0.35	0.35	3D, quasi-elliptic frequency response
[144]	3	1.03	1.04	0.96	0.34	0.38	circular loop
[145]	4.3	8.5	11.5	30	0.32	0.32	miniaturized ring
[146]	2.2	5.2	6.2	17.5	0.21	0.21	I slot, multiband
		8.7	10	13.9			
		13.5	14	3.63			

Frequency selective surfaces (FSS) have been widely used for applications such as filtering the incident electromagnetic waves in space depending upon the angle of incidence [140]. Many 2-D and 3-D FSS have been developed for bandstop or bandpass applications as shown in Fig. 2.12 and in Table 2.11. A 3-D FSS comprising a 2-D periodic array of multimode cavities has been proposed for performing a bandpass operation over the desired frequency that can be obtained by controlling the coupling with air [141]. Even cubical unit cells with hollow cavity and unit cells with two propagation paths have been discussed for bandpass filtering applications [142, 143]. Also, circular and elliptical loop resonators have been used to form thin 3-D FSS structures for obtaining single and dual-band bandstop operations [144]. It can be observed that the perimeter of the loop decides the operating frequency. Therefore choosing a proper shape profile of the loop in a unit cell for a fixed unit cell size can obtain a relatively lower operating frequency than a conventional circular loop. A special set of curves called Gielis curves have been used previously in metamaterial unit cell design, where it is shown that Gielis curves can further reduce the electrical size of the split-ring and complementary split-ring resonators.

2.3.2 Periodic leaky-wave antenna

Table 2.12: Periodic leaky wave antenna designs

Ref.	ε_r	$f_l(\text{GHz})$	$f_h(\text{GHz})$	FBW(%)	L (λ_g)	W(λ_g)	Type
[148]	2.2	4.8	6.8	34.4	0.64	0.4	rectangular patch unit cell
[149, 150]	2.42	2.2	-	-	0.38	0.57	rectangular patch unit cell with tapered width
[151]	4.4	4.5	7.5	50	0.98	0.17	supercell with tapered patches
[152]	10.2	12.1	16	27.7	0.57	0.62	non identical elements per unit cell

Planar leaky-wave antennas (PLWAs) shown in Fig. 2.13 and Table 2.12 have been widely used in broadband wireless communications and radar applications due to their advantages like low-profile, simple design, and the ability to produce narrow beams that can point either in the backward and forward direction [12]. Some PLWAs designs consist

Figure 2.13: Periodic leaky-wave antennas. (a) Series-fed uniform rectangular patch array [148]. (b) Series-fed non-uniform rectangular patch array [149, 150]. (c) Supercell with three periodic unit cells [151].

of a conventional rectangular-shaped patch (CRP) element as a radiating element in the unit cell of the periodic leaky-wave antenna structure [148–150]. Periodic leaky-wave antennas can also be designed using supercells instead of conventional unit cells. These supercells will comprise of more than one unit cell based on the required performance. Supercell with three periodic unit cells [151], two non-identical elements per unit cell [152]. The width of the RMP elements influence the surface currents and thereby can influence the radiation behavior of the PLWA. Hence changing the shape of the patch in the unit cell can improve the PLWA performance.

2.3.3 Linearly polarized grid-slotted patch antenna

Many antennas with grid-slotted patches, also known as metasurface structures, have been widely used to enhance the antenna characteristics such as bandwidth, gain, radiation pattern, etc., as shown in Fig. 2.14 and in Table 2.13. A grid-slotted patch array has been used as a metasurface over a slot fed by a microstrip line to enhance bandwidth or

Figure 2.14: Linearly polarized antennas with metasurfaces. (a) 4×4 array of square patches [154]. (b) Square patch array tilted diagonally [155]. (c) Hexagonal grid-slotted patch [156]. (d) Shorted square patch array (mushroom) [157]. (e) Patch array with varying width [158].

Table 2.13: Linearly polarized grid slotted patch antenna designs

Ref.	ϵ_r	f_l (GHz)	f_h (GHz)	FBW(%)	L (λ_g)	W (λ_g)	G_{max} (dB)	Type
[154]	3.38	4.6	6.2	29.6	1.36	1.36	6 to 9.8	grid-slotted square patch
[155]	3.38	4.6	6.5	34.23	1.24	1.24	7.7 to 10.3	diagonally tilted grid-slotted patch
[156]	2.2	4.89	6.89	33.9	1.23	1.23	9 to 11	hexagonal grid slotted patch
[157]	3.38	4.85	6.28	25.6	1.43	1.43	8 to 11	grid-slotted mushroom cells
[158]	10	1.85	5.65	101.3	1.56	0.91	2.1 to 7.32	index gradient patches

reduce the lower cut-off frequency [154]. To understand the underlying working principle of such metasurface structures, characteristic mode analysis (CMA) has been used as an alternate method [155] to observe the modes. Since the geometry or shape of the patches in the metasurface influences the behaviour of such structures, even another shape such as hexagon has been used for the patches in the metasurface to enhance

antenna characteristics [156]. Shorted square patch array is being excited by a slot [157] and a patch array with varying width on the same plane as the bowtie launcher [157, 158] for linear polarization has been proposed.

2.3.4 Same-sense circularly polarized grid-slotted patch antenna with wide axial ratio bandwidth

Circularly polarized (CP) antennas are widely known for their applications in satellite communications, GPS, GNSS, etc., and some of those antenna types are shown in Fig. 2.15. Bidirectional CP antennas, which are prominently used in microcellular base station relays, WLAN, indoor wireless access, etc., require same-sense CP radiation in forward and backward direction. Though many works have been done on antennas with opposite-sense CP radiation [159, 160], very few works have been done in CP antennas with same-sense as in Table 2.14. Other techniques to generate CP such as using spiral conductors [161], four radiating arms connected to a reconfigurable feeding network [162], inverted L-shaped parasitic strip at the open end of a microstrip line extension, and a slot modification [163], curved tapered slot in the form of ‘S’ [164] have been proposed. Some non-planar designs have been proposed where two rectangular metal strips are fed in phase, and they are placed at an inclination angle of 45° inside a metallic waveguide with a square aperture to produce a same-sense CP radiation [165–167]. Also, using an array of orthogonally placed dipole antennas fed by composite right/left-handed transmission lines (CRLH-TL), a same-sense CP radiation can be obtained [168]. Similar work with more dipoles in the array and special substrates [169] have been done. Many planar designs, including a single substrate layer with two closely spaced complementary dipoles in back-to-back configuration connected with a phase delay line for creating same-sense CP radiation, have been proposed [170].

The polarization for a CP antenna can be reconfigured between LHCP and RHCP radiation, based on the switching of the electrical dipoles while having a fixed mag-

Table 2.14: Comparison with previous same-sense circularly polarized planar antennas

Ref.	ε_r	f_l, f_h	IBW ^a	$L \times W \times H(\lambda_g)^b$	Gain ^c	f_l, f_h	ARBW
[170]	2.65	5.72, 5.88	2.76	$0.62 \times 0.7 \times 0.04$	4	5.72, 5.88	2.76
[171]	2.2	5.7, 5.9	3.44	$2.04 \times 1.62 \times 0.03$	5	5.7, 5.9	3.44
[172](1)	2.5	2.44, 2.54	4	$0.97 \times 0.97 \times 0.06$	NG	2.46, 2.5	1.61
[172] (2)	2.55	2.35, 2.46	4.57	$0.67 \times 1.01 \times 0.04$	NG	2.34, 2.45	4.56
[173]	2.5	2.44, 2.54	4.02	$0.97 \times 0.97 \times 0.06$	4 to 5	2.46, 2.51	2.01
[174]	4.4	2, 2.6	26.08	$0.78 \times 0.78 \times 0.32$	5.6	2.38, 2.57	7.67
[175]	4.4	2.28, 2.64	14.63	$0.94 \times 0.62 \times 0.39$	~ 5	2.3, 2.47	7.12
[176]	4.5	1.2, 1.65	31.58	$0.93 \times 0.93 \times 0.07$	2	1.54, 1.57	1.92
This work	4.1	4.05, 5.97	38.32	$0.95 \times 0.85 \times 0.2$	2 to 3.5	4.07, 5.02	21

f_l : lower cut-off; f_h : higher cut-off; ^a -10 dB Impedance bandwidth (IBW); ^b $\lambda_g \approx c/(f_{low}\sqrt{(\varepsilon_r + 1)/2})$; L : length; W : width; H : height; ^c Realized Gain (dBic).

Table 2.15: Circularly polarized grid slotted patch antenna designs

Ref.	ε_r	f_l (GHz)	f_h (GHz)	FBW(%)	L (λ_g)	W(λ_g)	G _{max} (dB)	Type
[178]	3.38	4.9	6.2	23.4	0.77	0.77	7 to 7.6	truncated corner square patch
[179]	3.38	4.2	6.7	45.8	1.09	1.09	6 to 8	grid slotted patches shorted diagonally
[180]	4.4	4.9	5.9	18.5	0.92	0.75	5 to 7	oblique slot excitation
[181]	2.2	5.8	8.2	34.2	1.46	1.46	10 to 10.9	shorted grid patches with L-slots
[182]	2.65	2.48 3.48	2.56 3.63	3.17 4.21	0.82	0.78	6 to 7	diagonal slot on grid slotted patch

netic dipole [171]. Some multi-layer designs involve back-to-back patches with a common ground and crossed slots for same-sense CP radiation [172, 173]. An H-shaped aperture is used to excite dual rotated patches [174] to generate same-sense CP. A bending strip, when placed above a narrow slot and flipping the strip below the slot, has been shown to give a same-sense CP radiation [175]. A frequency selective surface based cross-slot unit cell has been used as a polarization converting surface fed by a bidirectional CP slot antenna to convert the sense of CP wave in the backward direction, thereby making it same-sense [176]. However, the above-mentioned antennas provide narrow axial ratio bandwidth (ARBW). Metamaterial-based grid-slotted patch antennas have been proposed for directional linearly polarized [157, 177] and circularly polarized antennas to improve gain or impedance and AR bandwidths. The grid-slotted patches are connected

Figure 2.15: Circularly polarized antennas with metasurface. (a) Rectangular patch array fed by diagonal slot [180]. (b) Shorted patch array fed by L-shaped slot [181]. (c) Patch array with parasitic patch [179]. (d) Patch array with slots [182]. (e) Patch array fed by truncated slot [178].

to ground through via in some designs and are excited by a truncated patch or a slanted slot or L- and Z-shaped slots [178–181]. Some patch array with diagonal slots are used as metasurface for dualband circular polarization [182] given in Table 2.15. But not many grid-slotted patch designs have been reported for bidirectional same-sense CP radiation.

2.4 Gielis curves in the antenna designs

A type of closed-form formula has been used for designing the shape of antenna, dielectrics or metamaterials and for enhancing their performance by exploring unconventional dimensions and geometry. This closed-form formula is called the Gielis curve, given by Johan Gielis [183]. In general, circle, square, ellipse, and rectangle are all members of the set of superellipses defined as

$$|x/a|^n + |y/b|^n = 1 \quad (2.3)$$

Rotational symmetry (m)	$n_1=n_2$ $=n_3=1$	$n_1=1$ $n_2=n_3$	$n_1=n_2$ $=n_3=0.5$	$n_1=30$ $n_2=n_3=15$	$n_1=80$ $n_2=25, n_3=100$	$n_1=n_2$ $=n_3=0.5$ $a=2$
0		2
1		500
2		500
3		1980
4		1000
5		620
6		390
7		320
8		250

Figure 2.16: Various forms of Gielis shapes based on the parameter values [183].

The main disadvantage of the superellipses is their limited symmetry. The use of the polar coordinates $r = f(\phi)$ by substituting $x = r \cos \phi$ and $y = r \sin \phi$, and the introduction of argument $m/4$ of the angle ϕ introduces specific rotational symmetries. The rotational symmetry leads to a different equation where n_i and $m \in \mathbb{R}^+$ (positive real numbers) and for $a, b \in \mathbb{R}_0^+$ (positive real numbers but not zero) expressed as

$$r(\phi) = \frac{1}{\left[\left(\left| \frac{1}{a} \cos \left(\phi \frac{m}{4} \right) \right| \right)^{n_2} + \left(\left| \frac{1}{b} \sin \left(\phi \frac{m}{4} \right) \right| \right)^{n_3} \right]^{1/n_1}} \quad (2.4)$$

Some of the Gielis shapes generated based on the values of various parameters are shown in Fig. 2.16. These shapes from Gielis can be used in the designs of antennas or microwave circuit design, as they offer additional features in geometry, which can help to have better control over the antenna characteristics.

Figure 2.17: Gielis shapes in antenna design. (a) Dielectric resonator [184, 185]. (b) Bowtie patch [186]. (c) Saw-tooth edged patch [187]. (d) Characteristic mode analysis of Gielis patch [188]. (e) Various metamaterial unit cells [191]. (f) Flower-shaped monopole [189]. (g) Gielis in SRR and CSRR [190].

Some of the antenna types where the Gielis shapes are used in the antenna design are shown in Fig. 2.17. A dielectric resonator (DRA) having a Gielis-shaped contour is studied in [184, 185]. The Gielis-shaped DRAs can also be used in the generation of linearly and circularly polarized waves, while they exhibit wideband performance in terms of input impedance matching, radiation patterns, realized gain, and polarization properties. The proposed class of antennas shows wideband matching and stable patterns. In [186], the Gielis-shape formula is used to describe the optimum bowtie antenna design. The antenna with balanced-feed exhibits super wideband characteristics, and pattern stability is observed over the whole operating frequency range. A compact coplanar waveguide (CPW)-fed patch antenna for UWB operation using Gielis-shape results in a circular patch with sawtooth-like circumference [187]. In comparison to the conventional CPW-fed- circular patch antenna whose optimal diameter is 25 mm, the circular patch of the proposed design has a diameter of 18.2 mm. The antenna has an improved group delay, with similar radiation and gain characteristics. Even some analysis, such as the Characteristic Mode analysis of a drop-like Gielis-shaped patch antenna has been done [188]. The eigenanalysis is performed to determine the chracteristic modes through the

equivalent magnetic characteristic currents on the printed antenna aperture surface of the cavity-like subdomain of the simulated structure. In [189], the design and analysis of nature-inspired flower-shaped UWB antenna are shown. The contour profile of the antenna is obtained from the Gielis formula. The antenna works in the UWB range (3.1 GHz – 10.6 GHz) with VSWR less than 2. Also, it exhibits good advantages of being compact and having an almost constant gain and group delay. Gielis-shapes are also used for split-ring resonators (SRR), complementary split-ring resonators (CSRR) [190], and also in metamaterials to achieve double-negative refractive index with supershaped unit cells that optimize the Gielis parameters through evolutionary algorithms [191]. These structures show a 12% reduction of the resonance frequency for the same maximum radius, and close to 30% when the same unit cell area is considered.

Hence, by altering the shapes of the conductor patches or dielectric resonators, the performance of the antennas and dividers can be controlled or improved to a large extent. The modifications of the useful Gielis curve to avoid its shortcomings and their usage in antennas, power dividers, and antenna array type structures will be discussed in the forthcoming chapters.

Chapter 3

Improved Designs of Antennas Using Modified Gielis Curves

3.1 Introduction

The radiation characteristics and performance of any antenna depends on how the surface currents are distributed on the conductors involved in the antenna. Hence, the shape of the conductors in the design can greatly influence the surface currents and impedance matching, especially in planar antennas like microstrip or CPW-fed antennas.

3.2 Modified Gielis curves

In order to realize a complex shape for any conductor or slot profile using SMGC, firstly, the conventional Gielis curve described by [183] is revisited.

$$r(\theta) = \left(\left| \frac{\cos\left(\frac{m_1\theta}{4}\right)}{a} \right|^{n_2} + \left| \frac{\sin\left(\frac{m_2\theta}{4}\right)}{b} \right|^{n_3} \right)^{-\frac{1}{n_1}} \quad (3.1)$$

where the parameters a and b are used to scale the curve size, while $m_1 = m_2$, n_1 , n_2 , n_3 are used to create the desired curve profile. However, the profiles generated from (3.1)

Figure 3.1: Curves from conventional Gielis curves (a) G^I [$m_1 = 7, m_2 = 7, n_1 = 10, n_2 = 6, n_3 = 6$] and (b) G^{II} [$m_1 = 1, m_2 = 1, n_1 = 0.5, n_2 = 0.5, n_3 = 0.5$] for given ‘ a ’ and ‘ b ’ values.

may not always properly conform to the required shape and size because the conventional Gielis curves are inherently required to satisfy the condition $a = b$ to avoid discontinuities in the profiles. Therefore, the obtained profiles from the conventional Gielis formula can only be uniformly scaled along the X and Y directions. To better understand the disadvantages of the conventional Gielis curves, an example is given in Fig. 3.1. A curve with heptagon-based profile namely G^I can be generated with the Gielis parameters $m_1 = 7, m_2 = 7, n_1 = 10, n_2 = 6, n_3 = 6$ as shown in Fig. 3.1(a). It can be seen that if the parameters a and b are equal i.e., $a = b = 1$ or $a = b = 2$, then the generated curve is continuous and smooth. But when $a \neq b$, the generated curve is discontinuous and broken which is marked by the dotted red circle as indicated in Fig. 3.1(a) for $a = 1$ and $b = 2$. Similarly another curve geometry G^{II} that looks like a leaf can be generated with parameters $m_1 = 1, m_2 = 1, n_1 = 0.5, n_2 = 0.5, n_3 = 0.5$. Again, the curve gets discontinuous or broken when $a \neq b$ as shown in Fig. 3.1(b). Hence, the conventional Gielis cannot properly conform to the required scaling and will have distortions when $a \neq b$.

Figure 3.2: Examples of sectorially modified Gielis curves with (a) G^I with at least two scaling variables equal $S_{X1} = S_{X2}$, (b) G^I with different scaling variables $S_{X1} \neq S_{X2} \neq S_Y$, (c) G^{II} with at least two scaling variables equal $S_{X1} = S_{X2}$, and (d) G^{II} with different scaling variables $S_{X1} \neq S_{X2} \neq S_Y$.

To properly conform to the required curve profile, the modified form of conventional Gielis curve called Sectorially Modified Gielis curves (SMGCs) are proposed for generating many shapes for a patch or slot profile. This sectorially modified equation allows independent scaling along the $+X$, $-X$, and $\pm Y$ directions since the curve is modified based on two sectors given by the following equation.

$$\begin{aligned}
 X = x - c_x &= \begin{cases} r(\theta) \cos(\theta) S_{X1}, & \forall 90^\circ \leq \theta < 270^\circ \\ r(\theta) \cos(\theta) S_{X2}, & \forall 270^\circ < \theta \leq 450^\circ \end{cases} \\
 Y = y - c_y &= r(\theta) \sin(\theta) S_Y, \quad \forall 90^\circ \leq \theta \leq 450^\circ
 \end{aligned} \tag{3.2}$$

where S_{X1} , S_{X2} scale the curve along $-X$ and $+X$ directions respectively, while S_Y scales uniformly along $\pm Y$ directions with reference to the XY axes in Fig. 3.1(a) and Fig. 3.2(a). Thus, the SMGCs can provide additional degrees of freedom in the design. To avoid the scaling problems and discontinuities in the conventional Gielis curve, a and

b are set to 1 and the scaling variables S_{x1} , S_{x2} , and S_y are introduced as external factors depending on left and right sectors. The 360° curve is sectorized into two parts as left half ($-X$) and right half ($+X$), since θ is swept from 90° to 270° for the left half and from 270° to 450° for the right half as in (3.2). The parameters c_x and c_y give the center position for the SMGCs with reference to the origin of xy axes located at the bottom left corner. Using sectorially modified Gielis method for scaling G^I and G^{II} , it can be seen that no discontinuities can be observed when the curve is scaled independently along the X and Y directions as shown in Fig. 3.2. Both the conventional Gielis curves G^I and G^{II} have no discontinuities when the curve is scaled uniformly in both $-X$ and $+X$ directions i.e., $S_{x1} = S_{x2}$ or scaled non-uniformly along $-X$ and $+X$ directions i.e. $S_{x1} \neq S_{x2} \neq S_y$. Hence SMGCs can be used as a versatile formula to generate closed curves or even open curves for shaping a conductor in planar designs.

To further understand the effects of SMGC parameters on its geometry, each parameter is varied, and the behaviour of the shapes is observed.

Figure 3.3: Effect of parameters $m_1 = m_2$ on the geometry ($n_1 = n_2 = n_3 = 1$).

The SMGC parameters m_1 and m_2 are generally set as the same values for the curve to close properly, although there are exceptions in some cases. From Fig. 3.3, it can be seen that by varying $m_1 = m_2$ from 1 to 10, the number of vertices in the curve is increased accordingly. If $m_1 = m_2 = 5$, then there are five vertices in the SMGC curve.

Figure 3.4: Effect of parameters n_1 on the geometry ($m_1 = m_2 = 6$, $n_2 = n_3 = 1$).

The parameter n_1 is varied from 1 to 4 as shown in Fig. 3.4, while fixing $m_1 = m_2 = 6$. It is observed that for $n_1 < 2$, the edges between the vertices are concave in nature. For $n_1 = 2$, the edges or sides are straight and become a perfect hexagon. For $n_1 > 2$, the edges start to take a convex nature.

Figure 3.5: Effect of parameters n_2 on the geometry ($m_1 = m_2 = 6$, $n_1 = n_3 = 1$).

The effect of parameter n_2 is given in Fig. 3.5. It can be seen that the variation of n_2 parameter from 1 to 4 causes a ‘blunting’ (smoothing) effect on the odd-numbered vertices of the SMGC geometry.

Figure 3.6: Effect of parameters n_3 on the geometry ($m_1 = m_2 = 6$, $n_1 = n_2 = 1$).

The effect of parameter n_3 is given in Fig. 3.6. A similar effect is observed when the n_2 parameter is varied from 1 to 4, causing the ‘blunting’ effect on the even-numbered

vertices of the SMGC geometry.

Figure 3.7: Effect of parameters $n_2 = n_3$ on the geometry ($m_1 = m_2 = 6$, $n_1 = 1$).

The symmetry of the SMGC structures in more than one plane is usually achieved by setting $n_2 = n_3$. When both $n_2 = n_3$ is increased from 1 to 4, the roundedness of the vertices is increased (or the sharpness of the vertices is decreased), as shown in Fig. 3.7. Also, the scale of the geometry is increased when both n_2 and n_3 are increased. When $n_2 = n_3 = 2$, the vertices get rounded completely, and hence SMGC forms a circle geometry.

Figure 3.8: Simple SMGC modifications for a rectangular shaped patch. (a) Rectangular shape from SMGC. Variation of parameters (b) $n_1 = n_2 = n_3$, (c) $n_2 = n_3$, (d) n_1 , (e) n_2 , and (f) n_3 .

A simple SMGC modifications for a rectangular shaped patch is shown in Fig. 3.8. A simple rectangle shape is expressed by SMGC as shown in Fig. 3.8(a) with parameters $m_1 = m_2 = 4$, $n_1 = n_2 = n_3 = 100$, $S_{X1} = S_{X2} = 1.25$, $S_Y = 1$. It can be seen from

Fig. 3.8(b), that when parameters $n_1 = n_2 = n_3$ is increased from 1 to 9, the diamond shape is transformed to a rectangle with rounded edges. When parameters $n_2 = n_3$ are varied from 60 to 140, the rectangle with convex sides are changed to concave sides as shown in Fig. 3.8(c). A similar behaviour is observed if only n_1 is varied from 60 to 140, as shown in Fig. 3.8(d). If only n_2 is varied from 60 to 140, the shorter sides of the rectangle change from convex to concave as in Fig. 3.8(e), and if only n_3 is varied from 60 to 140, the longer sides of the rectangle change from convex to concave as in Fig. 3.8(f).

Figure 3.9: Simple SMGC modifications for a circle shaped patch. Variation of parameters (a) $m_1 = m_2$, (b) n_1 , (c) n_2 , (d) n_3 , (e) $n_2 = n_3$, and (f) $n_1 = n_2 = n_3$.

Fig. 3.9 shows the modifications done by using SMGC on a simple circle. Fig. 3.9(a) shows a basic circle generated using SMGC with parameters $m_1 = m_2 = 2$, $n_1 = n_2 = n_3 = 2$, $S_{x1} = S_{x2} = 1$, $S_y = 1$. Fig. 3.9(b) shows that if variable n_1 alone is changed from 2 to 10, there is no change in the geometry. Fig. 3.9(c) shows that if variable n_2 alone is changed from 2 to 10, then the circle tends to form a flatter side to its right. Similarly, from Fig. 3.9(d), if the variable n_3 alone is changed, then the circle tends to form a flatter side to its left. From Fig. 3.9(e), it can be seen that if both $n_2 = n_3$ is varied from 2 to 10, the circle extends into an ellipse vertically. Fig. 3.9(f) shows a similar behaviour i.e.,

if $n_1 = n_2 = n_3$ is varied from 2 to 10, the circle deforms in to an ellipse.

Figure 3.10: Simple SMGC modifications for a triangular shaped patch. Variation of (a) n_1 , $n_2 = n_3$, (b) n_1 and (c) $n_2 = n_3$.

Fig. 3.10(a) shows the modifications that can be made by SMGC on a simple triangle. The triangle is generated from SMGC using parameters $m_1 = m_2 = 3$, $n_1 = 50$, $n_2 = n_3 = 100$. Fig. 3.10(b) shows that vertices and sides get shortened or curved when n_1 is increased from 50 to 200. When $n_2 = n_3$ is increased from 100 to 250, the vertices get more sharpened, and the sides change from convex to concave, as shown in Fig. 3.10(c).

To highlight the advantage of SMGCs, an SMGC based microstrip-fed patch is generated with parameters $m_1 = m_2 = 32$, $n_1 = 0.4$, $n_2 = 3$, $n_3 = 3$ and compared to a similar sized conventional circular microstrip-fed patch as shown in Fig. 3.11(a) and (b).

Figure 3.11: Microstrip-fed patch antenna. (a) Circular patch. (b) Gielis patch. (c) S_{11} and realized gain of the antennas [units:mm].

The SMGC patch is fed by an exponentially tapered feeding marked as curve C given by equation

$$y = c_1 e^{Rx} + c_2 \quad (3.3)$$

$$c_1 = \frac{x_1 - x_2}{e^{(Ry_1)} - e^{(Ry_2)}} \quad (3.4)$$

$$c_2 = \frac{x_2 e^{(Ry_1)} - x_1 e^{(Ry_2)}}{e^{(Ry_1)} - e^{(Ry_2)}} \quad (3.5)$$

where (x_1, y_1) and (x_2, y_2) are starting and ending points, with $R=0.5$, $c_1=6.24$, $c_2=2.49$. Fig. 3.11(c) shows that the operating frequency of the conventional circular patch is at 3.35 GHz and that of the SMGC patch (having the same size as the conventional circular patch) is at 3.03 GHz. The maximum gain of the conventional circular patch is 7.3 dB, and that of the SMGC patch is 6.5 dB. The reduction in the gain of the SMGC patch is due to the reduced conductor area, but the difference in gain is less than 1 dB. It is to be noted that the 3 dB gain bandwidth of the conventional circular patch is 3.59%, and that of the SMGC patch is 1.65%. The parameter variation of the SMGC based monopole antenna in Fig. 3.11(b) are shown in Fig. 3.12. Fig. 3.12(a) shows variation of $m_1 = m_2$ from 16 to 32. It can be seen that when m_1 is varied from 16 to 32, the resonance frequency shifts from 3.07 GHz to 3.02 GHz. It should be noted that when m_1 is varied, the size of the antenna is not increased as only the number of spikes or vertices are increased. The gain is increased from 5.3 dB to 6.6 dB when m_1 is varied from 16 to 32. Fig. 3.12(b) shows the variation of n_1 from 0.5 to 0.8. As n_1 is increased, the frequency shifts from 3.02 GHz to 2.87 GHz as the patch becomes slightly bigger in size. The gain remains almost the same if n_1 is changed. Fig. 3.12(c) shows the variation of $n_2 = n_3$. When n_2 is varied from 0.4 to 0.7, the frequency shifts from 3.27 GHz to 2.75 GHz as the patch size gets slightly bigger with increase in $n_2 = n_3$. The gain almost remains the same, while n_2 is varied.

Figure 3.12: SMGC parameter variation for monopole patch antenna. S_{11} and gain variation with parameters (a) m_1 , (b) n_1 and (c) n_2 .

Figure 3.13: FSS unit cell. (a) Nested square. (b) SMGC nested in square. (c) S_{21} of the unit cells ($\epsilon_r = 4.3, h = 1.6$ mm) [units:mm].

Another illustration has been shown in Fig. 3.13 based on [51], to show the advantage of SMGC based structures in FSS. Fig. 3.13(a) shows nested square loops to provide a

dual-band band-stop response. Fig. 3.13(b) shows a modification in the nested square loop by replacing the inner square loop with an SMGC loop. The SMGC parameters for this inner loop is given by $m_1 = m_2 = 16$, $n_1 = 0.75$, $n_2 = 0$, $n_3 = 30$, $S_{x1}=S_{x2}=S_y=3.8$. Fig. 3.13(c) shows the S_{21} of the two FSS unit cell. It can be observed that the first stop-band frequency of the nested square and SMGC unit cells are 4.86 GHz and 4.78 GHz, respectively. The second stop-band frequency of the nested square and SMGC unit cells are 7.07 GHz and 6.63 GHz, respectively. Hence, using SMGC for the inner loop in the unit cell reduces the second stop-band frequency by 6.2% compared to the nested square unit cell. Even though the inner loop can be modified using a conventional meander, it is easier to generate/modify/tune such meander-like structures using SMGCs.

The main reason for the antenna with a Gielis-shaped edge to have a lower cut-off frequency is that the surface currents have to flow through a relatively long distance along the edge of the Gielis-shaped antenna. The perimeter of the SMGC shaped conductor is relatively longer than the conventional geometric shaped conductor. This perimeter increase makes the surface currents with slightly longer wavelength to resonate along the edge of the SMGC shaped antenna. The perimeter of SMGC shaped conductors cannot be made much longer when compared to conventionally shaped conductors. Making the perimeter longer will require the edge of the SMGC shape to have a more undulating or wavy structure, which will affect the higher-end frequency characteristics in the bandwidth. Hence, an optimal increase in the perimeter of the SMGC based antenna shape should be selected, which can be done by optimization. A better SMGC shape compared to the conventional geometry can be obtained or further improved by optimization.

The application of SMGCs in the generation of curve profiles for patches or slots with uniform scaling along both $-X$ and $+X$ directions i.e., $S_{x1} = S_{x2}$ are discussed in section 3.3 and 3.4. Then, the application of SMGCs for curve profiles with non-uniform scaling along $-X$ and $+X$ directions i.e., $S_{x1} \neq S_{x2} \neq S_y$ are discussed in section 3.5, 3.6 and 3.7.

Figure 3.14: (a) Printed UWB monopole structure with parasitic strips. (b) Normalized Gielis curve. (c) Scaled-up Gielis curve (axes in mm).

3.3 Unidirectional CPW-fed monopole antenna

In this section, a patch shape of the monopole and reflector [192] is designed using SMGC curve profiles with uniform scaling along both $-X$ and $+X$ directions i.e., $S_{X1} = S_{X2}$. The calculation of the curve profile needed for the monopole and the dimensions of the parasitic strips are discussed in subsection 3.3.1. The designed antenna is then simulated, and the corresponding prototype is fabricated. The results of the simulation and measurement such as return loss and gain are discussed in subsection 3.3.2.

3.3.1 Antenna structure

The antenna structure comprises of a printed monopole antenna based on supershape equation and has a reflector to create a unidirectional pattern, which is further enhanced by adding three parasitic microstrips as shown in Fig. 3.14(a).

A design flow for choosing the SMGC parameters for a particular design is shown in Fig. 3.15.

Figure 3.15: Design flow for selecting SMGC geometry.

The geometry of the circular monopole with reflector whose shape is to be modified is shown in Fig. 3.16. The following steps can help in giving an approach to choose SMGC parameters for this example.

Figure 3.16: Observing the geometry of circular monopole with reflector to be modified.

1. *Observing the geometry to be modified:* Observe the geometry of the patch to be modified in the antenna. The circumference of the circular patch antenna will greatly influence the lower cut-off frequency. Hence, to lower the cut-off frequency, the circumference or perimeter of the circular patch should be increased in the available space. The only way to increase the perimeter is to have a geometry

with a ‘wavy’ edge as the patch radiator. Therefore, the aim is to choose SMGC parameters that can provide such a ‘wavy’ edge for the circular monopole.

Figure 3.17: (a) Choosing number of vertices or symmetry planes. (b) Choosing parameters $n_1 < n_2 = n_3$. (c) Choosing parameters $n_1 > n_2 = n_3$. (d) Tuning $n_1, n_2 = n_3$.

2. *Determine the symmetry of the SMGC geometry:* The number of vertices in an SMGC geometry is decided by the parameters $m_1 = m_2$. In most cases, m_1 and m_2 are set as the same value for continuity of the curve. Initially, to generate a ‘wavy’ geometry $m_1 = m_2$ is set as 10, which means 10 vertices are created as shown in Fig. 3.17(a). For a given $m_1 = m_2$, there are m_1 unique symmetric planes. Then for n_1, n_2, n_3 , the initial value is set as 1.
3. *Choose the number of vertices required:* For a ‘wavy’ edge to be generated, the value of $m_1 = m_2$ should be set at least to 10. This will give a ‘ripple’ appearances for the geometry, and the perimeter can be increased. Later the value of $m_1 = m_2$ can be increased or decreased by observing the S_{11} and gain performance.
4. *Determine the smoothness of SMGC geometry:* Fig. 3.17(a) shows that the vertices are quite sharp when $n_1 = n_2 = n_3 = 1$. Hence, the parameters are varied with simple relation $n_1 < n_2 = n_3$. The values are initially chosen as $n_1 = 1, n_2 = n_3 = 3$ as shown in Fig. 3.17(b). It can be observed that the vertices have now become smooth, but still, the vertices are protruding to a considerable extent. This protrusion will affect the antenna’s performance at higher frequencies even though the perimeter can be increased appreciably. To reduce the height of these protrusions, another relation between variables is chosen i.e. $n_1 > n_2 = n_3$ with

$n_1 = 4, n_2 = n_3 = 3$ as shown in Fig. 3.17(c). The protrusions have become very smooth and appear like a ‘wavy’ edge as required. The ‘wavy’ edge can further be tuned or optimized based on the S_{11} performance of the antenna with some optimization goals as $\{S_{11}(\text{dB})|_{f_1}^{f_2} + 10 \text{ dB}\} < 0$, where f_1 and f_2 are the lower and upper cut-off frequencies of the antenna. While maintaining the relation $n_1 > n_2 = n_3$, the values can further be adjusted as $n_1 = 8, n_2 = n_3 = 5$ which is shown in Fig. 3.17(d).

Figure 3.18: Effect of scaling variables (a) S_{X1} , (b) S_{X2} , (c) S_Y , and (d) $S_{X1} = S_{X2} = S_Y$.

5. *Determine the scaling required (S_{X1}, S_{X2}, S_Y):* The effect of scaling variables is shown in Fig. 3.18. In this circular monopole case, symmetric scaling would suffice i.e. $S_{X1} = S_{X2} = S_Y$. In case if asymmetric scaling is required, then the scaling variables can be individually varied. In Fig. 3.18(a), it can be seen that the scaling can be done at $-X$ direction by varying $S_{X1}=2$. Similarly, the scaling can be done at $+X$ direction by varying $S_{X2}=2$ as shown in Fig. 3.18(b). If only the scaling need to be varied in $\pm Y$ directions, then S_Y is set as required, which is shown in Fig. 3.18(c). If the SMGC geometry need to be scaled uniformly, then all scaling variables can be uniformly increased $S_{X1} = S_{X2} = S_Y=2$ as in Fig. 3.18(d).

To justify whether the obtained SMGC shape is right, the general approach is to optimize the SMGC parameters to determine the shape of the antenna with simple optimization goal functions involving S_{11} or realized gain. Most of the reported designs are done using simple optimization and some using trial and error method.

The profile for the monopole is initially given by the conventional Gielis equation from

(3.1) where the values of parameters $a = 1$, $b = 1$, $m_1 = 20$, $m_2 = 20$, $n_1 = 15$, $n_2 = 4$, $n_3 = 4$ determines the nature of the curve. m_1 and m_2 are chosen as 20 to increase the number of vertices in the shape. When the inherent scaling variables a and b are set as 1, then the conventional Gielis curve can be called as a normalized Gielis curve obtained from (3.1) as shown in Fig. 3.14 (b). The normalized Gielis curve can then be scaled up after conversion from polar to rectangular co-ordinates. In cases where $S_{x1} = S_{x2}$, the general SMGC equation from (3.2) can be reduced to the following equation

$$\begin{aligned} X &= r(\theta) \cos(\theta) S_x \\ Y &= r(\theta) \sin(\theta) S_y \end{aligned} \quad (3.6)$$

where $S_x = 11.8$ and $S_y = 11.8$ are the scaling constants that are used to obtain the curve as shown in Fig. 3.14 (c) from (3.6). The center of the curve $Gc1$ is at (17.7 mm, 21.56 mm) from the origin xy . The antenna is designed on an FR4 substrate with dielectric constant $\epsilon_r = 4.3$ and substrate height = 1.6 mm. The dimensions of the antenna structure is given as $w = 56$ mm, $l = 44$ mm, $wf = 2$ mm, $s = 50.5$ mm, $w2 = 12.5$ mm, $w3 = 13.5$ mm, $lf = 10$ mm, $g1 = 1.5$ mm, $g2 = 1.5$ mm, $g = 16.33$ mm, $w1 = 1$ mm, $d = 8$ mm, $l1 = 16$ mm.

The patch is used as the radiator, and the edge of the reflector part facing the patch is shaped by the Gielis curve, which will influence the surface currents and the impedance matching. The placement of the reflector creates a unidirectional pattern, which is further enhanced by adding parasitic strips. The length of the parasitic strips is 0.112λ , the inter-distance between the parasitic strips is 0.056λ , and the width of strips is about 0.007λ , where λ is the wavelength at the cut-off frequency 2.1 GHz.

3.3.2 Simulation and measurement results

In order to better understand the effect of Gielis shape on the antenna, the monopole antenna without parasitic strips is analysed as shown in Fig. 3.19. The monopole antenna

with a reflector having a circular edge is shown in Fig. 3.19(a), and the antenna having a Gielis-shaped edge is shown in Fig. 3.19(b). The S_{11} for the antenna with circular and Gielis-shaped edge is given in Fig. 3.19(c). It can be seen that the lower cut-off frequency for the antenna with the Gielis-shaped edge is 2 GHz, and that of the circular edge is 2.14 GHz. Even though the reduction in lower cut-off frequency is not so high, it can be observed that the Gielis-shaped edge can help in reducing the lower cut-off frequency. Considering the Gielis-shaped monopole antenna with parasitic strips, the -10 dB impedance bandwidth of the antenna is from 2.1 GHz to around 10 GHz with a bandwidth of 130.58%. The simulated return loss and the measured return loss are shown and compared in Fig. 3.20. Some of the structures in the thesis are simulated both in CST and HFSS so that they can be fabricated with better confidence as any mistakes in simulation settings can be avoided. The plots of results from CST and HFSS are usually very close but may differ slightly. The slight difference between CST and HFSS results could be due to the different electromagnetic computation techniques, which uses different meshing techniques. However, the difference between CST and HFSS results for a given model is very small when higher mesh settings are set in the software. For S-parameters, the variation between CST and HFSS results below -15 dB can be neglected as they are very minute. The simulated and measured return loss agree quite well for frequencies up to 10 GHz and deviate slightly beyond 10 GHz, which could be due to fabrication errors. The printed monopole with reflector and parasitic strips is simulated, and its prototype is fabricated as shown in Fig. 3.21 (a).

The simulated realized gain is given in Fig. 3.21 (b), which shows the end-fire gain at $\phi = 0^\circ$, $\theta = 90^\circ$ (realized gain along the x axis) and the peak gain in the radiation patterns over the entire range of frequencies. The peak gain is around 5 dB to around 8 dB over the entire frequency range. Without the addition of strips, the realized gain would be less than 5 dB over the operating frequency range. The end-fire realized gain and the peak realized gain is quite the same for frequency ranges of 2 GHz to 5 GHz and

Figure 3.19: Printed monopole with reflector and no parasitic strips. Structure of (a) antenna with circular edge, (b) antenna with Gielis-shaped edge. (c) S_{11} of antenna with circular and Gielis-shaped edge.

Figure 3.20: Reflection coefficient of the UWB monopole structure with parasitic strips.

also from 10 GHz to 12 GHz. The realized gain patterns are shown in Fig. 3.22 (a) and (b), which illustrates the azimuthal (H-plane) and elevation (E-plane) patterns at 3 GHz and 10 GHz. The H-plane is given by the cut-plane $\phi = 0^\circ$ which is similar to xz plane and the E-plane is given by cut-plane $\theta = 90^\circ$ similar to xy plane. For H-plane and E-plane patterns, the x axis is along 270° . The parasitic strips which act as directors, can increase the directivity in addition to the presence of reflector. The main lobe magnitude along the x axis is comparatively greater than the back lobe. In the frequency ranges of 6 GHz to 9 GHz and 12 GHz to 14 GHz, the main lobe of the realized gain pattern is slightly shifted away from the x axis. The strips also help in reducing the side-lobes in the E-plane pattern at 10 GHz, as shown in Fig. 3.22(b).

Figure 3.21: (a) Fabricated prototype of the UWB monopole structure with parasitic strips. (b) Simulated endfire and maximum realized gain.

Figure 3.22: Radiation patterns at (a) 3 GHz (b) 10 GHz (— CST H-plane, -- HFSS H-plane, ... CST E-plane, - · HFSS E-plane).

In conclusion, a supershaped CPW-fed monopole antenna formed from SMGC curve profiles with uniform scaling along both the $-X$ and $+X$ directions i.e., $S_{x1} = S_{x2}$ having parasitic strips for unidirectional pattern has been proposed. A Gielis curve based patch with planar reflector and parasitic strips have been used to enhance the directivity. The characteristics of the antenna such as return loss, realized gain, and radiation pattern have been simulated, and the return loss has been compared with the measured results. The -10 dB impedance bandwidth of the designed antenna is from 2 GHz to around 10 GHz with peak gain around 4 dB or above for most of the frequency range. The next

section deals with using the SMGC profiles for the slot based antenna having $S_{x1} = S_{x2}$.

The printed monopole with reflector and parasitic strips discussed in Section 3.3 is designed for operating over a UWB band. The parasitic strips are not included in this discussion to compare the conventional and SMGC shapes easily. Fig. 3.20 in the thesis shows a comparison of antennas with circular and SMGC edge for the UWB band. If the SMGC structure has ‘deeper wavy modifications’ it may cause undesired distortion at high-frequency ranges. Hence, only SMGCs with a ‘mild wavy’ edge can be used to increase perimeter and in turn reduce the lower cut-off frequency slightly. The improvement by using the SMGC curve (instead of the circular curve) for this antenna is shown in Table 5.3 with a 6.5% reduction in lower cut-off frequency and 2% improvement in bandwidth.

To further illustrate the advantage of SMGCs in reflector monopole antenna, another illustration is shown in Fig. 3.23 for broadband operation.

Figure 3.23: CPW-fed monopole patch with reflector. (a) Circular. (b) SMGC. (c) S_{11} and gain of the antennas [units:mm].

A conventional circular monopole with reflector is shown in Fig. 3.23(a) and an SMGC based patch of same size is shown in Fig. 3.23(b). From Fig. 3.23(c), it can be seen that the lower cut-off frequency of the SMGC based patch is reduced to 1.9 GHz from 2.21 GHz, which is the lower cut-off frequency of the circular patch. The bandwidth of the SMGC based patch is 44.9% while that of the circular patch is 58%. Though there is a slight

reduction in bandwidth of the SMGC based patch, the gain of the SMGC antenna is higher than the circular patch, especially in the frequency ranges 1.9 GHz to 2.6 GHz and 2.9 GHz to 3.25 GHz.

3.4 Bidirectional CPW-fed slot antenna

In this section, the shape of the slot and the shape of the tuning stub patch [193] is also designed with SMGC curve profiles with uniform scaling along both $-X$ and $+X$ directions i.e., $S_{x1} = S_{x2}$. Here the contours from SMGCs helps in providing additional surface current paths, and the curve profile is easier to construct and optimize than using a combination of conventional geometries. The design steps for the generation of the curve profile that outlines the slot, tuning stub patch, and other dimensions of the antenna structure are discussed in subsection 3.4.1. The designed antenna is then simulated, and the antenna structure is fabricated. The results of the simulation and measurement, such as return loss and gain, are discussed in subsection 3.4.2.

3.4.1 Antenna structure

The shapes of the patch and the slot used in the antenna are based on the SMGCs, which can have a slightly different current distribution compared to a circular slot. The curve profile for the slot and patch is based on the proposed SMGC equation from (3.1), where the parameters $a, b, m_1, m_2, n_1, n_2, n_3$ are given in Table 3.1. The slot antenna geometry is given in Fig. 3.24(a) and the normalized curve is shown in Fig. 3.24(b).

The curve given by (3.1) is then scaled up after converting the polar form to rectangular form as expressed by the equation (3.6), where S_x and S_y are scaled equally as shown in Table 3.1. The curves $Gc1$ and $Gc2$ are scaled up curves as shown in Fig. 3.24(c) from (3.6). Only the scaling constants are different for curves $Gc1$ and $Gc2$ as shown in Table 3.1. The center for the curve $Gc1$ is at (40 mm, 30.6 mm) and $Gc2$ at (40 mm,

Figure 3.24: (a) CPW-fed slot antenna structure. (b) Normalized Gielis curve. (c) Scaled-up Gielis curves (— $Gc1$ curve, -- $Gc2$ curve; axes in mm).

Table 3.1: Parameters for curves $Gc1$ and $Gc2$ in CPW-fed slot

Curves	a	b	m_1	m_2	n_1	n_2	n_3	S_x	S_y
$Gc1$	1	1	10	10	20	6	6	9.5	9.5
$Gc2$	1	1	10	10	20	6	6	23	23

43 mm) from the origin. The antenna design is based on an FR4 substrate with dielectric permittivity $\epsilon_r = 4.3$ and substrate height $h = 1.6$ mm. The antenna dimensions are given as $w = 80$ mm, $wf = 3$ mm, $lf = 20$ mm, $s = 0.4$ mm, $g = 0.8$ mm. The gap g in the transition between the co-planar feed and the patch decides the impedance matching. The SMGCs are used to slightly alter the surface currents along the edges in an effort to increase the circumference of the slot and patch, which may help in a slight reduction of the cut-off frequency. Also, the complex shape can be easily generated from SMGCs rather than using a combination of many other standard shapes to form a complex shape.

3.4.2 Simulation and measurement results

The effect of SMGCs on the antenna compared to the effect of the conventional circular slot is shown in Fig. 3.25. The slot antenna with circular edge is shown in Fig. 3.25(a)

Figure 3.25: Structure of slot monopole antenna (a) with circular edge, (b) with Gielis-shaped edge. (c) S_{11} of antenna with circular and Gielis-shaped edge.

Figure 3.26: Reflection coefficient of the printed CPW-fed slot antenna using Gielis curves.

and the antenna with the SMGCs for the edge is shown in Fig. 3.25(b). The S_{11} of both the antennas are shown in Fig. 3.25(c), and it can be seen that the lower cut-off frequency for the Gielis edge is 2 GHz while that of the circular edge is 2.2 GHz. It is inferred that this reduction in the lower cut-off frequency is due to the usage of the Gielis profile for the edge. Considering the slot antenna with SMGC, the -10 dB impedance bandwidth is from 2.5 GHz to around 9 GHz, as shown in Fig. 3.26, which gives a bandwidth of 113.04%. The circular slot antenna using Gielis curves is simulated, and the corresponding prototype is fabricated, as seen in Fig. 3.27(a). The return loss from simulation agrees quite well with the measured return loss at the low frequencies but is slightly deviating at high frequencies beyond 9 GHz, which could be due to fabrication errors. The simulated realized gain are shown in Fig. 3.27(b) describing the gain along

Figure 3.27: Fabricated prototype of the CPW-fed slot antenna and comparison of realized gain.

Figure 3.28: Radiation patterns at (a) 3 GHz and (b) 6 GHz (— CST H-plane, – HFSS H-plane, $\cdots$ CST E-plane, $-\cdot$ HFSS E-plane).

the bore-sight direction and peak gain. The bore-sight gain is along $\theta = 0^\circ$ and $\phi = 0^\circ$. The bore-sight gain and the peak gain are also the same for the frequency range of 2 GHz to 4 GHz. The maximum gain point is slightly shifted from the bore-sight direction, which is a characteristic of the CPW monopole antenna. The peak is around 5 dB or above over the entire frequency range. The simulated realized gain patterns along the elevation plane (E-plane), and azimuthal plane (H-plane) are shown in Fig. 3.28(a) and (b) at two frequencies 3 GHz and 6 GHz. The E-plane is along the yz plane, and the H-plane is

along the xz plane, which can also be alternatively described as a cut plane $\phi = 0^\circ$ and $\phi = 90^\circ$ respectively. The E- and H-plane pattern at 3 GHz shows that the maximum gain lobe is along the bore-sight direction z . The front lobe and back lobe magnitudes are almost the same. The patterns at 6 GHz show that the main lobe is slightly shifted away from the bore-sight direction, which is common in CPW-fed monopole antennas.

It is concluded that broadband printed CPW-fed slot antenna using SMGCs with uniform scaling along $-X$ and $+X$ directions i.e., $S_x = S_{x1} = S_{x2}$ has been discussed. The super-shaped slot and the super-shaped patch have been used to provide additional surface current paths along the edges, which also influences impedance matching. The characteristics of the antenna, such as return loss, realized gain and radiation pattern have been simulated, and return loss has been compared with the measured results. The -10 dB impedance bandwidth of the designed antenna is from 2.5 GHz to around 9 GHz with the gain around 5 dB or above for most of the bandwidth.

The comparison between conventional and SMGC based coplanar slot antenna is given in Table 5.3 in the row ‘Bidirectional CPW-fed slot antenna’. The SMGC based slot antenna has reduced the lower cut-off frequency by 9.09% and has increased the -10 dB bandwidth by 5.27% compared to the conventional circular slot antenna.

Extending the SMGC curve profiles to form further complex shapes with different scaling along $-X$, $+X$, and $\pm Y$ directions i.e., $S_{x1} \neq S_{x2} \neq S_y$ will be discussed in next section.

3.5 Bidirectional microstrip-fed monopole antenna

In this section, the design of planar mushroom-shaped wideband monopole antenna [194] using SMGC curve profiles with different scaling along $-X$, $+X$, and $\pm Y$ directions i.e., $S_{x1} \neq S_{x2} \neq S_y$ is discussed. The ability to scale differently along $-X$, $+X$ and $\pm Y$ directions gives additional degrees of freedom to SMGCs than conventional Gielis curves.

Figure 3.29: (a) Planar mushroom-shaped monopole antenna. (b) Normalized Gielis curves. (c) Scaled-up Gielis curves (— $Gc1$ curve, — $Gc2$ curve [axes in mm]).

In subsection 3.5.1, two Gielis curves are used to design the complex shape profiles of the monopole conveniently. The expressions of the curve profiles and various dimensions of the monopole are discussed. In subsection 3.5.2, the performance characteristics of the antenna, such as return loss, realized gain, and radiation pattern are simulated, and the return loss is compared with the measured result.

3.5.1 Antenna structure

The planar mushroom-shaped wideband monopole antenna is shown in Fig. 3.29(a). The antenna is fed by a microstrip with a notched ground plane at the bottom layer. Two Gielis curves are used to design the complex shape profiles of the monopole. Using (3.1), with $(a = b = 1)$ for $Gc1$ and $Gc2$ indicates normalized Gielis curves as shown in Fig. 3.29(b), where θ is chosen to be swept from 90° to 450° . The parameters m_1 and m_2 determine the number of vertices to be produced in the Gielis curves depending on the ‘rectangle’ or ‘bell’ shape to be produced. The parameters a and b can be used for scaling, but their values are often high to form a required geometry and may cause distortion in

Table 3.2: Parameters for curves $Gc1$ and $Gc2$ in mushroom monopole

Curves	a	b	m_1	m_2	n_1	n_2	n_3	S_{x1}	S_{x2}	S_y
$Gc1$	1	1	5	5	4	4	4	2	15	10
$Gc2$	1	1	4	4	18	15	15	7.5	7.5	4

Figure 3.30: Structure of (a) monopole with rectangular and circular patch (MRCP), (b) mushroom antenna with Gielis shape and (c) S_{11} of the MRCP and Gielis-shaped mushroom antenna.

profile. Alternatively, the SMGCs are used more conveniently based on the sectors using (3.2) with the parameter values from Table 3.2 as seen in Fig. 3.29(c). The curve $Gc2$ is then rotated 90° counter-clockwise about its shape center. The shape centers (X, Y) of the curves $Gc1$ and $Gc2$ are then moved from the origin (x, y) to $C1$ (17.5 mm, 16 mm) and $C2$ (17.5 mm, 26.13 mm) respectively. FR-4 substrate with dielectric constant $\epsilon_r = 4.3$ and thickness $h = 1.6$ mm is used for the design. The parameters of the overall antenna design are: $w = 35$ mm, $l = 42.2$ mm, $gl = 11$ mm, $wf = 3$ mm, $sl = 1$ mm, $sw = 5$ mm, $g = 1$ mm, $pw = 23$ mm and $pl = 23$ mm.

3.5.2 Simulation and measurement results

A monopole with the rectangular and circular patch (MRCP), i.e., an antenna having a combined shape of a rectangle and circle for the patch is shown in Fig. 3.30(a). A mushroom-shaped antenna using Gielis curves is shown in Fig. 3.30(b). The S_{11} of both the antennas are shown in Fig. 3.30(c), and it can be inferred that the lower cut-off frequency of Gielis based mushroom is 2.5 GHz while that of MRCP is 2.5 GHz. Also, it

Figure 3.31: (a) Fabricated prototype. (b) S_{11} of the Gielis-shaped wideband mushroom monopole antenna.

can be seen that the impedance matching is better for Gielis shaped mushroom than the MRCP antenna in the frequency range of 2.5 GHz and 4 GHz. Hence, the Gielis shape profile can help in improving the impedance matching and reducing the lower cut-off frequency.

The mushroom-shaped monopole antenna is simulated using full-wave electromagnetic software CST and HFSS. The prototype is fabricated using milling technology, as shown in Fig. 3.31(a). The simulated return loss of the antenna is also compared with the measured result, as shown in Fig. 3.31(b). The -10 dB impedance bandwidth of the antenna is from 2.5 GHz to 9 GHz, which gives a wide bandwidth of 113%. The simulated and measured return loss results agree quite well except that there is a slight shift in the lower cut-off frequency, which could be due to the dielectric substrate. The simulated realized gain curves can be seen in Fig. 3.32, which shows the boresight gain and the peak gain versus frequencies. The boresight gain is along the direction of z axis where $\phi = 0^\circ$, $\theta = 0^\circ$ and the peak gain is the maximum gain of the 3-D radiation pattern. Notice that the peak gain is about 3 dB to 4.5 dB. The realized gain along the boresight direction and the peak gain are quite close from 2.5 GHz to 4 GHz. Above 4 GHz, the main lobe is slightly shifted away from the boresight direction, which is a common characteristic of

Figure 3.32: Realized gain of the mushroom-shaped wideband monopole antenna

planar monopole antennas.

The radiation patterns on the azimuthal plane (H-plane) and elevation plane (E-plane) are given in Fig. 3.33(a), (b), and (c) at the frequencies 4 GHz, 6 GHz, and 8 GHz, respectively. The H-plane is along the cut-plane $\phi = 0^\circ$ (xz plane) and the E-plane is along the cut-plane $\phi = 90^\circ$ (yz plane). The patterns show that the main lobes are initially along the $+z$ and $-z$ axes, but at higher frequencies, the beam is slightly shifted away from the boresight direction. The antenna patterns at 8 GHz on $\phi = 0^\circ$ and $\phi = 90^\circ$ planes are shown in Fig. 3.33(c). It can be seen that the antenna radiates at 8 GHz, though the main lobe has shifted to 25° from boresight due to distortion from the ground plane at high frequency.

The author would like to highlight the significant enhancement in this design using SMGC.

1. Better impedance matching at 3 to 4 GHz (at least 5dB reduction in return loss)
2. Gain improved by 0.5 to 1 dB over the entire operating bandwidth.

Figure 3.33: Radiation patterns at the frequencies (a) 4 GHz, (b) 6 GHz, and (c) 8 GHz. (— CST H-plane, - - HFSS H-plane, ... CST E-plane, - · HFSS E-plane)

3. The performance of the antenna is enhanced by the SMGC structure even if it has a lesser area and perimeter compared to the conventional structure.

Figure 3.34: Comparison between conventional and SMGC shaped microstrip-fed monopole (mushroom) antenna. (a) S_{11} . (b) Realized gain (max.).

Fig. 3.34(a) shows that the mushroom antenna (SMGC) has a 5dB better impedance matching at 3 to 4 GHz when compared to conventional shape. Fig. 3.34(b) shows that there is an improvement in the gain throughout the bandwidth. The gain of SMGC based antenna has increased by 0.5 to 1 dB compared to the conventional structure.

In conclusion, the design of the planar mushroom-shaped wideband monopole antenna using Gielis curves has been presented. The Gielis curves have been used to design the complex shape profiles of the monopole conveniently. The performance characteristics of the antenna, such as return loss, realized gain, and radiation pattern have been simulated, and the return loss has been compared with the measured result. The -10 dB impedance bandwidth of the antenna is from 2.5 GHz to 9 GHz, and the peak gain is about 3 dB to 4.5 dB. Similarly, a design of another microstrip-fed antenna such as antipodal bowtie antenna using SMGC curve profiles with different scaling along $-X$, $+X$ and $\pm Y$ directions i.e., $S_{x1} \neq S_{x2} \neq S_y$ will be discussed in next section.

3.6 Bidirectional antipodal bowtie antenna

In this section, a supershape formula involving SMGCs with $S_{x1} \neq S_{x2} \neq S_y$ is utilized to generate the shape of bowtie patch [195], which is compared with a conventional bowtie patch (CBP) with triangular shape. The design procedure for forming the curve profile that outlines the bowtie patch and other dimensions of the antenna structure are discussed in subsection 3.6.1. The designed antenna is simulated, fabricated, and measured. The results of simulation and measurement including, S parameter and gain are discussed and compared in subsection 3.6.2.

3.6.1 Antenna structure

The double-sided printed bowtie antenna is fed by a microstrip to stripline transition. The width of the stripline is tapered to increase the characteristic impedance gradually

Figure 3.35: (a) Conventional bowtie patch (CBP). (b) Bowtie patch using G_1 profile. (c) Normalized Gielis curve. (d) Scaled-up Gielis curve [units: mm].

toward the center of the bowtie. The CBP comprising triangular patch of dimensions $p_w = 13.6$ mm and $p_h = 18$ mm is shown in Fig. 3.35(a). To obtain better impedance matching at lower frequencies, the profile of the bowtie patch is designed using the supershape formula involving a modified form of Gielis curve G_1 as shown in Fig. 3.35(b). Both CBP and bowtie patch using G_1 profile have the same feeding part consisting of microstrip to stripline transition and same ground plane size. Using SMGCs from (3.1) with parameters $a = 1$, $b = 1$, $m_1 = 3$, $m_2 = 3$, $n_1 = 5$, $n_2 = 10$, $n_3 = 10$ are used to form the normalized Gielis curve of G_1 is shown in Fig. 3.35(c). This curve is then scaled using (3.2) where

the values $S_{x1} = 1.7$, $S_{x2} = 11$, $S_y = 5$ are assigned to get the curve profile shown in Fig. 3.35(d). The center point C_1 of the scaled-up curve is at (8.4 mm, 25.5 mm) from the origin. The inset in Fig. 3.35(b) gives the details about the connection of the strip line feed to the bowtie patch, which is sensitive for impedance matching of the antenna.

The antipodal bowtie antenna is designed on an FR4 substrate with dielectric constant $\epsilon_r = 4.3$ and substrate height $h = 1.6$ mm. After some parametric studies, the final antenna design is given by $w = 38.8$ mm, $l = 38$ mm, $w_f = 2.76$ mm, $g_l = 4.65$ mm, $w_1 = 2.3$ mm, $l_1 = 7.3$ mm, $w_2 = 2$ mm, $l_2 = 9.8$ mm, $w_3 = 1.12$ mm, $l_3 = 3.43$ mm, $w_c = 1.5$ mm. The microstrip feed of width w_f has a characteristic impedance of around 50Ω is connected to the tapered stripline sections of width w_1 , w_2 , w_3 whose characteristic impedances are 24.19Ω , 26.87Ω , 39.86Ω respectively. The length l_i ($i = 1, 2, 3$) of each stripline section affects the bowtie antenna bandwidth and impedance matching. The length g_l and width w of the ground plane affects the radiation pattern of the bowtie antenna. The triangular patch of the CBP and the bowtie patch using G_1 profile have an almost similar area, as observed from the dashed lines shown in Fig. 3.35(b). The effects of changing the bowtie patch profile on the antenna characteristics as compared to CBP can be seen from its current distribution along the patch edge. It can be noted from Fig. 3.36 that the surface current at 4 GHz along the edge of G_1 antenna is quite higher than that along the edge of the CBP antenna as marked by the red ellipses. The physical explanation for the bandwidth improvement or gain enhancement of the SMGC based antennas is due to the variation in the surface current path along the edge of the conductor patch modified by SMGC.

3.6.2 Simulation and measurement results

The bowtie patch using G_1 profile is simulated using CST and HFSS. The designed prototype is fabricated whose top and bottom layers are shown in Fig. 3.37. The simulated and measured S_{11} results for the G_1 bowtie patch are also shown in the figure. It can

Figure 3.36: Current distribution at 4 GHz: (a) CBP antenna and (b) G_1 antenna.

Figure 3.37: S parameter of double-sided printed bowtie using G_1 profile and CBP

be observed that the results agree quite well with each other. Two noticeable resonance peaks are observed at around 3.47 GHz and 5.65 GHz. The cut-off frequency is determined mostly by the size of the bowtie patch. From Fig. 3.37, it is also noted that the G_1 bowtie patch has better impedance matching at lower frequencies compared to CBP.

The peak values of realized gain for both the G_1 bowtie patch and CBP are shown in Fig. 3.38. The simulated and the measured peak gain values for the G_1 bowtie patch agree quite well. It is also observed that the peak gain for the CBP is slightly lower than that of the G_1 bowtie patch at lower frequencies due to the poorer impedance matching of CBP. The peak gain values vary within a range of about 3 dB to 5 dB. The realized gain at the boresight direction for both the G_1 bowtie patch and CBP at $\theta = 0^\circ$, $\phi = 0^\circ$ (along z axis) is shown in Fig. 3.39. The simulated and measured boresight gain results

Table 3.3: Comparisons of Bowtie Slot Antennas

Ref.	ε_r	f_{low}	f_{high}	FBW (%) ^a	L (λ_g) ^b	W (λ_g) ^b	Gain ^c
[66]	2.6	0.76	2.5	106.7	0.51	0.51	2 to 6 dB
[67]	4.4	3.1	10.6	109.4	0.5	0.5	0 to 3.5 dB
[68]	2.2	0.7	2.5	112.5	0.23	0.38	4.9 dB
[69]	2.2	3	10.6	111.7	0.4	0.26	NA ^d
[175]	6.15	3.1	10.6	109.5	0.7	0.7	2 to 3.4 dB
This work	4.3	3.1	11	112	6.4	6.5	2 to 4 dB

^a -10 dB Bandwidth; ^b $\lambda_g \approx c/(f_{\text{low}}\sqrt{(\varepsilon_r + 1)/2})$; ^c Realized Gain; ^d Not Available; f (GHz)

Figure 3.38: Peak gain of double-sided printed bowtie using G_1 profile and CBP.

are almost the same for the G_1 bowtie patch. Notice that the G_1 bowtie patch provides better boresight gain than the CBP. The boresight gain varies from 2 dB to 4 dB over the entire operating bandwidth. The boresight realized gain and the peak realized gain are quite close for the G_1 bowtie patch between 6 GHz and 8 GHz.

The normalized realized gain patterns of the bowtie patch using G_1 profile are shown at frequencies 3 GHz, 5 GHz, 7 GHz, and 9 GHz as in Fig. 3.40. The H-plane is the xz plane which is the cut-plane at $\phi = 0^\circ$, while the E-plane is the yz plane which is the cut-plane at $\phi = 90^\circ$. The peak gain is slightly off-centered from the boresight direction due to the presence of the ground plane and the direction of feed. At few frequencies, the maximum realized gain is directed away from the boresight as the surface currents are observed to support radiation in $\theta = 90^\circ$, $\phi = 90^\circ$. Still, the boresight realized gain

Figure 3.39: Boresight gain of double-sided printed bowtie using G_1 profile and CBP.

is at least above 2 dB for most of the frequencies, as can be seen in Fig. 3.39. The simulated and normalized realized gain patterns are almost close to each other. It can be observed that the patterns support radiation in the front lobe ($\theta = 0^\circ$, $\phi = 0^\circ$) and the back lobe ($\theta = 180^\circ$, $\phi = 0^\circ$) for most of the frequencies. From above, it is noted that using G_1 profile for the bowtie patch can improve impedance matching at lower frequencies when compared to CBP. Such improvement in impedance matching in turn increases the realized gain of the G_1 bowtie patch over the CBP. Therefore, bowtie patch using SMGC profiles can improve antenna characteristics over CBP while maintaining the antenna size.

In conclusion, a UWB double-sided printed bowtie antenna using SMGCs has been proposed. The bowtie patch fed by the microstrip to stripline transition has utilized a supershape formula involving a modified form of Gielis curve to generate complex shapes readily, which improve surface current and impedance matching. The performance characteristics of the antenna including S parameter, realized gain, and radiation pattern have been simulated and compared with measured results. The -10 dB impedance bandwidth of the designed antenna is from 3 GHz to around 12 GHz with a gain of around 4 dB over most of the bandwidth. It is inferred that an antipodal bowtie patch with SMGC can have boresight gain or even overall maximum gain higher than that of

Figure 3.40: Normalized realized gain patterns at frequencies (a) 3 GHz, (b) 5 GHz, (c) 7 GHz, (d) 9 GHz.

Figure 3.41: Structure of (a) TBSA (b) BSA with G_{52} slot profile ($W_g = 54$ mm, $L_g = 91$ mm, $a = 32$ mm, $b = 33.6$ mm, $l_f = 28$ mm, $w = 3$ mm, $g = 0.4$ mm, $w_1 = 0.2$ mm $g_1 = 1$ mm, $s = 1.26$ mm)

CBP. A further improved CPW-fed bowtie slot antenna will be discussed in the next section.

3.7 Design of bowtie slots using SMGCs

In this section, sectorially modified Gielis curves (SMGCs) will be used to form the slot profile in the bowtie slot antenna (BSA) [136]. To improve the impedance matching while preserving the antenna size, the SMGCs are used for the slot profile as they can overcome the shortcomings of the conventional Gielis curve, and they form complex shapes readily. The parametric studies of slot profiles, simulated S parameters, realized gains and current distributions for various curve parameters will be discussed in subsection 3.7.1, 3.7.2, 3.7.3. Simulated and measured results of two BSA prototypes using SMGCs will be discussed and compared with previous works in subsection 3.7.4.

3.7.1 Design of BSA using SMGCs

Fig. 3.41(a) shows the structure of conventional tapered bowtie slot antenna (TBSA) having a low profile with bi-directional radiation properties. The ground plane dimensions

Figure 3.42: Representative profiles of SMGCs using parameters from Table 3.4: (a) $m_1 = 1$, (b) $m_1 = 2$, (c) $m_1 = 3$, (d) $m_1 = 5$.

are fixed to be L_g and W_g while trying to make the antenna size as small as possible. The bowtie slot of dimensions a and b is placed at the distance l_f from the feed edge. The design is built on inexpensive and readily available FR4 substrate with dielectric constant $\epsilon_r = 4.3$ and substrate height $h = 1.6$ mm. A CPW feed of width w and gap g is used to provide a 50Ω input impedance. As the dimensions and shape of the slot profile may influence the impedance matching and bandwidth, BSA with the SMGC slot profile is designed as shown in Fig. 3.41(b). The conventional triangular slot profile is also depicted as dashed lines to show the modification. Besides, the CPW feed is tapered from width w and gap g at the input port to width and gap of w_1 and g_1 respectively at the center of the bowtie slot. This tapering provides an impedance transformation of about 50Ω to 138.9Ω . The opening of the slot at the center has a gap $s = 1.26$ mm to offer a better transition from CPW to slot.

Table 3.4: Parameters for SMGCs G_{mi}

G_{mi}	m_1	n_1	n_2	S_{x1}	S_{x2}	S_y	c_x	c_y
G_{11}	1	0.7	1	12	31.2	24	-33	28
G_{12}		0.8	1.1	12	31	22	-33	28
G_{13}		0.9	0.575	12	33.5	25	-33	28
G_{14}		1	0.53	12	33.5	25	-33	28
G_R	1	1	1.8	9	16.3	11.4	-19	22
G_{21}	2	0.7	1	6	32.5	24	-33	28
G_{22}		0.8	1	6	32.5	24	-33	28
G_{23}		0.9	1	6	32.5	24	-33	28
G_{24}		1	1	6	32.5	24	-33	28
G_{31}	3	0.7	1	22	22.8	19	-23	28
G_{32}		0.8	1	22	22.8	19	-23	28
G_{33}		0.9	1	22	22.8	19	-23	28
G_{34}		1	1	22	22.8	19	-23	28
G_{51}	5	0.5	1.6	12	27.2	17	-28	28
G_{52}		0.6	1.6	12	27.2	17	-28	28
G_{53}		0.7	1.6	12	27.2	17	-28	28
G_{54}		0.8	1.6	12	27.2	17	-28	28

3.7.2 SMGCs for slot profile

The slot profiles may need to be altered for obtaining a better impedance matching, which may result in an irregular complex shape. In order to realize a complex shape for the slot profile using SMGCs, different curve profiles with different values for parameters $m_1, n_1, n_2, n_3, S_{x1}, S_{x2}, S_y$ from (3.1) and (3.2) with reference to the XY axes is shown in Fig. 3.42. The parameters c_x and c_y give the center position in mm for the SMGCs with reference to the origin of xy axes located at the bottom right corner (along mirror plane).

Table 3.4 shows the parameters of various SMGCs G_{mi} , where subscript m denotes the class of curves that decides the number of vertices and subscript i denotes the iteration for a fixed class of curves. For illustration, Fig. 3.42 plots the representative profiles of

Figure 3.43: S_{11} characteristics of BSA using SMGCs G_{12} , G_{21} , G_{33} , G_{52} and TBSA

Figure 3.44: Realized Gain ($\theta=0$, $\phi=0$) and radiation efficiency characteristics of BSA using SMGCs G_{12} , G_{21} , G_{33} , G_{52} and TBSA.

SMGCs for each class m_1 using parameters from Table 3.4. The angle positions for theta from 90° to 450° are shown in Fig. 3.42(a). The dotted lines represent the triangular slot profile of the TBSA from Fig. 3.41(a). From Fig. 3.42, it can be observed that the slot profiles vary considerably for different classes and iterations of G_{mi} .

3.7.3 Effects of SMGCs on impedance and gain characteristics

Parametric studies are carried out using the parameter values indicated in Table 3.4. Fig. 3.43 shows the simulated results of the BSA using SMGCs with better S parameter (S_{11}) for each class m_1 . It can be inferred that the impedance matching is better for

Figure 3.45: E-plane (YZ) pattern at 1.75 GHz of BSA using SMGCs G_{12} , G_{21} , G_{33} , G_{52} and TBSA.

Figure 3.46: H-plane (XZ) pattern at 1.75 GHz of BSA using SMGCs G_{12} , G_{21} , G_{33} , G_{52} and TBSA.

the BSAs with SMGCs having lower cut-off frequencies than that of TBSA. It can be observed that the -10 dB impedance bandwidth of the TBSA is from 2.77 GHz to 4.13 GHz, whereas the impedance bandwidths of the BSA with SMGCs are from around 1.7 GHz to 4.5 GHz. Therefore, the latter provides a significant bandwidth improvement over the former. As shown in Fig. 3.44, the boresight realized gain ($\phi = 0^\circ, \theta = 0^\circ$) and the radiation efficiency are almost the same for all the SMGCs since the antenna size is unchanged, but their gain is higher than that of TBSA, especially at the lower frequencies.

The E-plane ($\phi = 90^\circ$) and H-plane ($\phi = 0^\circ$) patterns for the SMGCs and TBSA can

Figure 3.47: Surface current distribution at 2 GHz for (a) TBSA (b) BSA with G_{52} slot profile.

Table 3.5: Comparisons of Bowtie Slot Antennas

Ref.	ϵ_r	f_{low}	f_{high}	FBW (%) ^a	L (λ_g) ^b	W (λ_g) ^b	Gain ^c
[71]	4.7	2.1	3	35.29	0.89*	0.54*	NA ^d
[72]	3.2	1.6	3	60.87	0.79*	0.36*	NA ^d
[74]	4.4	0.4	1.5	115.78 ^e	0.78	0.34	NA ^d (max 6)
[77]	2.33	6	16	90.9	1.19	0.7	5 to 8
[78]	2.2	3.1	10.6	109.5	1.11	0.92	4 to 6
[79]	9.8	2.1	2.9	32	0.3	0.25	NA ^d
[80]	2.17	6	12	66.67	1.07	0.53	4 to 7.5
[81]	2.2	4.75	8	50.98	1.08	0.62	6 to 7.7
G_R	2.2	2.7	5.7	71.4	0.59	0.34	4 to 7
G_{52}	4.3	1.7	4.5	90.3	0.59	0.35	4 to 6

^a -10 dB Bandwidth; ^b $\lambda_g \approx c/(f_{\text{low}}\sqrt{(\epsilon_r + 1)/2})$; ^c Realized Gain (dB); ^d Not Available; ^e bandwidth noted at -8 dB; * Approx. values; f (GHz).

be observed from Fig. 3.45 and Fig. 3.46 respectively. Considering both planes, using SMGCs in BSA not just improves the boresight gain, but also the radiation pattern. It should be noted that the G_{52} BSA has better impedance matching at the lower and higher frequencies among the SMGCs; thereby, G_{52} will be chosen for further comparison. For frequencies above 4.5 GHz, its realized gain at boresight becomes less, and the mainlobe starts to shift away from the boresight direction. Hence the upper cut-off frequency is fixed as 4.5 GHz.

The improvement in the bandwidth is due to relatively longer curve length and the

Figure 3.48: Simulated and measured S parameter (S_{11}) and realized gain of G_{52} BSA and G_R BSA

low surface current reflections provided by the G_{52} slot profile and tapered CPW feeding, as shown in Fig. 3.47. It can be observed from the encircled regions in Fig. 3.47(a) and (b) that the amplitude of the current is increased along the curve of G_{52} as compared to that of the TBSA at 2 GHz, indicating a better impedance match. The prototype of G_{52} BSA will be fabricated and compared further with TBSA. Another slot profile G_R of Table 3.5 is also built on RO5880 and compared with previous work [81] as to be discussed in the next subsection.

3.7.4 Comparisons and measured results

Two prototypes of BSAs with G_{52} and G_R slot profiles are fabricated. The first prototype G_{52} BSA is designed on the FR4 substrate, same as TBSA. The simulated and measured S_{11} for the G_{52} BSA agree quite well as shown in Fig. 3.48. Table 3.5 shows that the G_{52} BSA is better than many other bowtie antennas either in terms of fractional bandwidth (FBW) or electrical size. The conventional BSA [71] fed by 50 Ω CPW has a lower FBW of 35.29% and a bigger size when compared to G_{52} BSA. Even some BSAs with tuning metal stubs [72], [74] have relatively lower FBWs when referred to -10 dB level,

Figure 3.49: Simulated and measured radiation patterns of normalized realized gain for G_{52} BSA co- and cross-polarizations at 2 GHz and 4 GHz on (a) E (YZ) plane (b) H (XZ) plane.

e.g. the FBW of [74] should be 75.86% only (not 115.78%) from 0.9 GHz to 2 GHz. Adding symmetric vias in the slot [77] may widen the bandwidth, but it has a larger electrical size than G_{52} BSA. Adding meander shaped slot to the conventional BSA [78] may lower the cut-off frequency, but it requires a larger ground plane (antenna size) than G_{52} BSA to obtain a stable radiation pattern. The flower-shaped arrangement of CPW-fed bowtie arms on a substrate with a higher ε [79], also has a lower FBW. The leaf-shaped slot on a lower ε substrate coupled-fed by microstrip [80] has lower FBW and larger electrical size as compared to G_{52} BSA. While altering the slot edge using sinusoidal curves along its longer sides [81] may improve the FBW at a higher frequency, it is still lower than that of G_{52} BSA. The second prototype G_R BSA (within Table 3.5 dashed box) is designed with the same substrate permittivity, size, and thickness as [81]

to have a further demonstration of using the SMGC slot profile for improving FBW and reducing antenna size. The G_R BSA has the dimensions as $w = 5$ mm, $g = 0.6$ mm, $w_1 = 0.3$ mm and $g_1 = 1$ mm. Also, the simulated and measured S_{11} for the G_R BSA agree quite well, as shown in Fig. 3.48. It provides an FBW of 71.4% from 2.7 GHz to 5.7 GHz, which is higher than [81]. Thus, using SMGCs for slot profile with tapered CPW feeding can indeed provide wider bandwidth and smaller electrical size.

Fig. 3.48 shows that the measured realized gain of G_{52} and G_R along the boresight direction varies from around 3 dB to 6 dB over the frequency range of operation. It is observed that the simulated and measured realized gains are almost similar, and the deviation is small. In Fig. 3.49 (a) and (b), good agreement can be seen between the simulated and measured patterns of normalized realized gain for G_{52} BSA co- and cross-polarization at 2 GHz and 4 GHz on E plane and H plane. The cross-polarization on the H plane is below -20 dB while that on E plane is negligible. For better pattern measurement, the reflections from coaxial cables, equipment, and antenna holder structures are minimized with the aid of absorber cover.

The conventional bowtie slot and SMGC bowtie slot are shown in Fig. 3.51. Both the antennas in comparison, have the same tapered feeding. The comparison of S_{11} and half-power beamwidth (HPBW) of both types of bowtie slot antennas are shown in Fig. 3.50.

Figure 3.50: Comparison between conventional and SMGC shaped bowtie slot antenna. (a) S_{11} . (b) Half-power beamwidth ($^{\circ}$).

Fig. 3.50(a) shows that the SMGC bowtie slot has a cut-off frequency reduced by 7.5% and bandwidth increased by 26.44%, observed at -15 dB. Fig. 3.50(b) shows that the HPBW of SMGC based bowtie slot has increased in the operating bandwidth. At least 5° to 10° improvement in HPBW is observed for SMGC based slot antenna which is indicated in Table 5.3.

In conclusion, a wideband bowtie slot antenna designed using SMGCs has been presented. The SMGCs have been used for the slot profile to improve the bandwidth while maintaining the antenna size. Two prototypes of BSAs have been built on FR4 and RO5880 substrates providing FBWs of about 90.3% (1.7 GHz to 4.5 GHz) and 71.4% (2.7 GHz to 5.7 GHz), respectively. The simulated S parameters, realized gains, current distributions, and radiation patterns along with measured results have been discussed.

3.8 Conclusion

A short summary of microstrip-fed monopole (mushroom) antenna in section 3.5, microstrip-fed antipodal bowtie antenna in section 3.6, and CPW-fed bowtie slot antenna in section 3.7 are given here, as the size of SMGC may appear to be larger than conventional ge-

ometry. The dimensions between conventional structure and SMGC structure have been compared since the proposed structures may appear to have larger dimensions.

The perimeter of a closed SMGC curve is given by

$$r(\theta) = \left(\left| \frac{\cos(\frac{m_1\theta}{4})}{a} \right|^{n_2} + \left| \frac{\sin(\frac{m_2\theta}{4})}{b} \right|^{n_3} \right)^{\frac{-1}{n_1}} \quad (3.7)$$

$$X = r(\theta) \cos(\theta) S_x \quad (3.8)$$

$$Y = r(\theta) \sin(\theta) S_y \quad (3.9)$$

$$\begin{aligned} \frac{dr(\theta)}{d\theta} &= \frac{-1}{n_1} \left(\left| \frac{\cos(\frac{m_1\theta}{4})}{a} \right|^{n_2} + \left| \frac{\sin(\frac{m_2\theta}{4})}{b} \right|^{n_3} \right)^{\frac{-1-n_1}{n_1}} \\ &\quad \cdot \left(\frac{-m_1 n_2}{4} \left| \frac{\cos(\frac{m_1\theta}{4})}{a} \right|^{n_2} \tan\left(\frac{m_1\theta}{4}\right) \right. \\ &\quad \left. + \frac{m_2 n_3}{4} \left| \frac{\sin(\frac{m_2\theta}{4})}{b} \right|^{n_3} \cot\left(\frac{m_2\theta}{4}\right) \right) \end{aligned} \quad (3.10)$$

$$\frac{dX}{d\theta} = \left(\frac{dr(\theta)}{d\theta} \cos(\theta) - r(\theta) \sin(\theta) \right) S_x \quad (3.11)$$

$$\frac{dY}{d\theta} = \left(\frac{dr(\theta)}{d\theta} \sin(\theta) + r(\theta) \cos(\theta) \right) S_y \quad (3.12)$$

$$Perimeter = \int_{\theta_1}^{\theta_2} \sqrt{\left(\frac{dX}{d\theta} \right)^2 + \left(\frac{dY}{d\theta} \right)^2} d\theta \quad (3.13)$$

$$Area = \frac{1}{2} \int_{\theta_1}^{\theta_2} \left(X \frac{dY}{d\theta} - Y \frac{dX}{d\theta} \right) d\theta \quad (3.14)$$

Figure 3.51: Overlapped geometry of conventional and SMGC structure for (a) microstrip-fed monopole (mushroom) antenna, (b) microstrip-fed antipodal bowtie antenna, and (c) CPW-fed bowtie slot antenna.

The perimeter and area of the enclosed of the SMGC curves is given by (3.13) and (3.14).

Table 3.6: Geometry comparison of conventional and SMGC shapes

Curve	Perimeter (mm)	Area (mm ²)	% $f_l \downarrow$	Gain* (dB) $\uparrow$	Improvement by using SMGC
C_{r1}	83.15	425.66	6.4	1	Better impedance bandwidth Increased gain
G_{r1}	81.56	397.4			
C_{r2}	49.8	132.6	-	1	Better impedance matching Increased Gain
G_{r2}	49.42	175.57			
C_{r3}	106	543.3	7.5	-	-15dB impedance bandwidth increased by 32% HPBW increased by 10°
G_{r3}	107.8	811			

$f_l \downarrow$: % of lower cut-off frequency reduced; * Maximum gain improvement by SMGC structure.

From Table 3.6, it can be seen that the perimeter of the conventional geometry and the SMGC geometry are quite the same. Only the areas of the two geometries are appreciably different, though the footprints of the geometries are almost similar. The frequency response of the antenna is mainly dependent on the perimeter of the conductor geometry. In these cases, the perimeter is almost the same, but the antenna performance is improved in terms of impedance matching, bandwidth, and gain.

Figure 3.52: Comparison between conventional and SMGC shaped microstrip-fed monopole (mushroom) antenna. (a) S_{11} . (b) Realized gain (max.).

Fig. 3.52(a) shows that the mushroom antenna (SMGC) in Fig. 3.51(a) has a 5dB better impedance matching at 3 to 4 GHz when compared to conventional shape. Fig. 3.52(b) shows that there is an improvement in the gain throughout the bandwidth. The gain of SMGC based antenna has increased by 0.5 to 1 dB compared to the conventional structure, as shown in Table 3.6.

Figure 3.53: Comparison between conventional and SMGC shaped microstrip-fed antipodal bowtie antenna. (a) S_{11} . (b) Realized gain ($\theta = 0, \phi = 0$).

Fig. 3.53(a) shows that the SMGC antipodal bowtie in Fig. 3.51(b) has a 5 to 10 dB better impedance matching at 3 to 6 GHz when compared to conventional shape. Fig. 3.53(b) shows that the gain has increased in most of the operating bandwidth. The gain of SMGC

based antenna has increased by 0.5 to 1 dB from 3.5 GHz to 5.5 GHz and 7 GHz to 10 GHz compared to the conventional structure.

Figure 3.54: Comparison between conventional and SMGC shaped bowtie slot antenna. (a) S_{11} . (b) Half-power beamwidth ($^{\circ}$).

Fig. 3.54(a) shows that the SMGC bowtie slot in Fig. 3.51(c) has a cut-off frequency reduced by 7.5% and bandwidth increased by 26.44%, observed at -15 dB. Fig. 3.54(b) shows that the half-power beamwidth (HPBW) of SMGC based bowtie slot has increased in the operating bandwidth. At least 5° to 10° improvement in HPBW is observed for SMGC based slot as seen in Table 3.6.

The SMGCs have been used in the design of planar antennas such as microstrip fed or CPW-fed antennas. Designs that require SMGCs with special case of equal scaling ($S_{x1} = S_{x2} = S_y$) and unequal scaling ($S_{x1} \neq S_{x2} \neq S_y$) have been discussed in the design of various cases of planar antennas. In section 3.3, an unidirectional antenna is discussed where the lower cut-off frequency of the S_{11} can be slightly reduced using SMGC. Similarly, in section 3.4, a slot antenna with tuning stub is designed, and using SMGCs can also slightly reduce the cut-off frequency. In section 3.5, a mushroom-shaped monopole is designed using SMGC as it can help generate a complex shape profile. It can also improve the impedance matching appreciably and help to reduce the lower cut-off frequency. In section 3.6, an antipodal bowtie antenna is designed using SMGC, and

it can provide a better impedance matching and a slight improvement in gain. Most importantly, SMGCs can offer a significant improvement in the performance of bowtie slot antennas with CPW-feed, as it can reduce the cut-off frequency to a large extent and also show a considerable increase in the gain, without any increase in the antenna size.

Using SMGCs for the generating the edge or shape profile of planar conductors such as the microstrip or the slot can provide the following

- SMGCs can provide additional degrees of freedom in the design of an antenna.
- SMGCs can help in exploring additional possibilities of spatial geometry that can be taken advantage for realizing new electrical behaviour of antenna.
- SMGCs can alter the edges of the conductor or the slot, thereby influencing the surface currents on the conductor edges which in turn will affect the antenna impedance bandwidth, input impedance and radiation gain.
- SMGCs overcome the shortcomings of conventional Gielis curves by avoiding discontinuities of the curve profile while scaling.

Chapter 4

Improved Designs of Power Dividers Using Modified Gielis Curves

4.1 Introduction

In many applications, such as circular array of planar antennas there is a necessity to design a PD network to be integrated on the same substrate as the planar antennas and the feed has to be located centrally from the circular array elements for in-phase operations. Hence, there is a need to have perpendicular coax to microstrip transition. Therefore for wideband applications, the perpendicular coax to microstrip transition has to support wideband operation.

4.2 Center-fed four-way power divider based on SIW and modified Wilkinson structures

In this section, a 1 to 4-way power divider is designed in two stages [135]. The first stage is a type of coax to SIW transition, which also acts as a 2-way PD. The second stage is a modified Wilkinson PD with frequency-selective coupling structures (FSCS) and short-ended meander stubs. The novelty is using an SIW based structure for per-

Figure 4.1: Structure of the 1 to 2 way SIW power divider (a) top view (b) side view of coaxial feed.

pendicular coaxial feed, i.e., when the feed cannot be placed at the edge of the substrate, for designing a broadband 1 to 4-way divider. The design and the S-parameters of the 2 power dividers will be discussed in subsection 4.2.1 and 4.2.2. The arrangement of the first and second stages of the PDs to form a 4-way PD and its S-parameters will be discussed in subsection 4.2.3. The PD designs are based on the FR-4 dielectric substrate with dielectric permittivity $\epsilon_r = 4.3$ and thickness of 1.6 mm as it could be cheap and easy to fabricate. The loss tangent of the substrate is $\delta = 0.025$.

4.2.1 Coaxial to SIW based two-way wideband power divider

The SIW based two-way PD fed by perpendicular coax has been modified from [130] and designed on the FR-4 substrate for wideband operation. This coax to SIW structure serves as a transition from the coaxial input and also as a power divider. The wedge-shaped triangle is implemented to obtain the $1/32^{th}$ mode inside the SIW structure. From Fig.4.1, the metal vias are placed along the end of the triangle and are shorted to the ground to create an electric wall. The hypotenuse and the adjacent sides form a magnetic wall. The bottom ground plane has a hole of diameter r_4 at the feeding point. The circular plate of diameter r_1 , r_4 and L_2 influences the impedance matching as shown in Fig. 4.1(b). The length of the triangular wedge L_1 affects the operating

ADS modeling of SIW based 1st stage PD

The SIW based 1st stage PD is modeled in ADS (Advanced Design Software) schematic circuit for initial design as shown in Fig. 4.2(a). In the schematic model, the SIW PD is modeled using tapered transmission lines that are shorted at the end and a coaxial transmission line at the center. The S_{11} , S_{21} , and S_{31} are given in Fig. 4.2(b), where the operating center frequency is around 2.8 GHz having a bandwidth range from 2.1 GHz to

Figure 4.3: Simulated and measured S parameters of SIW based two-way power divider. (a) Fabricated prototype. (b) Reflection and Transmission coefficient. (c) Output reflection and isolation. (d) Transmission phase [superscript 's' denotes simulation, 'm' denotes measurement].

3.4 GHz. The insertion loss is around 3.5 dB. The S_{22} and S_{32} are shown in Fig. 4.2(c), but they are still higher than -10 dB. The phase at the output ports is almost the same as shown in Fig. 4.2(d). The ADS model is done to evaluate the preliminary design. To perform a fine analysis, a full-wave simulation is performed.

The fabricated prototype of the first stage PD, which is the coaxial to SIW based PD, is given in Fig.4.3(a). The simulated and measured S parameters S_{11} , S_{21} , S_{22} are given in Fig. 4.3(b). The -10 dB impedance bandwidth of S_{11} is from 1.7 GHz to 4 GHz. The electric field components fed by the coaxial input are converted to SIW mode, which is shown by better impedance matching. The transmission coefficients (S_{12} and S_{13}) are observed to be around -3.5 dB. To improve the output port matching and isolation, it

is cascaded with a second stage. From Fig. 4.3(c), the output reflection coefficient S_{22} and the isolation coefficient S_{32} are above -10 dB in the frequency range of interest. In Fig. 4.3(d), it is noted that the power arrives in phase at the two output ports of the SIW based PD, thereby having a negligible phase difference. Hence the SIW based PD serves as a better transition structure for coaxial to microstrip conversion, especially if the coax is fed perpendicular to the substrate. The fractional bandwidth of this SIW based power divider is 80.7%.

4.2.2 Modified Wilkinson wideband power divider with frequency selecting coupling structure and short-ended meander stubs

The second stage is implemented using a two-way power divider using a modified Wilkinson wideband power divider with FSCS and short-ended meander stubs for harmonic suppression. The structure is designed on the FR-4 board using the guidelines from [125]. Fig.4.4 shows the structure of Wilkinson PD with FSCS and short-ended meander line

Figure 4.4: Structure of Wilkinson PD with FCFS and short-ended meander line

stubs. The FSCS unit comprises of the coupled lines of width cw and length cl connected in series with a stub of width sw and length sl . It is inferred from [125], that the FSCS unit helps in suppressing the harmonics for a very wide bandwidth and the short-ended

meander lines help in providing an extra transmission zero and also block the DC component. The meander lines are shorted using a metal via of diameter 0.6 mm, and there could be a slight coupling between the two meander lines. The dimensions of the modified Wilkinson PD is obtained as $w1 = 3.1$ mm, $w2 = 1.65$ mm, $L4 = 3.3$ mm, $sp = 2$ mm, $cs = 0.3$ mm, $cw = 0.3$ mm, $cl = 4$ mm, $sw = 2.2$ mm, $sl = 4$ mm, $mw = 1.2$ mm, $ms = 0.2$ mm, $ml1 = 8.2$ mm, $ml2 = 4.4$ mm, $R = 100 \Omega$.

The authors would like to clarify why the S_{11} spectrum is shown from 1 to 10 GHz. An illustration is provided by referring to a graph plot from [121] as shown in Fig. 4.5. It

Figure 4.5: S_{11} of the two-way power divider from [121]

can be seen from Fig. 4.5, that even though the operating frequency range of the power divider is from 1.6 GHz to 2 GHz, the S_{11} spectrum is shown from 0 to 9 GHz. The fundamental operating frequency is $f_o = 1.8$ GHz. Hence, a power divider with a good harmonic suppression is expected to suppress at least the next few harmonics, i.e., $2f_o$, $3f_o$, $4f_o$, etc. Therefore, if $f_o = 1.8$ GHz, then it is expected to show S_{11} up to $4f_o$, i.e., around 7.2 GHz. In Fig. 4.5, S_{11} is shown up to 9 GHz to show that harmonics are suppressed up to $4f_o$, i.e., $|S_{11}| < -10$ dB.

Similarly, to show the harmonic suppression of a particular power divider in the thesis, their S_{11} is shown up to 10 GHz. So two plots are given for S_{11} , where one plot will show the frequency range of 0 to 10 GHz to see the harmonic suppression characteristics and

another plot to show only the operating range 2 to 3 GHz for clarity.

Figure 4.6: Simulated and measured S parameters of modified Wilkinson PD. (a) Fabricated prototype. (b) Reflection and Transmission coefficient. (c) Output reflection and isolation. (d) Transmission phase. (e) Reflection and Transmission coefficient (1 to 4 GHz). [superscript 's' denotes simulation, 'm' denotes measurement].

The fabricated prototype of the second stage modified Wilkinson PD is shown in Fig. 4.6(a). The S parameters S_{11} , S_{21} of the PD is shown in Fig.4.6(b), and it can be

seen that the harmonics are suppressed over a very wide bandwidth due to the presence of the FSCS unit and short-ended meander stubs. It can be seen that the S_{12} and S_{13} are less than -10 dB for frequencies greater than 5 GHz. The -10 dB input impedance is from 1.35 GHz to 3.4 GHz.

The transmission coefficients for both the output ports are around -3.5 dB, and it is maintained throughout the entire operating bandwidth. From Fig. 4.6(c), the output reflections S_{22} and S_{23} is also less than -10 dB for most of the operating bandwidth. The isolation between ports 2 and 3 are also less than -10 dB. The phase at the output ports from the input port is given in Fig.4.6(d). The phase is almost the same at both the ports, which indicates minimal phase imbalance. The fractional bandwidth of the modified Wilkinson with FSCS and short-ended meander stub is calculated to be about 86.32%. A plot showing S_{11} for the operating frequency range of 1 to 4 GHz given in Fig. 4.6(e) is added for clarity.

This SIW based power divider in subsection 4.2.1 is just the stage I power divider of the whole four-way PD discussed in subsection 4.2.3. This SIW-based two-way PD is first analyzed before proceeding to the four-way PD design. The output port reflection and isolation is indeed not very good, which is why this SIW-based PD is cascaded with the modified-Wilkinson based PD as stage II, which is discussed in subsection 4.2.2. The combination of these stages I and II can give a better output port reflection and isolation for the four-way PD.

4.2.3 Simulation and discussion of 1 to 4 Way wideband power divider

The outputs of the first stage, which is the coax to SIW based power divider is cascaded with the second stage of Wilkinson power divider with FSCS and short-ended meander

Figure 4.7: 1 to 4 way power divider using SIW based power divider and modified Wilkinson power divider using FSCS

stubs, as shown in Fig. 4.7 to form the 1 to 4 Way Wideband PD and the simulation results are discussed. Transmission lines of lengths $L5$, $L6$, and $L7$ are used to connect the two stages and are having a characteristic impedance of 50Ω . The two stages are integrated on the same FR-4 substrate with a thickness of 1.6 mm.

Some dimensions of the overall structure is given as $L = 46$ mm, $W = 80$ mm, $L5 = 8.55$ mm, $L6 = 9.45$ mm and $L7 = 8.45$ mm. The spacing between the coupled line cs is slightly tuned to 0.2 mm to improve the impedance matching of the overall divider.

A prototype of the fabricated four-way power divider after the cascading is shown in Fig. 4.8(a). The simulated input reflection S_{11} and transmission S_{21} of the designed 4-way divider is given in Fig.4.8(b). The -10 dB impedance bandwidth at the input port is from 1.6 GHz to about 3.5 GHz. The power transmitted to the four output ports is around -6.5 dB. The insertion loss could also be due to the loss tangent $\delta=0.025$ of the substrate. The amplitude imbalance is minimal at the four output ports.

The simulated output reflection and isolation at the output ports is shown in Fig. 4.8(c). The reflection at the output port S_{22} is less than -10 dB for the frequency ranges of 2.25 GHz to about 4 GHz. The isolation between the output ports 2 and 3 which are denoted by S_{32} is less than -10 dB for the frequency ranges of 1.75 GHz to 3.5 GHz. The

Figure 4.8: Simulated and measured S parameters of 4-way PD by combining SIW and modified Wilkinson PD. (a) Fabricated prototype. (b) Reflection and Transmission coefficient. (c) Output reflection and isolation. (d) Transmission phase. (e) Reflection and Transmission coefficient (1 to 4 GHz). [superscript 's' denotes simulation, 'm' denotes measurement].

phase of the power at the output ports 2, 3, 4 and 5 are observed to be in phase from Fig.4.8(d). The difference in phase between the output ports are negligible.

The frequency range of 2 to 4 GHz has been chosen as it can be more suitable for

supporting WiMAX technology. WiMAX for nomadic application (frequency band of 2.3-2.4 GHz) and mobile application (frequency band of 3.3 to 3.7 GHz) is likely to be a prominent candidate for wireless data in the near future. Therefore, there is a need to develop a multibeam horizontally polarized antenna array that can support both WiMAX applications including frequencies 2.3 GHz and 3.3 GHz. Hence, the power divider with 2 to 4 GHz is designed so that it can be integrated with the antenna array.

The input power from a perpendicular coax is fed to an SIW structure operating at $1/32^{th}$ mode, which is in turn fed to a modified Wilkinson for a 4-way power division operation for wideband application. The calculated fractional bandwidth of the overall 4-way PD after integrating the first and second stages of PD is about 74.5%. The proposed divider can be used as a feeding network for array antennas, especially when the feed is supposed to be placed at the center and perpendicular to the substrate. For example, in the design of horizontally polarized omnidirectional antennas involving Vivaldi antennas with common ground plane [86], the proposed feeding network can be used as an alternative for feeding antenna arrays in wideband applications.

Power dividers are used in the front-end along with the antenna in any wireless communication system. The even- and odd-mode analysis employed in a conventional Wilkinson power divider, it can be observed that high order harmonics will introduce spurious passbands, which affects the performance of the circuit drastically, especially when the divider is applied in wireless systems. For example, antennas with power divider operating at 2.4 GHz may also receive spurious signals at harmonic frequencies, which are multiples of 2.4 GHz, i.e., 4.8 GHz, 7.2 GHz, 9.6 GHz, etc., which is undesired.

Figure 4.9: S_{11} of power dividers from (a) [122], (b) [123], and (c) [124] showing frequency range out of operating band to show harmonics.

The S_{11} of some power dividers with harmonic suppression have been shown in Fig. 4.9, from [122], [123], and [124]. It can be seen that the S_{11} spectrum is shown at least up to 10 GHz to convey that the 2^{nd} , 3^{rd} or even 4^{th} harmonics have been suppressed.

In this section, a 1 to 4-way wideband power divider using substrate integrated waveguide and modified Wilkinson structures has been presented. Two stages of a 2-way power divider and their arrangement to form a 4-way power divider for feeding input perpendicular to the substrate have been discussed. The simulated transmission coefficients, reflection coefficients, and isolation for the 2-way power dividers and the 4-way power dividers have also been discussed. The designed 4-way power divider with substrate integrated waveguide and modified Wilkinson has a -10 dB impedance from 1.6 GHz to 3.5 GHz having a fractional bandwidth of 74.5%. However, the harmonic suppression of the above four-way PD needs to be improved, which will be discussed in the next section.

4.3 Geometry of stubs using SMGCs

Table 4.1: SMGC parameters for conventional geometries in stubs

SMGCs	m_1	m_2	n_1	n_2	n_3	S_{X1}	S_{X2}	S_Y	c_x	c_y
Rectangle	4	4	1000	1000	1000	1.5	1.5	3	0	0
Taper 1	3	3	6.75	12	12	3	2	1	0	1
Taper 2	3	3	6.75	12	12	3	2	1	0	-1

The shape of the stubs can greatly influence the transmission zeros or the S_{21} char-

Figure 4.10: (a) Type of open stubs using conventional geometries. (b) Shape profiles of the conventional geometries and the SMGC shape based geometries. (c) S_{21} of open stubs for various conventional geometries.

acteristics of the stubs. To illustrate the importance of stub geometry, three types of stubs are considered to be placed across a two-port transmission line of characteristic impedance 50Ω . The stubs are built on an FR4 substrate with permittivity 4.3 and thickness 1.6 mm. The ground plane sizes are the same, and the lengths of the stubs are the same as shown in Fig. 4.10(a). Three types of stubs are involved, namely ‘rectangular’, ‘taper 1’, and ‘taper 2’. The shape profiles of these conventional geometries are shown in Fig. 4.10(b). Also, the shape profiles of the SMGCs that are equivalent to that of the conventional geometries are shown and labeled. It should be noted from Fig. 4.10(c), that the transmission zero or the lowest S_{21} occurs at 1.75 GHz for ‘rectangular’ stub, 1.42 GHz for ‘taper 1’ stub, and 2.16 GHz for the ‘taper 2’ stub. The above observation emphasizes that even if the stubs are of the same length and similar size, the type of tapering or non-tapering profiles can shift the frequency (at which the transmission zeros occur) to a great extent. But multiple geometries for the stub need more variables, and the control of those variables is cumbersome. To avoid these problems, SMGCs can be used to generate many shapes for the stubs, including these conventional geometries.

The parameters of the curves SMGC1, SMGC2 and SMGC3 to produce the shapes of stubs with ‘rectangular’, ‘taper 1’, and ‘taper 2’ geometries as shown in Fig. 4.10(b) are given in Table 4.1 using (3.1) and (3.2). To get the SMGC plots similar to Fig. 4.10(b), the SMGC curves should be plotted as follows: SMGC1 (X_{SMGC1} , Y_{SMGC1}), SMGC2 ($-Y_{SMGC1}$, X_{SMGC1}), SMGC3 ($-Y_{SMGC1}$, $-X_{SMGC1}$) using (3.2).

4.4 Improved center-fed four-way power divider with modified Gielis curves for wideband harmonic suppression

Figure 4.11: Simulated S -parameters (a) S_{11} and S_{21} and (b) S_{22} and S_{32} of the four-way PD from [135].

In our previous work [135] that was discussed in the previous section, a four-way PD is designed using substrate integrated waveguide and modified Wilkinson structures, but the harmonics appear at 5.5 GHz, and the S_{22} is not less than -10 dB over the operating frequency range of 1.6 GHz to 3.5 GHz. Very few works have been done in center-fed or perpendicular coax-fed four-way PD. Considering the PD in the previous section, the proposed PD has some novelty herein:

1. Sectorially modified Gielis curves (SMGCs) are used to form the tapered open and

short stubs along with coupled-line structures to provide easy modifications in the tapering profile and additional degrees of freedom (DOF) while designing.

2. Open or short stubs are used to influence the position of transmission zeros, thereby suppressing the harmonics (at the region of $2f_o$ and $3f_o$) over a wideband than many other planar four-way PDs such as [94, 97, 108, 130].
3. A stub-loaded coupled line structure is used to suppress harmonics at $4f_o$.
4. An expression for the ABCD matrix of the stub-loaded coupled line structure with different even and odd mode electrical lengths ($\theta_e \neq \theta_o$) is presented.
5. The proposed center-fed four-way PD can provide a wideband harmonic suppression of up to $4f_o$ as compared to [94, 97, 108, 130, 135].

In this section, a center-fed four-way PD with wideband harmonic suppression is proposed. The proposed four-way PD is designed in two stages. The first stage is a center-fed two-way PD with stubs. The second stage is a modified two-way Wilkinson PD with coupled-line structures. The stubs and coupled-line structures are useful for harmonic suppression, while the SMGCs are used to form the profiles of the stubs and coupled-line structures. In subsections 4.4.2, 4.4.3, 4.4.4, 4.4.5, the theoretical analysis and design of the two stages of proposed four-way PD will be discussed in detail, along with the effect of stubs on harmonic suppression. The four-way PD is then analyzed with full-wave simulation, and its results are compared with the measured results in subsection 4.4.6.

4.4.1 Design and analysis of four-way power divider with wideband harmonic suppression

The layout of the proposed four-way PD is shown in Fig. 4.12(a). The harmonic suppression proposed in this work is basically created by introducing two transmission zeroes

Figure 4.12: Proposed four-way PD (a) Overall circuit layout and (b) Overall schematic of equivalent circuit.

between 3 GHz and 9 GHz, while the desired operating bandwidth is approximately between 1 to 2 GHz. Hence, the center frequency of operating bandwidth is f_1 , and the center frequency of the stopband between 3 to 9 GHz is given as $f_2 = 6$ GHz. Three structures are involved in creating the harmonic suppression, namely (1) a tapered open stub to produce the first transmission zero between 3 and 6 GHz, (2) a tapered short stub to produce the second transmission zero between 6 and 9 GHz, and (3) a coupled line with loaded stubs to suppress harmonics between 7 GHz to 9 GHz as shown in the schematic in Fig. 4.12(b). The physical dimensions of the first and second stage marked by a dashed red box in Fig. 4.12(a) will be later discussed.

Figure 4.13: Effects of SMGC profiles on the stubs. (a) Structure, (b) SMGC profiles, (c) S_{21} for open stub ($m_1=m_2=1, n_1=0.5, n_2=n_3=0.3, S_{X2}=15$); (d) Structure, (e) SMGC profiles, (f) S_{21} for shorted stub ($m_1=m_2=3$); (g) Structure, (h) SMGC profiles, and (i) S_{21} for coupled-line structure ($m_1=m_2=1, S_{X1}=2, S_{X2}=6, S_Y=8$)

4.4.2 Design of open stub, shorted stub and loaded coupled-line structure using SMGCs

In order to have a smooth transition of the impedance in the tapered stub arms, the shape profiles of the stubs are generated using (3.1) and (3.2). Based on various SMGC parameters, the different forms of curves that are used for shaping the stubs are shown in Fig. 4.13.

Effects of SMGCs on open stub

The stub profile is described by a few SMGC curves whose parameters are used for better tuning of the S parameters of the open stub, as shown in Fig. 4.13(a)-(c). The simulation setup for the open stub is shown in Fig. 4.13(a) with $50\ \Omega$ port impedance. The shape and size of the open stub for only few pairs of S_{x1} and S_y are shown in Fig. 4.13(b) for brevity, without changing n_1 , n_2 , and S_{x2} variables. Fig. 4.13(c) shows the variations of S_{21} for the corresponding values of S_{x1} and S_y . Such open stubs are used to influence the position of transmission zeros, thereby suppressing the harmonics at $2f_o$. It can be noted that even for small changes in the shape profile, the location of the transmission zero changes noticeably, even though the length of the stub is not changed.

Effects of SMGCs on shorted stub

The structure of the tapered shorted stub for second transmission zero is shown in Fig. 4.13(d) with $50\ \Omega$ ports. The shorted via has a diameter of 0.5 mm and distance 0.5 mm between them. With the length of the shorted stub almost fixed, the frequency at which the transmission zero occurs changes appreciably even for a slight change in the width of the profile in Fig. 4.13(e)-(f). The short stubs are used to influence the position of transmission zeros to suppress the harmonics at around $3f_o$.

It can also be seen that the transmission zeroes of the shorted stub can be easily tuned using SMGCs as even small changes in its parameters can influence the transmission zero.

Effects of SMGCs on loaded coupled-line structure

The coupled-line having width 0.4 mm, spacing 0.4 mm, and length 5 mm are loaded with stubs using SMGC profiles, as shown in Fig. 4.13(g) with $70\ \Omega$ port impedance. The SMGC profiles are used for the loaded stubs to form a smooth tapering and better tuning, as shown in Fig. 4.13(h) instead of using a rectangular shaped loaded stubs. Fig. 4.13(i)

shows that for smaller n_1 and n_2 values, greater is the frequency above which stopband characteristics occur. The other parameters $m_1 = m_2$, $S_{X1} = S_{X2}$ and S_Y can also be tuned for different characteristics of the coupled-line stub but are not discussed here for brevity. In Fig. 4.13(i), it can be seen that it can suppress the harmonics from 7 GHz until around 8.5 GHz as the S_{21} becomes lower. The stub-loaded coupled line structure is used to suppress harmonics at $4f_o$ and, therefore, can enhance the harmonic suppression in higher frequency ranges in addition to the transmission zeros provided by the open and short stubs, respectively.

SMGCs are used to form the tapered open and short stubs along with coupled-line structures to provide easy modifications in the tapering profile and additional degrees of freedom (DOF) while tuning. The proposed four-way PD consists of two stages, i.e., center-fed two-way PD and modified Wilkinson PD, as shown in Fig. 4.12(b). Even and odd mode analysis are performed on the PD to characterize its performance. The even and odd bisected schematics are shown in Fig. 4.14, respectively, with appropriate terminations.

4.4.3 Even mode analysis

The schematic circuits for the even mode analysis of the two stages of the proposed four-way PD are shown in Fig. 4.14(a) and (b) based on the overall circuit in Fig. 4.12. Under even mode operation, the input and output terminations are $Z_{01} = 2Z_0$ and $Z_{02} = Z_0$, respectively, where Z_0 is the port impedance. The open stub with the SMGC profile is approximated with sub-sections of TLs, as shown in Fig. 4.14(a). Each stub is modeled with sub-sections of transmission lines having an equal electrical length and impedance profile following that of the SMGC profile along its length. The input impedance $Z_{in,OC}^{SMGC}$ of the open stub is obtained by calculating the total input impedance of cascaded sub-

Figure 4.14: Schematic for (a) Even mode 1st stage, (b) Even mode 2nd stage, (c) Odd mode 1st stage, and (d) Odd mode 2nd stage of the proposed four-way PD.

Figure 4.15: Port numbering for derivation of ABCD matrix of coupled lines loaded with stub.

sections successively as

$$Z_{in\{n\}} = Z_n \frac{Z_{in\{n-1\}} + j Z_n \tan \theta_n}{Z_n + j Z_{in\{n-1\}} \tan \theta_n} \quad (4.1)$$

where n is the sub-section number in a tapered section. A similar approach is used to obtain the input impedances $Z_{in,SC}^{SMGC}$ for shorted tapered stub and $Z_{L,CL}^{SMGC}$ for the loaded stub in coupled line structure. A series tapered section with characteristic impedance ranging from Z_{t1} to Z_{t2} of electrical length θ_{t1} and a shorted tapered stub with characteristic impedance ranging from Z_{s1} to Z_{s2} of electrical length θ_{s1} are also used. The junction between the series tapered section and the shorted tapered stub produces a junction capacitance indicated by equivalent capacitance C .

In the 2nd stage, the input impedance $Z_{in,e2}$ appears as a load impedance on the port I of the coupled-line structure with stubs having SMGC profile as labeled in Fig. 4.14(b).

The coupled line circuit with loaded stub is shown in Fig. 4.15, based on which the ABCD for coupled lines loaded with stubs can be derived. The ports 1', 2', 3', 4' are actual port convention of the coupled lines, but ports 1' and 3' are seen as ports 1 and 2 when considering the coupled lines as a part of the entire power divider. The stub can be accounted as an impedance $Z_{L,CL}^{SMGC}$ loaded to the two adjacent ports of coupled lines. The voltage and current equations for a coupled line with a load $Z_{L,CL}^{SMGC}$ connected between ports 2 and 3 can be written as

$$\begin{aligned}
V_1^e - \cos(\theta_e) V_2^e + j Z_{oe} \sin(\theta_e) I_2^e &= 0 \\
I_1^e - j \frac{\sin(\theta_e)}{Z_{oe}} V_2^e + \cos(\theta_e) I_2^e &= 0 \\
V_1^o - \cos(\theta_o) V_2^o + j Z_{oo} \sin(\theta_o) I_2^o &= 0 \\
I_1^o - j \frac{\sin(\theta_o)}{Z_{oo}} V_2^o + \cos(\theta_o) I_2^o &= 0 \\
V_1 - V_1^e - V_1^o &= 0 \\
V_2 - V_2^e - V_2^o &= 0 \\
V_4 - V_2^e + V_2^o &= 0 \\
I_1 - I_1^e - I_1^o &= 0 \\
I_2 - I_2^e - I_2^o &= 0 \\
I_4 - I_2^e + I_2^o &= 0 \\
V_2^e - 2I_2^e Z_{L,CL}^{SMGC} &= 0 \\
V_2^o &= 0 \\
V_3 &= V_1^e - V_1^o \\
I_3 &= I_1^e - I_1^o
\end{aligned} \tag{4.2}$$

These stubs are also approximated with sub-sections of the transmission line for derivation. The expression for the $ABCD$ matrix of the stub-loaded coupled line structure with different even and odd mode impedances (Z_{oe}, Z_{oo}) and electrical lengths ($\theta_e \neq \theta_o$) is given as follows.

$$[ABCD]_c = \begin{bmatrix} An/d & Bn/d \\ Cn/d & Dn/d \end{bmatrix}$$

$$\begin{aligned} d &= Z_{oe} Z_{oo} \cos\theta_e \sin\theta_o - Z_{oe}^2 \cos\theta_o \sin\theta_e \\ &\quad + 2j Z_{in,3} (Z_{oe} \cos\theta_e \cos\theta_o \\ &\quad + Z_{oo} \sin\theta_e \sin\theta_o) \\ An &= -Z_{oe}^2 \cos\theta_o \sin\theta_e - Z_{oe} Z_{oo} \cos\theta_e \sin\theta_o \\ &\quad + 2j Z_{in,3} (Z_{oe} \cos\theta_e \cos\theta_o \\ &\quad - Z_{oo} \sin\theta_e \sin\theta_o) \\ Bn &= -2 Z_{oe} Z_{oo} \sin\theta_o (2Z_{in,3} \cos\theta_e \\ &\quad + j Z_{oe} \sin\theta_e) \\ Cn &= 2 \cos\theta_o (j Z_{oe} \cos\theta_e - 2 Z_{in,3} \sin\theta_e) \\ Dn &= An \end{aligned} \tag{4.3}$$

where $Z_{in,3}$ is calculated similarly from (4.1). Having found the $ABCD$ matrix, the input impedance $Z_{in,e5}$ can be found using voltage and current relations as follows

$$\begin{aligned} V_{II}^e / I_I^e &= -(D Z_{in,e2} + B), \\ I_{II}^e / I_I^e &= -(C Z_{in,e2} + A), \\ Z_{in,CL(e)}^{SMGC} &= V_{II}^e / I_{II}^e \end{aligned} \tag{4.4}$$

where V_I^e , I_I^e and V_{II}^e , I_{II}^e are the even mode voltages and currents facing the port I and II

of the stub-loaded coupled-line structure as in Fig. 4.14(b). Considering the output port impedance to be $Z_{02} = Z_0$, the S_{11} parameter for the even mode is given by

$$S_{11,e} = \frac{Z_{in,e3} - Z_{02}}{Z_{in,e3} + Z_{02}} \quad (4.5)$$

It is observed that the total S_{11} of the entire four-way PD is same as $S_{11,e}$ obtained from (4.5).

The transmission coefficient S_{21} can be obtained by forming the $ABCD$ matrices for the 1st and 2nd stage of the PD. The S_{21} of the 1st and 2nd stage are denoted as S_{21}^{1st} and S_{21}^{2nd} which are converted from their $ABCD$ matrices respectively, with their appropriate port terminations. The total S_{21} of the PD in the even mode is given by

$$\begin{aligned} S_{21,e} &= \frac{(S_{21}^{1st}/\sqrt{2}) \times (S_{21}^{2nd}/\sqrt{2})}{1 - S_{22}^{1st} S_{11}^{2nd}} = \frac{1}{\sqrt{2}} S_{21}^{1st} \times \frac{1}{\sqrt{2}} S_{21}^{2nd} \\ &= 0.5 S_{21}^{1st} S_{21}^{2nd} \end{aligned} \quad (4.6)$$

which is the total S_{21} of the four-way PD.

4.4.4 Odd mode analysis

The schematic circuits for the odd mode analysis of the two stages of the proposed four-way PD are shown in Fig. 4.14(c) and (d). Under odd mode operation, the input port becomes short-circuited, and hence the odd mode circuit does not influence the S_{11} . The odd mode analysis can be used in determining the isolation resistor between the output ports. In the odd mode analysis of the 2nd stage, the $ABCD$ matrix of loaded coupled-lines from (4.2), (4.3), along with $Z_{in,o4}$ can be used to find the input impedance as

follows

$$\begin{aligned}
Z_{in,o2} &= j Z_2 \tan \theta_2 \\
V_{II}^o / I_I^o &= -(D Z_{in,o2} + B), \\
I_{II}^o / I_I^o &= -(C Z_{in,o2} + A), \\
Z_{in,CL(o)}^{SMGC} &= V_{II}^o / I_{II}^o.
\end{aligned} \tag{4.7}$$

Subsequently from $Z_{in,o3}$, the input impedance $Z_{in,o4}$ at the output of the 2nd stage in the odd mode can be given by

$$Z_{in,o4} = \frac{Z_{iso} Z_{in,o3}}{Z_{iso} + 2 Z_{in,o3}} \tag{4.8}$$

where $Z_{iso} = R + jX$ is the isolation impedance between ports 2 and 3. Z_{iso} can be obtained by equating $Z_{in,e3}$ and $Z_{in,o4}$ from which the isolation resistance R can be subsequently found as follows

$$R = \frac{2 Z_{in,e3} Z_{in,o3}}{Z_{in,o3} - Z_{in,e3}} \tag{4.9}$$

The theoretical S_{11} and S_{21} are calculated from (4.5) and (4.6) respectively using the tuned characteristic impedance and electrical length values shown in Table 4.2.

4.4.5 Effects of stubs on the harmonic suppression from S_{11} and S_{21} response

Effects of impedance and electrical length of the stubs on S_{11}

To further understand the effects of the stub, some parameters of the stubs are varied, while other parameters used are the same as from Table 4.2 shown in Fig. 4.16. From Fig. 4.16(a), it is observed that with greater the range between the characteristic impedance of the open stubs Z_{o1} and Z_{o2} , the undesired resonance at around 3 GHz

Table 4.2: Parameters of the Four-Way PD

Z, θ, f	Z_{o1}	Z_{o2}	$\theta_{o1}@f_1$	Z_{t1}	Z_{t2}
Phy.*	$w_{a5}=0.66$	$w_{a4}=1.35$	$l_{a5}=11.2$	$w_{a2}=0.63$	$w_{a3}=2$
Elec.	103	77.6	48.3	105	64.1
Z, θ, f	$\theta_{t1}@f_1$	Z_{s1}	Z_{s2}	$\theta_{s1}@f_1$	Z_1
Phy.	$l_{a1}=9.45$	$w_{a2}=2$	$w_{a3}=4.8$	$l_{a2}=12.5$	$w_1=3.1$
Elec.	36.4	64	37.8	48.1	50
Z, θ, f	$\theta_1@f_1$	Z_2	$\theta_2@f_1$	C	Z_{L1}
Phy.	$l_1=31.4$	$w_{b2}=1.6$	$l_{b1}=5.4$		$w_{b4}=1.7$
Elec.	121	71.7	20.6	1.1	69.6
Z, θ, f	Z_{L2}	$\theta_L@f_2$	Z_{ec}	Z_{oc}	θ_e, θ_o $@f_2$
Phy.	$w_{b5}=0.1$	$l_{b2}=1.8$	$w_{b3}=0.4,$ $s_{b3}=0.4$	$w_{b3}=0.4,$ $s_{b3}=0.4$	$l_{b3}=4.6$
Elec.	172.7	23.3	171.1	71.5	58.78, 54.16
Z, θ, f	f_1	f_2	R	ε_r	h
Phy.			100		1.6
Elec.	1.8	6	100	4.3	

units: w, l, h (mm), Z (Ω), θ (deg), C (pF), f (GHz).

*Physical dimensions calculated for conductors with zero thickness.

becomes lower. The tapered open stub at the input A has impedance ranging from 100 Ω to 70 Ω and the initial electrical length θ_{o1} is chosen as $\lambda/8$ at operating center frequency $f_1 = 1.8$ GHz to provide the 1st transmission zero. The variation in the electrical length of the open stubs θ_{o1} and θ_{o2} affects the bandwidth and also the resonance at around 3 GHz as shown in Fig. 4.16(b). The tapered shorted stub at junction B has impedance ranging from 50 Ω to 30 Ω , and the initial electrical length $\theta_{s1} = \lambda/8$ at f_1 is chosen based on tuning to provide the 2nd transmission zero. It is also observed from Fig. 4.16(c) that choosing the appropriate length for the shorted stub θ_{s1} helps in suppressing the harmonics around 7 GHz. Choosing the length of the parallel TL stubs θ_L and the coupled-line section θ_c also plays a role in suppressing the harmonics at around 6 to 8 GHz as shown in Fig. 4.16(d). The TL in the 2nd stage of PD with $Z_3 = 70.7$ Ω is chosen initially as it is a Wilkinson-based PD where $Z_3 \approx \sqrt{2}Z_0$ with relatively smaller electrical length θ_1

Figure 4.16: Effects of (a) open stub impedance, (b) open stub length, (c) short stub length, and (d) loaded coupled-line section length on the S_{11} parameter.

at f_1 based on tuning. The impedances $Z_{ec} = 190 \Omega$ and $Z_{oc} = 60 \Omega$ of the coupled line section are chosen farther apart for better harmonic suppression, with electrical length $\theta_{ec} = \theta_{oc} \approx \lambda/6$ at stopband center frequency f_2 . The impedance of the loaded parallel stubs range from $Z_{L1} = 69 \Omega$ to $Z_{L2} = 172 \Omega$ having electrical length $\theta_L \approx \lambda/20$ at f_2 approximately. SMGC profiles are used to influence the dimensions of loaded parallel stubs for better tuning as it can suppress harmonics between 7 to 8.5 GHz.

Effects of stubs on S_{21} (transmission zeros) and harmonic suppression

Furthermore, by finding the position of transmission zeros with respect to frequency, as shown in Fig. 4.17, the harmonic suppression of the four-way PD can be analyzed.

Figure 4.17: Dependence of transmission zero locations on the electrical length of (a) Open stubs θ_{o1} , θ_{o2} , (b) Shorted stubs θ_{s1} , (c) Loaded coupled-line structure θ_L/θ_c with $\theta_c = \theta_{ec} = \theta_{oc} = 60^\circ$, and (d) Effects of coupled-line length θ_c on S_{21} ($\theta_L = 16^\circ$).

Fig. 4.17(a) shows the variation of transmission zeros with varying lengths of the open stubs θ_{o1} . The location of the first zero z_1 is mostly influenced by θ_{o1} . Initially, θ_{o1} is chosen to be 48.3° at 1.8 GHz ($\approx 90^\circ$ at 3.6 GHz) such that a transmission zeros z_1 is placed at 3.6 GHz. Similarly, the length of the shorted stub θ_{s1} is chosen to be 48.1° at 1.8 GHz ($\approx 180^\circ$ at 6.9 GHz) so that another transmission zero z_2 is placed 6.9 GHz as shown in Fig. 4.17(b). The length of shorted stub θ_{s1} mostly influences the location of the second zero z_2 . Also, the variation of the transmission zero locations with the length ratio θ_L/θ_c is shown in Fig. 4.17(c), where θ_c is fixed as 60° at 6 GHz. It can be seen that the location of two transmission zeros z_1 and z_2 does not vary with θ_c , but it can be observed from Fig. 4.17(d) that the S_{21} at around 7 GHz is suppressed below -10 dB

Table 4.3: SMGC parameters for open stub and coupled-line structure

G_{mi}	$m_1 = m_2$	n_1	$n_2 = n_3$	S_{x1}	S_{x2}	S_Y
G_1	1	0.5	0.3	2	15	2
G_2	3	5.5	10	9	7	1.6
G_3	1	0.5	0.3	2	6	8

when θ_c is increased from 30° to 60° . The addition of this coupled-line structure does not introduce transmission zeros, but it exhibits a suppression around 8 GHz. Hence a length ratio θ_L/θ_c is chosen to be 0.28 for suppressing harmonics and better impedance matching at operating bandwidth.

4.4.6 Simulation and measurement of the 4-way PD

The electrical parameters of the four-way PD are then converted into their physical dimensions, as shown in Table 4.2 based on the chosen substrate FR4 with dielectric permittivity $\varepsilon_r = 4.3$, loss tangent $\tan(\delta) = 0.025$ and thickness 1.6 mm. The zoomed-in view and dimensions of the 1st and 2nd stage of the four-way PD (marked in Fig. 4.12) are shown in Fig. 4.18.

Fig. 4.18(a) shows the structure of center-fed two-way PD with stubs along with its dimensions. The coaxial inner conductor at port 1 is connected to a circular patch of radius r_{a1} to improve the impedance matching. Two tapered transmission lines originate from the circular patch and split the power to the two output ports. The ends of the tapers are shorted with metallic vias to ground with a diameter 0.5 mm and distance 0.5 mm between them. The location of the output transmission lines on the tapered transmission lines at l_{a1} is based on the length θ_{t1} . The two open stubs (cf. dashed part of Fig. 4.18(a)) are tapered based on impedances Z_{o1} and Z_{o2} . The SMGC profiles G_1 and G_2 used for the open stub and the shorted stub are given in Table 4.3, respectively. The second stage of the four-way PD involves a modified two-way Wilkinson PD, whose

Figure 4.18: (a) Structure of center-fed two-way PD with stubs (1st stage) [$w_{a1} = 3.1$, $w_{a2} = 0.6$, $w_{a3} = 4.8$, $w_{a4} = 1.3$, $r_{a1} = 3$, $l_{a1} = 23$, $l_{a2} = 9.45$, $l_{a3} = 8.55$, $l_{a4} = 1$, $l_{a5} = 11$] and (b) Structure of modified Wilkinson (2nd stage) [$w_{b1} = 3.1$, $w_{b2} = 1.6$, $w_{b3} = 0.4$, $w_{b4} = 5.4$, $l_{b1} = 5.4$, $l_{b2} = 2.7$, $l_{b3} = 5$, $l_{b4} = 3.4$, $s_{b1} = 2.5$, $s_{b2} = s_{b3} = 0.4$ (unit: mm)].

structure is shown in Fig. 4.18(b) based on the physical parameters in Table 4.2. It uses a loaded coupled-line structure based on SMGC to improve the harmonic suppression of the whole four-way PD between 7 GHz to 8.5 GHz. The coupled-line structures of line width w_{b3} and spacing s_{b2} are based on parameters Z_{ec} , Z_{oc} . The SMGC profile G_3 from Table 4.3 is used for the loaded TLs to form a smooth tapering and provide additional DOF for tuning. The inset figure in Fig. 4.18(b) shows the G_3 curve as dashed lines.

The outputs of the center-fed two-way PD with stubs are connected to the inputs of the modified two-way Wilkinson PDs with coupled-line structures using TLs with Z_1 and θ_1 . In Fig. 4.12, the resultant proposed four-way PD is shown. Fig. 4.19 shows the simulated and measured results of S_{11} and S_{21} parameters for the proposed four-way PD along with

Figure 4.19: Simulated and measured results of the input reflection S_{11} and transmission S_{21} coefficients of the proposed four-way power divider.

its fabricated photo in the figure inset. The simulated and measured responses of S_{11} and S_{21} are quite similar in the operating bandwidth, and they follow the curve patterns approximately. It can be seen that the -10 dB bandwidth for the measured S_{11} is from 1.4 to 2.23 GHz with an FBW of 46% while the simulated S_{11} FBW is 52% from 1.35 GHz to 2.3 GHz. The harmonics in the input reflection co-efficient are suppressed over the frequency range from 2.5 to about 8.5 GHz. The proposed four-way PD can provide a wideband harmonic suppression of up to $4f_o$. The measured transmission coefficients ($S_{21}, S_{31}, S_{41}, S_{51}$) in the operating range are between 6.5 dB and 7.2 dB, which is due to the lossy substrate FR-4 and agree closely with that of the simulation results. They are suppressed to below -15 dB over the frequency range from 3 to 8 GHz. Fig. 4.20(a) shows the simulated and measured output reflection coefficients ($S_{22}, S_{33}, S_{44}, S_{55}$) parameters of the proposed four-way PD. The measured output reflection coefficients for the four-way PD is less than -10 dB over the operating frequency range and are similar to simulated output reflection coefficients. The difference in the phase of the measured transmission coefficients ($\angle S_{21} - \angle S_{31}, \angle S_{41} - \angle S_{51}, \angle S_{41} - \angle S_{21}$) are less than 7° . The measured isolation coefficients ($S_{23}, S_{24}, S_{25}, S_{34}, S_{35}, S_{45}$) between the output ports are also less than -10 dB in the operating frequency range and are similar to simulated

Figure 4.20: Simulated and measured results of (a) Output return loss and (b) Isolation between output ports of the proposed four-way power divider.

results as shown in Fig. 4.20(b). It can also be seen that the amplitude imbalances in the transmission coefficients ($||S_{21}| - |S_{31}||$, $||S_{41}| - |S_{51}||$, $||S_{21}| - |S_{41}||$) are less than 5×10^{-3} .

Table 4.4: Comparison of some four-way power dividers

Ref.	FBW (%) ^a	L×W (λ _g) ^b	Harmonics suppressed	Feed type	Structure	Insertion loss (dB)	Power split ratio (P ₂ : P ₃ : P ₄ : P ₅)	Isolation (dB)	Remarks
[100]	22.78	2.46×0.98	NA	EF	SIW	2	1:1:1:1	> 12	unsuitable for low frequency
[101]	54.9	3.87×3.32	NA	EF	SIW	0.7	1:1:1:1	NA	unsuitable for low frequency
[102]	85.71	1.48×1.48	none	EF	P	NA	1:1:1:1	NA	no HS
[111]	100	0.42×0.42	up to 2f _o	EF	P	0.37	1:1:1:1	>10	less HS
[118]	6.28	0.8×0.8	up to 2f _o	EF	P	<1	1:1:1:1	>8	narrow band
[94]	30.92	2.8×2.8	NA	CF	WG	0.25	1:1:1:1	>30 ^c	unsuitable for low frequency
[130]	32	0.62×0.62	none	CF	SIW	<0.9	1:1:1:1	>10.5	no HS
[97]	18.18	3.66×3.66	NA	CF	SIW	<1.5	1:1:1:1	>10	narrow band, larger size
[108]	50	0.3×0.3	up to 2f _o	CF	P	<1	1:1:1:1	>20	less HS
[134]	88.8	0.82×0.82	up to 3f _o	CF	P	<0.3	1:1:1:1	>11	S ₁₁ <-5dB between harmonics
[105]	42.4	1.7×1.7	none	CF	P	1	1:1:1:1	>12	no HS
[104]	14.9	1.4×1.4	none	CF	P	<1	1:1:1:1	>6	no HS
This work	52.05	0.75×0.45	up to 4f _o	CF	P	<1	1:1:1:1	>12	wideband HS

^a -10 dB Bandwidth; ^b guided wavelength $\lambda_g \approx c/(f_{low} \sqrt{(\epsilon_r + 1)/2})$; ^c orthogonal modes; NA: Info Not Available; SIW: substrate-integrated waveguide; CF: center-fed; P: planar; WG: waveguide; HS: harmonic suppression; P₂, P₃, P₄, P₅ are the output power at ports 2, 3, 4, 5.

Table 4.4 shows the comparison of many previous four-way PD and our proposed work. The SIW based cavity dividers in [100], [101] require relatively large sizes of $2.46 \lambda_g$ and $3.87 \lambda_g$ respectively. The matched stripline junction PD [102] does not suppress any harmonics. The Wilkinson PD in [111] can suppress harmonics only up to $2f_o$. Coupled square-ring resonators in PD [118] makes it relatively larger ($0.8 \lambda_g \times 0.8 \lambda_g$) and has narrow bandwidth compared to proposed work. Moreover, all the previously stated designs are end-fed and not center-fed. Thus, they cannot be used for applications such as feeding a circular antenna array. While the work in [94] using a rectangular waveguide cavity is center-fed for dual-polarization PD application, it is not suitable as they are larger with a relatively narrow bandwidth of 30.92%. Other SIW based power dividers [130] and [97] also suffer from relatively narrow bandwidth and it cannot suppress harmonics. The isolation network in [108] requires air-bridge connecting wires making its fabrication complex and can only suppress harmonic components up to $2f_o$. Some previous center-fed four-way PDs are built with stepped impedance lines [134], matched sectorial component [105], and disk configuration [104]. But these PDs exhibit no harmonic suppression and $S_{11} < -5\text{dB}$ between successive harmonics. Since the previous center-fed four-way PD structures exhibit undesired operation at harmonic frequencies, a new four-way power divider is to be designed with wideband harmonic suppression. Our proposed four-way PD is center-fed with a wide operating bandwidth of 52% and provides wideband suppression of harmonics up to $4f_o$.

The overall performance of the 4-way power divider when using conventional simple stubs will be similar to that of the SMGC based stubs. In contrast, the SMGC based stubs will occupy a slightly smaller area than conventional stubs. The role of SMGC stubs in this center-fed four-way PD is that they can offer more degrees of freedom in tuning a stub when compared to conventional stub (width and length).

In order to show the advantage of SMGC for a power divider, a simple illustration is provided based on a two-way PD given in [137].

Figure 4.21: Structure of two-way power divider with (a) conventional stubs [137] and (b) SMGC stubs. $[l_{p1} = 22, l_{p2} = 16, w = 0.65, l_1 = 9.9, l_2 = 7.4]$ (units:mm). Lumped resistor: 92Ω . SMGC parameters $[G_1:m_1 = m_2=1, n_1=0.5, n_2=0.5, n_3=0.5, S_{X1}=3, S_{X2}=12, S_Y=2; G_2:m_1 = m_2=1, n_1=0.5, n_2=0.5, n_3=0.5, S_{X1}=4, S_{X2}=9, S_Y=2]$

Fig. 4.21(a) shows two-way PD with conventional stubs based on work [137] and Fig. 4.21(b) shows SMGC stubs on the same two-way PD. The lengths of the stubs of SMGC are chosen the same as that of the conventional stubs. Only the widths of the SMGC stubs are modified to show the significance of SMGC stubs.

Figure 4.22: S parameters of the two-way PD with conventional and SMGC stub. (a) S_{11} and (b) S_{21} .

Fig. 4.22(a) shows that the operating frequency is at 1.8 GHz, and another resonance occurs at 4.5 GHz for PD with the conventional stub. However, the same PD with SMGC based stub has no such resonance at 4.5 GHz. In Fig. 4.22(b), it shows that the

transmission zeros of the conventional stub at 5.5 GHz and 7.4 GHz is shifted to 4.5 GHz and 6.4 GHz, respectively, when SMGC stubs are used. Hence, SMGC based stubs can provide transmission zeros at frequencies lesser than conventional stubs, even if both kinds of stubs have the same length as shown in Table 5.4. This property of SMGC stubs can be used in reducing the size of the power divider.

4.4.7 Application of proposed four-way power divider for antenna array

Table 4.5: Dimensions of conventional 4-way Wilkinson PD

Parameters	w_1	w_2	w_3	l_1	l_2	l_3	s_1	L_1	W_1
Values	1	1.1	3.1	5	1.1	20	2	58.4	14.2
Parameters	w_{d1}	w_{d2}	w_{d3}	w_{d4}	w_{d5}	l_{d1}	l_{d2}	l_{d3}	l_{d4}
Values	0.1	3.1	0.6	0.3	75	44.5	31.5	23.5	49.5
Parameters	l_{d5}	l_{d6}	l_{d7}	l_{d8}	r_{d1}	r_{d2}	r_{d3}		
Values	18.4	26	30.3	43	100	13.5	9.9		

Table 4.6: Dimensions (mm) of antenna array

w_{d1}	w_{d2}	w_{d3}	w_{d4}	w_{d5}	l_{d1}	l_{d2}	l_{d3}
0.1	3.1	0.6	0.3	75	44.5	31.5	23.5
l_{d4}	l_{d5}	l_{d6}	l_{d7}	l_{d8}	r_{d1}	r_{d2}	r_{d3}
49.5	18.4	26	30.3	43	100	13.5	9.9

The proposed four-way PD is used to feed an antenna array with four horizontally polarized Vivaldi elements, as shown in Fig. 4.23(c). Another similar antenna arrays in Fig. 4.23(a) is fed by the four-way PD with tapered transmission line (TTL) whose dimensions are given in Table 4.6 and in Fig. 4.23(b) where the vivaldi array is fed by conventional Wilkinson-based PD whose dimensions are given in Table 4.5. The feeding

Figure 4.23: Antenna array with four horizontally polarized Vivaldi elements fed by (a) four-way PD with TTL, (b) four-way conventional Wilkinson PD, and (c) proposed four-way PD with harmonic suppression.

network is on the top layer, and the Vivaldi elements serving as its ground plane are on the bottom layer. The total antenna radius is given by r_{d1} . The exponential tapering and aperture of the Vivaldi elements are chosen such that the operating frequency is similar to that of the four-way PDs used. All the Vivaldi elements are simultaneously fed in-phase by the PDs.

The input reflections of the antenna array fed by three different feeding networks of the four-way PD, i.e., with TTL, with conventional Wilkinson, and the proposed four-way PD are shown in Fig. 4.24. It can be seen that when the proposed four-way PD is used to feed the antenna structure, the S_{11} -10 dB-bandwidth is from 1.45 to 2.1 GHz. The input reflection is well above -3 dB over the frequency range from 3 to 8 GHz, which indicates that the harmonics are suppressed and are not radiated. On the other hand, the four-way PD with TTL causes harmonics at frequencies above 3 GHz, where S_{11} is less than -10 dB. The four-way PD with the Wilkinson-based circuit also suffers from similar harmonics. These harmonics can be radiated by the antenna and could cause undesired

Figure 4.24: Input reflection of antenna array fed by four-way PD with TTL and proposed four-way PD.

Figure 4.25: Realized gain of antenna array fed by four-way PD with TTL and proposed four-way PD.

interference at frequencies outside the operating frequency range. The realized gains of the horizontally polarized Vivaldi antenna arrays fed by the three feeding networks are shown in Fig. 4.25. It can be inferred that the gains are almost similar in the operating frequency range of 1.6 to 2.5 GHz, but the gain for the antenna fed by four-way PD with TTL and conventional Wilkinson is higher than 0 dB at around 4.2 GHz. The antenna fed by the proposed four-way PD shows lower radiation at the harmonic frequency of 4.2 GHz. Fig. 4.26(a) and (b) shows the radiation patterns in the yz - and xy - planes for

Figure 4.26: Radiation patterns in yz - and xy - planes for the antenna array fed by the proposed four-way PD at (a) 1.8 GHz and (b) 2.3 GHz.

the antenna array fed by the proposed four-way PD at 1.8 GHz and 2.3 GHz respectively. The xy -plane patterns show that the four main beams are created by the four Vivaldi elements with almost the same gain while the yz -plane patterns show doughnut shape. Hence, the proposed four-way center-fed PD can provide wideband harmonic suppression that can be integrated with omnidirectional or multibeam planar antennas.

4.5 Conclusion

SMGCs can be used to form the shape profile of stubs or tuning patches, since they can provide additional degrees of freedom in tailoring the impedance bandwidth profile of the stubs. A center-fed four-way PD with wideband harmonic suppression has been designed

with SMGC stubs. Adding the SMGC-based stubs and coupled-line structures have been useful for enhancing the harmonic suppression and lowering the center frequency. The design and analysis of the proposed four-way PD have been discussed in detail, along with their simulated and measured S parameters. The proposed PD has been center-fed and can provide wideband harmonic suppression of up to $4f_o$, while providing a relatively wide -10 dB impedance bandwidth of 52%. The designed four-way PD can be integrated with horizontally polarized antenna arrays and can prevent the radiation at harmonic frequencies.

Chapter 5

Improved Designs of Antenna Arrays Using Modified Gielis Curves

5.1 Introduction

Antenna arrays or periodic structures comprise of antennas or unit cell elements placed at a regular inter-element distance, i.e., the period. The design of the unit element in an array decides the performance of the array. SMGCs can be widely used for such unit cell element design, which can improve the overall array performance and explore new design characteristics.

5.2 Three-Dimensional Bandstop Frequency Selective Structures

In this section, a Gielis shape profile is used for the full- and half-loop resonators in a 3-D FSS to obtain a reduction in the frequency of bandstop operation [196], when compared with circular loop resonators for a fixed unit cell size. The simulated S parameters of the circular and Gielis-shaped full- and half-loop resonators have been discussed in the next section along with the equivalent circuit of the proposed unit cell using [144].

Figure 5.1: Unit cells using (a) circular full loop, (b) Gielis shaped full loop, Half loop structures with (c) circular shape, and (d) Gielis shape. ($W = 7.5$, $L = 7$, $w = 1.1$, $d = 1.6$, $h = 2.6$, $r = 3.5$, $r_1 = 0.6$, $v_1 = 0.95$, $v_2 = 2.25$, $v_3 = 0.4$ (units : mm)).

5.2.1 Thin Bandstop FSS Based on Gielis-Shaped Loop Resonators

The 3-D FSS is designed on an FR4 board with $\epsilon_r = 4.3$ and thickness d . A circular-shaped loop resonator unit cell is shown in Fig. 5.1 (a) consisting of a metal trace width of w with a loop radius r on both sides of the unit cell. The unit cell size is fixed to be W and L . The unit cell simulation setup is similar to [144], and the incident waves are parallel to the yz -plane. A Gielis shaped full loop resonator is shown in Fig. 5.1 (b) on a similar substrate. The Gielis profile is chosen such that it could generate a loop with meander-like formations, which can increase the perimeter and can help in reducing the operating frequency. A modified form of Gielis curve is used for the metal trace obtained from (3.1) and (3.6), where the scaling constants are set as $S_x = S_y = 1.4$. In Fig. 5.3, it can be seen that the operating frequency of the unit cell with circular full-loop is at 10.5 GHz, whereas the unit cell with the Gielis-shaped full loop is reduced to 9.15 GHz, providing about 10% reduction in frequency. An FBW of about 4% is obtained for the Gielis-shaped full loop. The peak S_{11} is about -1.5 dB as the substrate used has $\tan\delta = 0.025$.

Figure 5.2: S parameter of the unit cells with full loop. Figure 5.3: S parameter of the unit cells with half loop.

To reduce the size of the loops, shorted vias are used at the point on the loop where the current is maximum [144]. Hence full-loop unit cells are made into circular half-loop and Gielis-shaped half-loop unit cells, as shown in Fig. 5.1 (c) and (d). From Fig. 5.2 and Fig. 5.3, it is observed that the operating frequency of circular half-loop is around 9.9 GHz, which is shifted to 8.8 GHz when half Gielis loops are used with shorted vias. The FBW of the stopband is about 6.4%. Furthermore, the equivalent circuit has been modeled for the half Gielis loop based on the method in [144]. The equivalent circuit consists of air and substrate transmission lines, capacitances, and inductance, as shown in the inset of Fig. 5.4. The equivalent circuit of the unit cell is manually tuned in ADS software and compared with the simulated S parameter results from CST software. The tuned values of components in equivalent circuit is given as $L_v = 57$ nH, $Z_a = 143 \Omega$, $Z_s = 88.1 \Omega$, $C_a = 0.196$ pF, and C_s is negligible. θ_a and θ_s are 180° at 19 GHz and 24.1 GHz respectively. Fig. 5.4 shows that the S parameters of the equivalent circuit are almost similar to that of the full-wave CST simulation for the Gielis half-loop. The 3-D

Figure 5.4: S parameters from equivalent circuit model and CST simulation for half Gielis loop.

Figure 5.5: 3-D structure of the proposed FSS using Gielis shaped half loop.

arrangement of the resultant FSS using Gielis half-loop resonator is shown in Fig. 5.5, where FSS layers of thickness d are placed apart by period h , and the incident wave propagates along z direction.

In this section, a three-dimensional bandstop frequency-selective structure (FSS) based on the Gielis-shaped loop resonator has been proposed. A Gielis shape profile for the full- and half- loop resonator has been used to provide a reduction in operating frequency by 10% when compared to a circular profile for a fixed unit cell size. S parameters of an equivalent circuit model for the half Gielis loop unit cell have also been discussed.

5.3 Periodic leaky-wave antenna

In this section, a periodic leaky-wave antenna with modified Gielis-shaped patch (MGP) elements [197] has been presented to enhance the operating bandwidth of the antenna. The effects of a conventional CRP and MGP elements on the extracted attenuation and phase constants are discussed in subsection 5.3.1. The simulation and measurement

Figure 5.6: Structure of (a) CRP unit cell, (b) MGP unit cell, and (c) Extracted attenuation and phase constants. ($w_1 = 1.4$ mm, $w_2 = 12$ mm, $w_3 = 11$ mm, $d_1 = 6.3$ mm, $d_2 = 12.3$ mm)

results, such as return loss, realized gain, and patterns are also compared in this section.

5.3.1 Periodic Leaky-Wave Antenna with MGP Elements

In this design, the number of unit cells, their period p , substrate width, and length are fixed for the conventional and proposed designs. Fig. 5.6(a) shows a CRP unit cell designed on an FR-4 substrate with permittivity $\epsilon_r = 4.4$ and thickness 1.6 mm, while Fig. 5.6(b) shows the MGP unit cell, which has a similar size as CRP unit cell. The SMGC curve profile G for the proposed unit cell is obtained from 3.1, 3.2, with the parameters $m = 4$, $n_1 = 4$, $n_2 = 1$ which determine the profile of the curve. The scaling constants are set as $S_{x1} = 6$, $S_{x1} = 10$, $S_y = 5$, to create the curve profile G as shown in Fig. 5.6(b).

The extracted attenuation and phase constants are normalized and are shown in Fig. 5.6(c) using the method as in [152]. It can be seen that the attenuation due to the MGP unit cell is lesser than that of CRP unit cell, especially in the frequency ranges of 7.5 GHz to 9 GHz. Therefore the MGP unit cell has a higher bandwidth than that of CRP unit cell. The phase constant of MGP unit cell is similar to that of CRP unit

Figure 5.7: Planar leaky-wave antennas with (a) CRP unit cells and (b) MGP unit cells. ($p = 24.6$ mm, $l_p = 10$ mm, $w_p = 3.1$ mm)

Figure 5.8: S_{11} of the planar leaky wave antennas.

cell and radiates in backward direction for frequency region 4.5 GHz to 6 GHz where $-1 \leq \beta/k_o \leq 0$ and forward direction for frequency region 6 GHz to 8.5 GHz where $0 \leq \beta/k_o \leq 1$.

Referring to Fig. 5.6, it can be seen that the attenuation due to MGP unit cell is lesser than that of CRP unit cell, especially in the frequency ranges of 7.5 GHz to 9 GHz. The attenuation of the CRP unit cell is about 0.25 higher than that of the MGP unit cell in the frequency range of 7.5 GHz to 9 GHz. Therefore the MGP unit cell has a higher bandwidth than that of CRP unit cell. From Fig. 5.10, the gain is observed to be much higher for the leaky-wave antenna with MGP unit cells than that of CRP unit cells in the frequency range of 7.5 GHz to 9 GHz.

The PLWA structure using CRP and MGP unit cells with the same substrate size is

Figure 5.9: S_{21} of the planar leaky wave antennas

Figure 5.10: Peak realized gain of the planar leaky wave antennas.

shown in Fig. 5.7 (a) and (b). A matching section of length l_p is used to match at the feeding point. A fabricated prototype of the MGP based antenna is shown in the inset of Fig. 5.8. It can be seen that the S_{11} of MGP based antenna has lower input reflection than CRP based antenna in the region of 7 GHz to 8.5 GHz. Hence the MGP based antenna has a relatively higher bandwidth than CRP based antenna. The measured S_{11} is close to simulated S_{11} but slightly shifted due to a small change in the ϵ_r of the fabricated prototype. Fig. 5.9 shows the transmission coefficient of the antennas with port 2 connected to 50Ω . It is observed that the S_{21} of MGP based antenna is higher than CRP based antenna over the region of 7 GHz to 8.5 GHz, indicating the leakage radiation of the antenna. The simulated and measured S_{21} are similar to each other. Fig. 5.10 shows that the gain bandwidth for MGP based antennas is higher than CRP

Figure 5.11: Normalized gain of the planar leaky wave antenna on xz -plane. ($\cdots$ simulated, $—$ measured)

based antenna. The gain of MGP based antenna is slightly lower at around 6.5 GHz as S_{21} is lower at that frequency. The simulated and measured normalized radiation patterns of the antenna on the xz -plane for various frequencies are almost the same and are shown in Fig. 5.11. The main lobe direction for frequencies 5 GHz, 6 GHz, 7 GHz and 8 GHz are at -37° , -4° , 11° , 28° respectively.

To explain the bandwidth improvement by using the modified shape, the surface current distribution along the length of the 4th patch element in the array and the input impedance of the leaky wave antenna is observed. The line along which the surface current is observed on the 4th patch element is shown in Fig. 5.12.

Figure 5.12: Line on 4th unit cell patch in the leakywave antenna along which surface current is observed. (a) CRP. (b) MGP.

Figure 5.13: Comparison between the CRP- and MGP-based leakywave antenna. (a) Surface current distribution along the length of the unit cell at 8 GHz. (b) Input impedance Z_{11} of the antenna.

The surface currents on the 4th patch element in both CRP and MGP leaky wave antennas at 8 GHz are shown in Fig. 5.13(a). It can be seen that the surface current is much higher along the unit cell length of the MGP patch at 8 GHz, which means that the MGP leaky-wave antenna radiates better at 8 GHz. Also, the real part of the input impedance of the MGP leaky-wave antenna at 8 to 8.5 GHz is lower than that of the CRP leaky-wave antenna, as shown in Fig. 5.13(b). Hence, increased surface current and lower real part of input impedance at around 8 GHz have favoured MGP leaky-wave antenna to extend its bandwidth of operation.

In this section, a periodic leaky-wave antenna with modified Gielis-shaped patch elements has been proposed. Modified Gielis curve profiles have been used for the patch elements for bandwidth enhancement of the PLWA. The antenna has an operating bandwidth from around 4 GHz to 8.5 GHz, and its characteristics such as return loss, insertion loss, realized gain, and radiation pattern have been simulated and compared with the measured results of the fabricated prototype.

Figure 5.14: Structures of (a) CPW fed slot antenna, (b) Antenna with square-shaped patch, (c) Square and G_1 file (d) Antenna with G_1 shaped patch. ($W_1 = 60, w_2 = 3, w_3 = 3.5, w_4 = 1.2, w_5 = 9, w_6 = 10, L_1 = 60, l_2 = 29.4, l_3 = 11.75, g_1 = 0.5, g_2 = 1.2, g_3 = 10.2, g_4 = 1, h_1 = 1.6, h_2 = 3.2$ (units : mm)).

5.4 Linearly Polarized Grid-Slotted Patch Antenna with Gielis-Shaped Patch

In subsection 5.4.1, a CPW fed slot is used to excite two metasurface layers placed in front and back of the slot to obtain bidirectional radiation. Also, a Gielis-shaped patch is used for the metasurface patch to reduce the lower cut-off frequency, which is observed through CMA [198]. A prototype with a Gielis-shaped patch is fabricated, and its measured results of reflection coefficient, realized gain, and patterns are compared and discussed in that subsection.

5.4.1 Bidirectional Linearly Polarized Grid-slotted Patch Antenna with Gielis-shaped Patch

The antenna consists of a CPW fed slot to excite the two metasurface layers placed in front and back for bidirectional radiation. A CPW fed slot is chosen instead of a conventional microstrip-line fed slot to reduce the number of printed circuit board layers. The proposed antenna is designed on an FR4 substrate with $\varepsilon_r = 4.1$ and $\tan\delta = 0.025$. Fig. 5.14 (a) shows the CPW fed slot antenna on the FR4 substrate with thickness h_1 that will excite the metasurface layers placed over the slot at height h_2 as shown in Fig. 5.14 (b). Square-shaped patches with width w_5 and gap g_2 are used to form the metasurface. As the shape and size of the patch can influence the operation over a frequency range, SMGC curve G_1 is used for the patches given by equation (3.1), (3.2). The values of S_{x1} , S_{x1} , S_y are shown in Fig. 5.14(c) and compared with square profile. The proposed antenna with G_1 shaped patch for metasurface is shown in Fig. 5.14 (d).

In Fig. 5.15, the antenna with G_1 shaped patches has a slightly lower cut-off frequency of 3.75 GHz as compared to the square-shaped patch. Hence lower cut-off frequency is reduced from 3.92 GHz by 4%. The fractional bandwidth for G_1 shaped patch is 43.75% from 3.75 GHz to 5.85 GHz. It can be noted that adding the metasurface patches improves the impedance matching for CPW fed slot antenna, which is a characteristic of the metasurface. The inset picture shows the fabricated prototype of the grid-slotted antenna with G_1 shaped patch. The measured S_{11} is quite close to the simulated result, and the deviation could be due to fabrication error or misalignment of FR4 layers.

To further observe the operating modes of the antenna, CMA is performed on the antenna. From [153], the total complex power associated with an arbitrary excitation $\vec{E}^{inc}$ of a structure at a given frequency can be written in terms of characteristic modes

Figure 5.15: S_{11} of CPW slot and grid-slotted patch antennas.

Figure 5.16: Modal significance of the grid-slotted patch antennas.

(CMs) of the structure as

$$S = P + jQ = \sum_{n=1}^{\infty} |\alpha_n|^2 (1 + j\lambda_n) \quad (5.1)$$

where

$$\alpha_n = \langle \bar{J}_n, \bar{E}^{inc} \rangle / (1 + j\lambda_n) \quad (5.2)$$

The symmetric product in the numerator of (5.2) is defined as the dot product of the two functions integrated over the surface of the conducting structure. Quantity $\bar{J}_n(\bar{r})$ is the current density of the n^{th} mode at location $\bar{r}$ on the structure and it is normalized such that the power radiated by the n^{th} mode is 1 W. The CM analysis involves ranking the modes based on their eigenvalue λ_n , with the smallest magnitude implying modal

dominance. The modal excitation coefficient (5.2) contains the term $1/(1 + \lambda_n)$, its magnitude

$$\Lambda(n) = 1/\sqrt{1 + \lambda_n} \quad (5.3)$$

can be considered to be a form of ‘significance measure’ for the n^{th} mode that determines the importance of a particular mode as far as its contribution to the use of the structure as a radiator is concerned. This modal excitation coefficient can also be referred to as the modal significance. The modal significance ranges from 0 to 1. The mode resonates and radiates the most efficiently when modal significance is 1.

To further observe the operating modes of the antenna, CMA is performed on the square and G_1 shaped metasurface patches using CST software to obtain the modal significance as shown in Fig. 5.16. The first mode $\mathbf{J}_1$ and third mode $\mathbf{J}_3$ for each of the metasurface patch shapes are shown. Mode $\mathbf{J}_2$ is not shown as it occurs at the same frequency of $\mathbf{J}_1$. It can be inferred that modes of the square-shaped patches are shifted towards lower frequencies when G_1 shaped patches are used, as they have slightly longer perimeter than the square-shaped patches, thereby providing a lower cut-off frequency.

Figure 5.17: Realized gain of grid-slotted patch antennas with square and Gielis-shaped patches.

From Fig. 5.17, it can be seen that the adding metasurface patches can dramatically increase the realized gain of the CPW fed slot antenna. The realized gain of the square-shaped and G_1 shaped metasurface patch are almost the same, as their aperture size

Figure 5.18: Normalized realized gain patterns of grid-slotted with G_1 shaped patch (a) xz -plane (b) yz -plane at 4 GHz.

is almost the same. The simulated and measured realized gain of the antenna with G_1 shaped metasurface patch are quite similar. The normalized gain patterns are shown in Fig. 5.18 (a) and (b) on the xz - and yz - plane respectively. It can be observed that the radiation patterns show bidirectional radiation in front (0°) and back (180°) directions, having almost symmetric patterns. The measured gain patterns are close to the simulated gain patterns.

The linearly polarized grid-slotted antenna with SMGC patch has reduced the cut-off frequency by 4.33% and has increased the bandwidth by 10.73%, as shown on Table 5.3.

Another advantage of using the SMGC patch for this linearly polarized grid-slotted patch antenna is that it can reduce the cross-polarization level of the antenna compared to square patches. The comparison of cross-polarization radiation of square and SMGC patch is shown in Fig. 5.19 for frequencies from 4.5 GHz to 6 GHz.

Figure 5.19: Cross polarized radiation of the grid-slotted patch antenna with square patch and SMGC patch cells for the frequencies 4.5 GHz, 4.75 GHz, 5 GHz, 5.25 GHz, 5.5 GHz, 5.75 GHz, 6 GHz.

It can be noted from Fig. 5.19, that the SMGC based patch can achieve at least 2 to 5 dB reduction in the cross-polarization compared to the square patch over the entire operating frequency range.

In this section, a bidirectional linearly polarized grid-slotted patch antenna with a Gielis-shaped patch has been presented. A CPW fed slot has been used to excite two metasurface layers with the Gielis-shaped patch for bi-directional radiation and to reduce the lower cut-off frequency. A prototype with the Gielis-shaped patch has been fabricated. Its simulated and measured results of antenna characteristics have been discussed.

5.5 Same-Sense Circularly Polarized Grid-Slotted Patch Antenna with Wide Axial Ratio Bandwidth

In this section, a same-sense circularly polarized grid-slotted patch antenna with a wider ARBW is presented. In beforehand, the ability of SMGCs to scale square or rectangular based shape profile is illustrated in subsection 5.5.1. The stripline feed excites the flipped Z-shaped slots pair which in turn excites the grid-slotted patches without any need for vias to facilitate easy fabrication. The antenna design, dispersion characteristics of the unit cell patches, relation between resonant frequencies, and the effects of the unit cell dimensions on the resonant frequencies and ARBW are discussed in subsections 5.5.2, 5.5.3 and 5.5.4. The simulated and measured antenna characteristics such as impedance bandwidth, ARBW, realized gain, and patterns are discussed in subsection 5.5.5.

5.5.1 SMGCs for square or rectangular shaped patches

Figure 5.20: Square and rectangular geometry generated from SMGC.

Usually, geometries that are generated by the formula will have difficulties or inconsistencies in generating geometric shapes with right-angled edges. But with SMGCs,

right-angled geometries such as square- or even rectangular-shaped geometries can be generated and scaled in the desired way. This ability of SMGCs to generate square or rectangle shaped edges can be used in the design of patches in metamaterials or electronic bandgap structures or in general any periodic structures. The different square or rectangular profiles that can be generated and scaled using SMGC curves are shown in Fig. 5.20. The rectangle shape for the patches obtained from SMGCs can be used in the grid-slotted patch antenna, as shown in Fig. 5.21(b).

5.5.2 Same-sense circularly polarized grid-slotted patch antenna design

The structure of the proposed grid-slotted patch antenna is shown in Fig. 5.21, and its dimensions are shown in Table 5.1. The antenna is built on FR-4 substrate layers of permittivity $\varepsilon_r = 4.1$ and loss tangent $\tan(\delta) = 0.025$ with thicknesses h_1 and h_2 . Fig. 5.21(a) shows the preliminary structure of antenna where two Z-shaped slots of widths w_{s1} , w_{s2} are flipped on the two ground planes to form a flipped Z-shaped slot pair and are excited by a stripline feed. The proposed antenna is shown in Fig. 5.21(b) where the unit cell patches of width w and length l are placed on the superstrates in the front and back directions to create a bidirectional right-hand circularly polarized (RHCP) radiation of same-sense with wide ARBW.

Table 5.1: Dimensions of the Proposed Same-Sense Circularly Polarized Grid-Slotted Patch Antenna (Unit: mm)

w_g	l_g	w_{f1}	w_{f2}	w_{s1}	w_{s2}	l_{s1}	l_{s2}
40	45	1.4	3.2	1.5	1.4	17.3	14.75
h_1	h_2	w	l	g_x	g_y	r_s	ε_r
1.6	3.2	8.4	8.8	0.6	0.8	5	4.1

The Z-shaped slot creates magnetic currents that help in providing orthogonal feeding with equal amplitudes and also a 90° phase difference, which are required to create a CP

Figure 5.21: Structures of Z-shaped slots pair fed by strip line (a) without grid-slotted patch and (b) with grid-slotted patch.

radiation [179]. However, the design in [179] is unidirectional and requires metallic vias in the unit cell patches along with some parasitic patches to improve the CP radiation. Unlike the previous work, the proposed design is not restricted to equal gaps and equal patch dimensions, i.e., $g_x \neq g_y$, $w \neq l$, which can improve the ARBW.

Figure 5.22: Simulated results of Z-shaped slots fed by stripline without the grid-slotted patches (a) S_{11} and realized gain and (b) Axial ratio and phase difference.

Figure 5.23: (a) Paths (A-B-C) along x -direction and (D-E-F) along y -direction where the E-fields are observed along the center of the slot in the ground plane. (b) Simulated E-field components E_x and E_y normalized by the maximum magnitude of E_y along the path A-B-C, and (c) Simulated E-field components E_x and E_y normalized by the maximum magnitude of E_x along the path D-E-F of the slot at 4.5 GHz for the antenna in Fig. 5.21(b) at various frequencies.

To understand the resonance and radiation characteristics of the flipped Z-shaped slot pair alone, the antennas without the grid-slotted patch are simulated, as shown in Fig. 5.22. The S_{11} and the realized gains in the front and back directions along the boresight are shown for the antenna without the grid-slotted patch in Fig. 5.22(a). The two cases ($\theta=0^\circ$ and $\theta=180^\circ$) are the boresight directions of the antenna in front and back direction. Since the feeding microstrip is in the center layer of the antenna, and the Z-shaped slot layers are placed in an inverted mirror arrangement, the radiation pattern in front and back directions are the same. Hence, the plots for two cases ($\theta=0^\circ$ and $\theta=180^\circ$) in simulation are identical. It can be seen that the first resonance occurs at 3.95 GHz while the structure radiates at the second resonance, which is around 6 GHz with a gain of 4.5 dBi. The total length ($l_{s1} + 2l_{s2}$) of the Z-shaped slot is around $1\lambda_g$ at 4 GHz where λ_g is the guided wavelength, and this causes resonance at 3.95 GHz. Also, the minimum AR magnitude is above 5 dB, and the phase difference ($\angle E_y - \angle E_x$) is not maintained around 90° , as shown in Fig. 5.22(b). The AR suggests that the flipped Z-shaped slot pair is not yet circularly polarized. Hence, the grid-slotted patch is placed as a superstrate to obtain the desired characteristics.

The slot lengths (l_{s1}, l_{s2}) and widths (w_{s1}, w_{s2}) can influence the ARBW of the antenna with the grid-slotted patch. Hence, the magnitudes of E-field components E_x and E_y along the center of the slot in the ground plane of the grid-slotted patch antenna are observed as shown in Fig. 5.23. The path along which the E-fields are observed on the ground plane is shown in Fig. 5.23(a). The path is aligned with the center of the slot and traversed along the slot length as indicated by a dark solid line. The magnitudes of the E-field components for the observed frequencies are normalized by the magnitude of either of the two components E_x or E_y , whichever is maximum at 4.5 GHz. The component E_y becomes dominant when the slot is along x direction as shown in Fig. 5.23(b). It can be seen that for frequencies 4 GHz to 5 GHz, the ratios of the magnitudes are less than 1.8 i.e., $|E_{y,B}|/|E_{y,A}| \leq 1.8$, $|E_{y,B}|/|E_{y,C}| \leq 1.8$, where $|E_{y,A}|$, $|E_{y,B}|$, $|E_{y,C}|$ are

Figure 5.24: (a) Dimensions of a unit cell (Top view). (b) Simulation setup of the unit cell for obtaining its dispersion characteristics.

the magnitudes of E_y component at the points A, B and C, respectively. Therefore, the component $|E_y|$ is relatively uniform along the x -directed slot (from point A to C) for the range 4 GHz to 5 GHz. For frequencies 3.8 GHz, 3.9 GHz, 5.1 GHz and 5.2 GHz, the ratios $|E_{y,B}|/|E_{y,A}| > 2$ or $|E_{y,B}|/|E_{y,C}| > 2$ which means greater variation of $|E_y|$ along the x directed slot at those frequencies. For example, $|E_y|$ at the points B and C along the x directed slot at frequency 5 GHz is given by $|E_{y,B}|=1.16$ and $|E_{y,C}|=0.67$. Among the $|E_y|$ components from 4 GHz to 5 GHz, the $|E_y|$ at 5 GHz has more deviation in magnitude, and hence a higher ratio $|E_{y,B}|/|E_{y,C}| = 1.73$ is observed for 5 GHz. Similarly, E_x becomes dominant when the slot is along y direction as shown in Fig. 5.23(c). The magnitude of E_x at the starting edge D of y -directed slot will be similar to that of E_x at the midpoint E of y -directed slot from 4 GHz to 5 GHz. Again for example, $|E_x|$ at the points E and F along the y directed slot at frequency 5 GHz is given by $|E_{x,E}|=1.07$ and $|E_{x,F}|=0.73$. Among the $|E_x|$ components from 4 GHz to 5 GHz, the $|E_x|$ at 5 GHz has more deviation in the magnitude and hence a higher ratio $|E_{x,E}|/|E_{x,F}| = 1.46$ is observed for 5 GHz. It can be inferred from Fig. 5.23(b) and (c) that the dominant E-field components are relatively uniform along the length of the slot in a given direction for the frequency range of 4 GHz to 5 GHz, which is a factor that can influence the ARBW of the proposed

antenna. In addition to the slot, the dimensions of a unit cell in the grid-slotted patch layer will also determine the ARBW of the antenna.

5.5.3 Dispersion characteristics of the unit cells

The grid-slotted patch layer consists of N_x and N_y unit cells along the x - and y - directions, respectively. The unit cells are placed with period p_x and p_y along the x - and y - directions, where $p_x = w + g_x$ and $p_y = l + g_y$. To understand the grid-slotted patch behaviour, theoretical resonant frequencies f_1 of TM_{10} and f_2 of antiphase TM_{20} modes can be calculated from the extracted propagation constant $\beta p/\pi$ of a unit cell with period p based on a simulated two-port model [157,177]. In this work, the simulation setup has the unit cells bounded by periodic boundaries along x - and y -directions while having perfect electric conductors (PEC) in z -direction as shown in Fig. 5.24. The PEC boundary on the top is at least $h_u = \lambda_g/4$ at the lowest operating frequency 4 GHz. This simulation setup is then solved using Eigen solver in CST. The dispersion diagram from the setup can provide the frequencies at which the modes TM_{10} , TM_{20} (along x -axis) and TM_{01} , TM_{02} (along y -axis) exist, as observed from the intercepts in Fig. 5.25(a).

The extracted phase constants $\beta_x p_x/\pi$ and $\beta_y p_y/\pi$ are shown in Fig. 5.25(a) for unit cells with period p_x and p_y , respectively. Using (5.4)-(5.7), the intercepts made by $(1 - 2\beta_{eff,x,1}\Delta L_x/\pi)/4$ and $(1 - 2\beta_{eff,x,2}\Delta L_x/\pi)/2$ on $\beta_x p_x/\pi$ gives the first two predicted resonant frequencies $f_{1,x}^p$, $f_{2,x}^p$ (superscript ‘ p ’ denotes predicted) for p_x which can be theoretically expressed as [157]

$$\beta_{x,1} p_x N_x + 2 \beta_{eff,x,1} \Delta L_x = \pi \quad (5.4)$$

$$\beta_{x,2} p_x N_x/2 + 2 \beta_{eff,x,2} \Delta L_x = \pi \quad (5.5)$$

$$\beta_{eff,x,1} = 2\pi f_{1,x}^p \sqrt{\varepsilon_{eff,x}}/c \quad (5.6)$$

$$\beta_{eff,x,2} = 2\pi f_{2,x}^p \sqrt{\varepsilon_{eff,x}}/c \quad (5.7)$$

Figure 5.25: (a) Dispersion characteristics for unit cells along x - and y - directions with periodic boundaries. (b) Simulated S_{11} of the proposed antenna.

where $\beta_{x,1} p_x$ and $\beta_{x,2} p_x$ are the extracted phase constants at frequencies $f_{1,x}^p$ and $f_{2,x}^p$ respectively for period p_x from unit cell simulation. $\beta_{eff,x,1}$, $\beta_{eff,x,2}$ are the phase constants for the extended fringing region ΔL_x at $f_{1,x}^p$ and $f_{2,x}^p$. Similar equations can also be used for period p_y to obtain resonant frequencies $f_{1,y}^p$, $f_{2,y}^p$ using equations below.

$$\beta_{y,1} p_y N_y + 2 \beta_{eff,y,1} \Delta L_y = \pi \quad (5.8)$$

$$\beta_{y,2} p_y N_y / 2 + 2 \beta_{eff,y,2} \Delta L_y = \pi \quad (5.9)$$

$$\beta_{eff,y,1} = 2\pi f_{1,y}^p \sqrt{\epsilon_{reff,y}} / c \quad (5.10)$$

$$\beta_{eff,y,2} = 2\pi f_{2,y}^p \sqrt{\epsilon_{reff,y}} / c \quad (5.11)$$

The resonances $f_{1,x}^p$ and $f_{2,x}^p$ are due to TM_{10} and TM_{20} , while $f_{1,y}^p$ and $f_{2,y}^p$ are due to TM_{01} and TM_{02} , respectively as it can be seen from the intercepts in Fig. 5.25(a). It should be noted that the lines $(1 - 2\beta_{eff,x,1}\Delta L_x/\pi)/4$ and $(1 - 2\beta_{eff,y,1}\Delta L_y/\pi)/4$ have very similar values, hence they are simply noted together as $(1 - 2\beta_{eff}\Delta L/\pi)/4$ in the legend of Fig. 5.25(a). From the intercepts on the extracted phase constants $\beta_x p_x/\pi$ and $\beta_y p_y/\pi$, the predicted resonances are noted as $f_{1,y}^p = 4.55$ GHz, $f_{2,y}^p = 5.29$ GHz and $f_{1,x}^p = 4.6$ GHz, $f_{2,x}^p = 5.44$ GHz. It can be noted that the first resonances $f_{1,x}^p$ for p_x and $f_{1,y}^p$ for p_y are quite close as indicated by dashed green circle in Fig. 5.25(a), but their corresponding second resonances $f_{2,x}^p$, $f_{2,y}^p$ are appreciably different which is indicated by a dotted green circle. The simulated S_{11} for the antenna with grid-slotted patch is shown in Fig. 5.25(b). The resonances at $f_{1,slot}^{sim} = 3.85$ GHz and $f_{2,slot}^{sim} = 5.8$ GHz (superscript ‘sim’ denotes resonance from simulated S_{11} of the proposed antenna) are caused by the Z-shaped slots pair as discussed before. The resonances of the Z-shaped slots pair are slightly shifted to a lower frequency in the presence of the grid-slotted patch as the superstrate. The predicted first resonant frequencies $f_{1,x}^p$, $f_{1,y}^p$ are very close to the simulated resonant frequency $f_1^{sim} = 4.57$ GHz. The predicted second resonant frequency $f_{2,x}^p = 5.44$ GHz is quite close to the simulated resonance $f_2^{sim} = 5.45$ GHz. It is inferred that the TM_{01} mode along with TM_{10} mode are excited in the grid-slotted patch by the flipped Z-shaped slots pair to cause the resonance f_1^{sim} with circular polarization and the resonance f_2^{sim} is caused by TM_{20} mode with linear polarization.

The relation between the two resonant frequencies for periods p_x and p_y can be derived from equations (5.4)-(5.11) as

$$\frac{f_{2,x}^p}{f_{1,x}^p} \approx \frac{1 - B_{x,2} N_x/2}{1 - B_{x,1} N_x}; \quad \text{along } x \text{ direction} \quad (5.12)$$

$$\frac{f_{2,y}^p}{f_{1,y}^p} \approx \frac{1 - B_{y,2} N_y/2}{1 - B_{y,1} N_y}; \quad \text{along } y \text{ direction} \quad (5.13)$$

where $B_{x,i} = \beta_{x,i} p_x/\pi$ and $B_{y,i} = \beta_{y,i} p_y/\pi$ in which $i = 1, 2$. In this design, the number of cells along the x - and y -axis are $N_x = N_y = 4$. Considering (5.13), since $f_{2,y}^p$ is supposed

to be the 2nd resonant frequency which will be greater than $f_{1,y}^p$, it is observed that the relation $(2B_{y,1} - B_{y,2}) > 0$ should be met in order to make the ratio in (5.13) to be greater than 1. From extracted phase constants, it is noted that $B_{x,1} = 0.206$, $B_{x,2} = 0.399$, $B_{y,1} = 0.208$ and $B_{y,2} = 0.400$. For instance, substituting $f_{1,y}^p$, $B_{y,1}$ and $B_{y,2}$ in equation (5.13) gives an approximate estimate for the second resonance as 5.41 GHz along y direction which is close to the predicted $f_{2,y}^p = 5.29$ GHz.

5.5.4 Effects of unit cell dimensions on resonant frequencies and axial ratio bandwidth

Moreover, a parametric study is performed to observe the effects of unit cell dimensions such as length l , gap g_y and height h_2 of the grid-slotted patch on the first two resonant frequencies of the antenna, namely f_1^{sim} and f_2^{sim} (not considering the resonance at 3.85 GHz and 5.8 GHz caused by the Z-shaped slots pair) of the antenna as shown in Fig. 5.26. The parameters w , g_x are not varied since the dimension l_{s1} of the Z-shaped slot is fixed. Initially, p_x is fixed as $\lambda_g/4$ at 4.6 GHz and later tuned to obtain $w = 8.4$ mm and $g_x = 0.6$ mm while l/w ratio is varied as shown in Fig. 5.26(a). It can be observed that by increasing l for a given w and g_y , the resonant frequencies corresponding to modes TM_{10} , TM_{20} , TM_{01} , TM_{02} are decreasing. From Fig. 5.26(b), it is noted that by increasing the g_y for a fixed l , the frequencies at mode TM_{01} remains almost same while TM_{02} slightly decreases. Also, the frequencies at modes TM_{10} and TM_{20} are increasing. It is inferred that by increasing the height h_2 of the superstrate, the frequencies at modes TM_{10} and TM_{01} are uniformly decreasing as shown in Fig. 5.26(c). Similar behaviour can be seen for the modes TM_{20} and TM_{02} . Hence, the resonant frequencies can be greatly influenced by tuning l , g_y , and h_2 . Moreover, the ground plane size, strip-line feed length can also slightly affect the resonances in the S_{11} of the antenna.

The influence of the unit cell and Z-shaped slot dimensions on the variation of ARBW and the position of the two frequency points $f_{1,AR}^{min}$, $f_{2,AR}^{min}$ are shown in Fig. 5.27. The

Figure 5.26: Variation of resonant frequencies $f_{1,x}^p, f_{2,x}^p, f_{1,y}^p, f_{2,y}^p$ for the corresponding modes $TM_{10}, TM_{20}, TM_{01}, TM_{02}$ in the unit cell patch based on the ratio of (a) length l and width w ($w = 8.4$ mm), (b) gap g_y and length l ($l = 8.8$ mm), and (c) height h_2 and period p_y ($l = 8.8$ mm, $g_y = 0.8$ mm).

frequency points $f_{1,AR}^{min}, f_{2,AR}^{min}$ are the local minimas where the AR is below 3 dB as labeled in Fig. 5.28. It is observed from Fig. 5.27(a) that the ARBW is better over the range $0.93 < l/w < 1.08$ with the maximum bandwidth of 21.2% at $l/w = 1.024$. It can also be noted that this increase in ARBW over that range is due to the introduction of a second AR local minimum $f_{2,AR}^{min}$ aside from $f_{1,AR}^{min}$. Similarly from Fig. 5.27(b), it can be seen

Figure 5.27: Variation of AR bandwidth and frequency points (at which AR has a local minimum) with ratio of (a) l/w ($w = 8.4$ mm), (b) g_y/g_x ($g_x = 0.6$ mm), and (c) l_{s2}/l_{s1} ($l_{s1} = 17.3$ mm) for the proposed antenna.

that the ARBW is around 20% when the gap ratio g_y/g_x is in the range $1.3 < g_y/g_x < 1.75$ with the maximum at $g_y/g_x = 1.33$. The ARBW decreases when the g_y/g_x is increased from 1.5 to 2.5 because the $f_{2,AR}^{min}$ also decreases, thereby affecting the bandwidth. ARBW

can be maintained at 20% if the slot-length ratio is in the range of $0.8 < l_{s2}/l_{s1} < 0.92$ except at some points where the 3 dB ARBW drops. The maximum ARBW occurs at $l_{s2}/l_{s1} = 0.809$. The ARBW abruptly drops to almost 10% for the ratios $l_{s2}/l_{s1} = 0.82$ and $l_{s2}/l_{s1} = 0.866$, because at those dimensions the axial ratio at the center frequency of the ARBW slightly exceeds 3 dB. For example, when $l_{s2}/l_{s1} = 0.82$ the ARBW drops to 13% from 20% because the axial ratio at 4.7 GHz is 3.1 dB and therefore, the upper cut-off frequency for the ARBW is chosen as 4.7 GHz instead of 4.9 or 5 GHz. It can be observed that both $f_{1,AR}^{min}$ and $f_{2,AR}^{min}$ will decrease with increase in the l_{s2}/l_{s1} ratio from 0.81 to 0.96, hence maintaining the bandwidth. Fig. 5.28 shows the simulated AR of the proposed antenna in front and back directions with 3 dB bandwidth of 21% from 4.07 GHz to 5.02 GHz and the phase difference is maintained around 90° within $\pm 10^\circ$ deviation.

Table 5.2 shows some of the recent works done in same-sense CP antennas. In [170], planar design involving complementary dipoles placed in a back-to-back position has a very narrow ARBW of 2.76%. In [171], the dipoles arranged in back-to-back configuration can provide only a narrow ARBW of 3.44%. Even some configurations of slot-coupled back-to-back patches [172,173] can suffer from a narrow ARBW of less than 5%. In [174], the rotated patches in back-to-back configuration excited through an H-shaped aperture can provide a narrow ARBW of 7.67%. Some three-element slot designs with strip-coupling patches [175] have a lower ARBW of 7.12% and need complex structure. An asymmetric FSS used as a polarization converting surface in [176] for CP same-sense radiation also suffers from narrow ARBW of less than 2%, since FSS structures usually operate over a narrowband. The proposed work with the grid-slotted patch fed by flipped-Z shaped slots pair provides a wider ARBW of 21% in both directions with appreciable gain.

The proposed antenna with same-sense circular polarization (SSCP) radiates both in front and back boresight directions. The wide axial ratio bandwidth of this SSCP antenna

Figure 5.28: Simulated axial ratio and phase difference.

Table 5.2: Comparison with previous same-sense circularly polarized planar antennas

Ref.	ϵ_r	f_l, f_h	IBW ^a	$L \times W \times H(\lambda_g)^b$	Gain ^c	f_l, f_h	ARBW
[170]	2.65	5.72, 5.88	2.76	$0.62 \times 0.7 \times 0.04$	4	5.72, 5.88	2.76
[171]	2.2	5.7, 5.9	3.44	$2.04 \times 1.62 \times 0.03$	5	5.7, 5.9	3.44
[172](1)	2.5	2.44, 2.54	4	$0.97 \times 0.97 \times 0.06$	NG	2.46, 2.5	1.61
[172] (2)	2.55	2.35, 2.46	4.57	$0.67 \times 1.01 \times 0.04$	NG	2.34, 2.45	4.56
[173]	2.5	2.44, 2.54	4.02	$0.97 \times 0.97 \times 0.06$	4 to 5	2.46, 2.51	2.01
[174]	4.4	2, 2.6	26.08	$0.78 \times 0.78 \times 0.32$	5.6	2.38, 2.57	7.67
[175]	4.4	2.28, 2.64	14.63	$0.94 \times 0.62 \times 0.39$	~ 5	2.3, 2.47	7.12
[176]	4.5	1.2, 1.65	31.58	$0.93 \times 0.93 \times 0.07$	2	1.54, 1.57	1.92
This work	4.1	4.05, 5.97	38.32	$0.95 \times 0.85 \times 0.2$	2 to 3.5	4.07, 5.02	21

f_l : lower cut-off; f_h : higher cut-off; ^a – 10 dB Impedance bandwidth (IBW); ^b $\lambda_g \approx c/(f_{low}\sqrt{(\epsilon_r + 1)/2})$; L : length; W : width; H : height; ^c Realized Gain (dBic).

is due to the combined effect of Z-shaped slot pair and the grid-slotted patches. From Fig. 5.23, it can be seen that the dominant electric field components in the Z-shaped slot are quite uniform for frequencies 4 to 5 GHz. Hence, the Z-shaped slot can support the excitation of this frequency range. The predicted resonances are noted as $f_{TM_{01}} = 4.55$ GHz, $f_{TM_{02}} = 5.29$ GHz, $f_{TM_{10}} = 4.6$ GHz and $f_{TM_{20}} = 5.44$ GHz.

The current distributions on the grid-slotted patches are given in Fig. 5.29 for frequencies from 4.1 GHz to 5.4 GHz. The directions of the surface currents are shown for phases 0° , 45° , 90° , and 135° , where the black arrow represents the direction along which most of the surface current vectors are oriented. It can be inferred that the proposed antenna exhibits RHCP radiation based on the direction in which the surface current vec-

Figure 5.29: Surface current distribution of the CPSS antenna at frequencies (a) 4.1 GHz, (b) 4.2 GHz, (c) 4.3 GHz, (d) 4.4 GHz, (e) 4.5 GHz, (f) 4.6 GHz

tors rotate for different phases. It can be seen from Fig. 5.29(a)-(i), that the sequential rotation of the surface current vectors indicate that the radiation is circularly polarized

Figure 5.29: Surface current distribution of the CPSS antenna at frequencies (contd.) (g) 4.7 GHz, (h) 4.8 GHz, (i) 4.9 GHz, (j) 5 GHz, (k) 5.1 GHz

for frequencies 4.1 GHz to 4.9 GHz. Fig. 5.29(j)-(n) shows that there is no such sequential rotation of surface current vectors for frequencies 5 GHz to 5.4 GHz. This shows that frequencies 5 GHz to 5.4 GHz can only support linear polarization. Hence, the modes TM_{01} and TM_{10} support the circular polarization. The combination of Z-shaped slot pairs and grid-slotted patch arrangements help in achieving wide axial ratio bandwidth for same-sense circular polarization.

Figure 5.29: Surface current distribution of the CPSS antenna at frequencies (contd.) (l) 5.2 GHz, (m) 5.3 GHz, and (n) 5.4 GHz

Figure 5.30: Simulated and measured S_{11} results of fabricated prototype of the proposed antenna.

5.5.5 Simulation and Measurement Results

The fabricated prototype is shown in the inset picture of Fig. 5.30. Excluding the resonance at 3.85 GHz and 5.8 GHz due to the Z-shaped slots pair, the measured -10 dB

Figure 5.31: Normalized RHCP and LHCP gain patterns on the xz (H-) plane and yz (E-) plane at frequencies (a) 4.2 GHz, (b) 4.5 GHz and (c) 4.8 GHz respectively.

Figure 5.32: Simulated and measured results of realized RHCP gain and LHCP gain.

Figure 5.33: Simulated and measured results of Axial ratio.

impedance bandwidth is 38.32% from 4.05 GHz to 5.97 GHz. The resonances in the measured S_{11} are similar to that of the simulated S_{11} . The impedance matching is lower than -10 dB, and RHCP gain is appreciable within the highlighted yellow region. The highlighted region denotes the frequency range for which AR is less than 3 dB as can be seen from Fig. 5.30, 5.32 and 5.33.

The simulated and measured RHCP and LHCP normalized gain patterns on xz (H)- and yz (E)- planes are shown in Fig. 5.31(a), (b), and (c) at frequencies 4.2 GHz, 4.5 GHz, and 4.8 GHz, respectively. It can be seen that the RHCP is radiated bidirectionally at $\theta = 0^\circ$ and 180° on both H- and E- planes. The LHCP gain pattern is usually 10 dB below the RHCP gain in both front and back directions. From Fig. 5.32, the measured realized RHCP gain varies from 2 to 3.5 dB within the AR bandwidth in the front ($\theta = 0^\circ$, $\phi = 0^\circ$) and back ($\theta = 180^\circ$, $\phi = 0^\circ$) directions which agree closely with the simulation. Similarly, the left-hand circularly polarized (LHCP) gain is almost 10 to 12 dB lower than the RHCP for most of the operating frequency range, i.e. 4 GHz to 5 GHz. Slight differences between simulated and measured results of RHCP and LHCP gain pattern could be due to fabrication errors and minor misalignments in the experimental setup. The simulated radiation efficiency is noted to be in the range of 60% to 75% within the AR bandwidth, which is due to the usage of a lossy substrate, and the gain could be improved by using low-loss substrates. The simulated and measured AR are shown in Fig. 5.33. The simulated AR(<3dB) is from 4.07 GHz to 5.02 GHz with an ARBW of 21% in both front and back directions. The measured ARBW in both directions are around 4.13 GHz to 5 GHz, which are about 19%. The measured ARBW are wide and are relatively similar in front and back direction, which depends on the dimensions of the Z-shaped slots pair and the grid-slotted patch.

In this section, a same-sense circularly polarized grid-slotted patch antenna with wide axial ratio bandwidth has been proposed. The rectangular grid-slotted patch elements on top and bottom planes have been excited through flipped Z-shaped slots pair by a stripline

feed to generate same-sense circular polarization with wide axial ratio bandwidth. The dispersion characteristics of the unit cell in the grid-slotted patch have been analyzed, and a relation between the first and second resonant frequencies has been provided. The variations of the resonant frequencies, the axial ratio bandwidths, and local minimas with the design parameters have been discussed. The antenna has a 3 dB axial ratio bandwidth of about 21% from 4.07 GHz to 5.02 GHz with the realized gain ranging from 2 to 3.5 dBic.

The author would like to reiterate the novelties done in this work, though some of them may not be very significant.

1. Flipped Z-shaped slots pair are excited using a stripline feed to generate same-sense circular polarization which is a simple concept but not yet reported before in literature for same-sense CP antennas.
2. A relationship between the first and second resonant frequencies for periods p_x and p_y is given as

$$\frac{f_{2,x}^p}{f_{1,x}^p} = \frac{1 - B_{x,2}N_x/2}{1 - B_{x,1}N_x}; \quad \text{along } x \text{ direction}$$

$$\frac{f_{2,y}^p}{f_{1,y}^p} = \frac{1 - B_{y,2}N_y/2}{1 - B_{y,1}N_y}; \quad \text{along } y \text{ direction}$$

3. Unlike previous works, the proposed design is not restricted to equal gaps and equal patch dimensions i.e., $g_x \neq g_y$, $w \neq l$ in the grid-slotted patch layer.

5.6 Conclusion

The SMGC can also be used to generate shapes with right-angled edges such as square, rectangle, etc., which forms the shape of the patch elements while designing unit cells of metamaterial structures. A bidirectional linearly polarized grid-slotted patch antenna with SMGCs has been presented. A CPW fed slot has been used to excite two metasurface

layers with the Gielis-shaped patch for bi-directional radiation and to reduce the lower cut-off frequency. A prototype with the Gielis-shaped patch has been fabricated. A same-sense circularly polarized grid-slotted patch antenna with wide axial ratio bandwidth has been designed. The designed CP antenna has a 3 dB axial ratio bandwidth of about 21% from 4.07 GHz to 5.02 GHz with the realized gain ranging from 2 to 3.5 dBic.

SMGC curves will be more advantageous for wideband antennas. It is difficult to give the variables of SMGC in terms of wavelength, but the parameters of the SMGCs can be chosen such that the perimeter of the curve can be expressed in terms of wavelengths given in Table 5.3.

The author would like to highlight the significant improvement in some of the works, as shown in Table 5.3. Most of the works have at least 10% improvement in some characteristics of the antenna when SMGC based structures are used.

Table 5.3: Improvements in antenna performance by using SMGC ($\varepsilon_r = 4.3$)

Structure	Type	Figure	f_l (GHz)	f_h (GHz)	FBW (-10dB)	$L(\lambda_g)^b$	$W(\lambda_g)$	$P(\lambda_g)$	$l(\lambda_g)$	$w(\lambda_g)$	f_l % $\downarrow$	FBW % $\uparrow$	S_{11} /Gain/ Radiation improvement	Improvement Remarks
Printed monopole with reflector	conv.	3.4	2.14	14	146.9	0.65	0.82	1.02	0.32	0.32	6.54	2.06	-	Slight improvement in FBW and reduction in f_l
	SMGC	3.4	2	14	150	0.6	0.77	1.03	0.3	0.3				
Bidirectional CPW-fed slot antenna	conv.	3.8	2.2	10	127.86	1.21	1.21	s=2.16 p=0.94	s=0.69 p=0.3	s=0.69 p=0.3	9.09	4.27	-	improvement in FBW and reduction in f_l
	SMGC	3.4	2	10	133.33	1.1	1.1	s=2.04 p=0.87	s=0.65 p=0.28	s=0.66 p=0.26				
Bidirectional microstrip-fed monopole antenna	conv.	3.14	2.66	9	108.74	0.77	0.64	1.52	0.43	0.42	6.01	3.95	-	Around 5 dB reduction in return loss at lower frequency range from 2.5 GHz to 4 GHz
	SMGC	3.4	2	10	133.33	1.1	1.1	s=2.04 p=0.87	s=0.65 p=0.28	s=0.66 p=0.26				
Bidirectional antipodal bowtie antenna	conv.	3.20	3	10	107.69	0.78	0.8	1.03	0.37	0.28	-	-	Gain: 0.5-1dB at 3-5GHz. 0.5-2dB at 7-10GHz. S_{11} : At least 5dB reduction in return loss at 3 -6GHz	Improvement in Gain up to 2dB Improvement in matching at the input port
	SMGC	3.20	3	10	107.69	0.78	0.8	1.02	0.33	0.29				
Bidirectional bowtie slot antennas	conv. (straight feed)	3.26	3.18	3.8	17.76 ^a	2	1.18	2.33	0.7	0.72	41.82	348.49	10-20dB reduction in return loss in operating BW	Improvement in matching and FBW and HPBW
	conv. (taper feed)	3.26	2	3.84	63 ^a	1.25	0.74	1.46	0.44	0.46	7.5	26.44	5-9dB reduction in return loss in operating BW. 5°-10° increase in HPBW	
	SMGC	3.26	1.85	4.3	79.67 ^a	1.16	0.69	1.37	0.42	0.48				

Table 5.3: Improvements in antenna performance by using SMGC ($\varepsilon_r = 4.3$) (Contd.)

Structure	Type	Figure	f_l (GHz)	f_h (GHz)	FBW (-10dB)	$L(\lambda_g)$	$W(\lambda_g)$	$P(\lambda_g)$	$l(\lambda_g)$	$w(\lambda_g)$	f_l % $\downarrow$	FBW % $\uparrow$	S_{11} /Gain/ Radiation improvement	Improvement Remarks
Frequency selective surface	conv. full loop	5.2	10	11.8	16.51 ^c	0.51	0.48	1.52	0.48	0.48	11.5	-	-	Reduce operating frequency
	SMGC full loop	5.2	8.85	9.56	7.71 ^c	0.45	0.42	1.76	0.42	0.42				
	conv. half loop	5.3	9.3	10.4	11.16 ^c	0.48	0.22	0.7	0.44	0.22	10.75	1.75		
	SMGC half loop	5.3	8.3	9.3	11.36 ^c	0.43	0.2	0.82	0.4	0.2				
Periodic leaky-wave antenna	conv.	5.8	4	7.5	60.86 ^a	5.99	0.95	1.27	0.33	0.3	-	18.28	Gain: 4-6dB increased at 7.5-8.5GHz	FBW increased by 1GHz at higher frequency end
	SMGC	5.8	4	8.5	72 ^a	5.99	0.95	1.13	0.38	0.3				
Linear grid slotted patch	conv.	5.13	3.92	5.85	39.5	1.62	1.62	0.97	0.24	0.24	4.33	10.73	Xpol decreased by 2-5dB in the operating frequency range	slight improvement in FBW.
	SMGC	5.13	3.75	5.85	43.75	1.55	1.55	0.92	0.23	0.23				Improvement in Xpol reduction.

^a -15 dB Bandwidth of S_{11} ; ^b guided wavelength $\lambda_g \approx c/(f_{low}\sqrt{(\varepsilon_r + 1)/2})$; ^c -6 dB S_{21} bandwidth; f_l : lower cut-off frequency; f_h : higher cut-off frequency; L, W : length and width of substrate; P : perimeter of geometry; l, w : length and width of geometry; $f_l \downarrow$: %decrease in f_l of SMGC compared to conv. geometry; FBW $\uparrow$: %increase in FBW of SMGC compared to conv. geometry; (green font) to highlight improvements; 's'-slot; 'p'-patch.

Table 5.4: Improvements in two-way power divider example using SMGC stubs

Structure	Type	Figure	f_l (GHz)	f_h (GHz)	f_c (GHz)	1 st Harmonics observed	1 st TZ	2 nd TZ	1 st TZ % ↓	2 nd TZ % ↓	Improvement Remarks
Two-way power divider	conv. [137]	22	1	2.2	1.85	4.6	5.5	7.2	16.9	12.5	SMGC stubs can have TZ at lower frequency as compared to conventional stubs
	SMGC	22	1	2	1.7	8.7	4.57	6.3			

f_l : lower cut-off frequency; f_h : higher cut-off frequency; f_c : center frequency; TZ: Transmission zero frequency; TZ↓: % reduction in Transmission zero frequency.

Chapter 6

Conclusions and Future Work

6.1 Conclusion

This thesis focuses on the design of planar structures such as antennas, power dividers, and antenna arrays using SMGCs. Based on the concept that the shape profile of the conductor patches or slots will influence the surface current distribution on the conductor patch, the SMGCs are used in the design of planar structures. The following are the conclusions drawn from this thesis.

An extensive literature review has been done on planar antennas, power dividers, and antenna arrays. The application of SMGCs in various antennas and power divider circuits have been explored and implemented. The improvements in antenna or power divider or antenna array characteristics are analyzed.

The SMGCs have been used to improve the impedance bandwidth characteristics or reduce the lower cut-off frequency of antennas or power dividers. The SMGCs can provide additional degrees of freedom while designing a structure. A patch with planar reflector and parasitic strips are designed using SMGCs to enhance the directivity. The -10 dB impedance bandwidth of the designed antenna is from 2 GHz to around 10 GHz, with the peak gain around 4 dB or above for most of the frequency range. Then, SMGCs

are applied to the slot and the tuning stub patch to provide additional surface current paths along the edges, which controls the impedance matching. The -10 dB impedance bandwidth of the designed antenna is from 2.5 GHz to around 9 GHz with a gain around 5 dB or above for most of the bandwidth. Another design involves the planar mushroom-shaped wideband monopole antenna using SMGCs, and it is shown that the SMGCs can conveniently design the complex shape profiles of the monopole. The -10 dB impedance bandwidth of the antenna is from 2.5 GHz to 9 GHz, and the peak gain is about 3 dB to 4.5 dB. Also, the antipodal bowtie patch fed by the microstrip to stripline transition utilizes a modified form of Gielis curve to generate complex shapes readily, which improve surface current and impedance matching. The -10 dB impedance bandwidth of the designed antenna is from 3 GHz to around 11 GHz with peak realized gain around 3 dB to 4 dB over most of the bandwidth. The design of wideband bowtie slot antenna (BSA) using sectorially modified Gielis curves (SMGCs) can greatly improve the impedance bandwidth while maintaining the antenna size. Two prototypes of BSAs are built on FR4 and RO5880 substrates providing fractional bandwidths of about 90.3% (1.7 GHz to 4.5 GHz) and 71.4% (2.7 GHz to 5.7 GHz), respectively.

A center-fed four-way power divider with wideband harmonic suppression is achieved using SMGCs to form the tapered open and short stubs along with coupled-line structures. SMGCs provide easy modifications in the tapering profile and additional degrees of freedom (DOF) while designing. The proposed four-way power divider is center-fed and can provide wideband harmonic suppression of up to $4f_o$, while providing a relatively wide -10 dB impedance bandwidth of 52%. An expression for the ABCD matrix of the stub-loaded coupled line structure with different even and odd mode electrical lengths ($\theta_e \neq \theta_o$) is presented.

A three-dimensional bandstop frequency-selective structure (FSS) based on SMGC-shaped loop resonator can provide a reduction of 10% in operating frequency when compared to a circular loop resonator having a fixed unit cell size. A bidirectional linearly

polarized grid-slotted patch antenna with SMGC-shaped patch can reduce the lower cut-off frequency by 4%. The rectangular grid-slotted patch elements from SMGCs on the top and the bottom planes are excited through flipped Z-shaped slots pair using a stripline feed to generate same-sense circular polarization with better axial ratio bandwidth. The dispersion characteristics of the unit cell in the grid-slotted patches have been discussed. An approximate relation between the first and second resonance frequencies and the axial ratio bandwidth variations have been discussed. The antenna has a 3 dB axial ratio bandwidth of about 21% from 4.07 GHz to 5.02 GHz with realized gain ranging from 2 to 3.5 dBic.

Hence, SMGCs can be used to modify the geometric parameters in the design of antenna, power dividers, and antenna arrays in order to improve their performance in terms of impedance bandwidth, impedance matching, gain, etc.

6.2 Future Work

6.2.1 Sectorially Modified Gielis curves in wiregrid array antennas

The usage of SMGCs can be extended to other planar antennas such as microstrip wire-grid array antennas. The SMGCs can be used for shaping the profile of the vertical radiating elements in the wire-grid array antennas to improve the matching. The horizontal elements only act as transmission lines and will not contribute to radiation. Instead of using conventional geometries like ellipse in the vertical radiators of the wire-grid array, some additional geometries from SMGCs can be explored to have a better matching.

Some existing designs in wire-grid array antennas use elliptical-shaped vertical radiating elements, as shown in Fig. 6.1. In [199], non-uniform elliptically shaped radiation elements along with a linearly tapered ground plane are designed to enhance the band-

Figure 6.1: Wiregrid antenna with elliptical shaped vertical patch radiators from (a) [199] and (b) [200].

Figure 6.2: Wiregrid antenna with astroid-shaped radiating elements [201].

width of the wire-grid array, as shown in Fig. 6.1(a). A similar type of wire-grid antenna is designed with uniform elliptically shaped radiation elements [200], but the ground plane is not tapered with respect to the substrate as shown in Fig. 6.1(b) to increase the gain and impedance bandwidth. Another wire-grid array consisting of 33 radiating elements with amplitude tapering techniques for automotive UWB radar sensors [201] has been shown in Fig. 6.2.

Since the shape of wire-grid's vertical radiation elements greatly influences the band-

width and gain, SMGCs can be used for these radiating patches to enhance the antenna performance. Using ellipse-shapes can only offer limited possibilities of patch geometry and dimensions, but SMGCs can offer a lot more geometries, dimensions, and better control over the design.

6.2.2 Extending Gielis curves to 3D SMGCs

Figure 6.3: 3D plot of Gielis curves [202].

The modified Gielis curves discussed in this thesis is based on the 2D formula. It is possible to extend the formula to 3, 4, or n dimensions, by means of the spherical product of Gielis formulas. For example, the 3D parametric surface is obtained by multiplying two Gielis formulas, say r_1 and r_2 . The coordinates are defined by the relations [203]

$$\begin{aligned}
 x &= r_1(\theta) \cos \theta \cdot r_2(\phi) \cos \phi \\
 y &= r_1(\theta) \sin \theta \cdot r_2(\phi) \cos \phi \\
 z &= r_2(\phi) \sin \phi.
 \end{aligned} \tag{6.1}$$

where ϕ (latitude) varies between $-\pi/2$ and $\pi/2$ and θ (longitude) between $-\pi$ and π .

A few examples of the possible 3D Gielis shapes for a given m, n_1, n_2, n_3 are shown in Fig. 6.3. Using (6.1), a 3D Gielis based shape can be used to form the shape of dielectric resonators based on the application.

3D Gielis in lens antennas

Figure 6.4: 3D Gielis shapes for lens antennas. (a) Lens fed by waveguide [203] and (b) Lens fed by a patch antenna array [204].

The 3D Gielis shapes have been used in some lens antennas, where a dielectric in the shape of a 3D Gielis is excited by waveguide [203] or placed as a superstrate over feeding antenna structure [204] as shown in Fig. 6.4. Based on analyzing the multiple wave reflections occurring within the lens using physical optics, the effect of these reflections on the radiation properties of the antenna can be evaluated. Even though 3D Gielis shapes have already been used for such lens antennas, this 3D Gielis shape is still based on conventional Gielis (3.1), which has issues when scaling along independent x or y axis. An attempt to produce 3D Gielis curves using two SMGC curves can be given as follows

$$\begin{aligned} X_{SMGC} &= r_1(\theta) \cos \theta \cdot r_2(\phi) \cos \phi \cdot S_x(\theta, \phi) \\ Y_{SMGC} &= r_1(\theta) \sin \theta \cdot r_2(\phi) \cos \phi \cdot S_y(\theta, \phi) \\ Z_{SMGC} &= r_2(\phi) \sin \phi \cdot S_z(\phi), \end{aligned} \tag{6.2}$$

where S_x , S_y and S_z are the 3D scaling constants for a given θ and ϕ range. Based on the selection of scalar constant for a spherical sector, the 3D SMGC shape can be modified.

3D Gielis in pattern reconfigurability

Figure 6.5: (a) Geometry of spherical dielectric resonator antenna (Side view). (b) Top view of two microstrip patches, and (c) Lens fed by a patch antenna array.

A spherical dielectric resonator antenna is fed by a microstrip resonating structure [205, 206] as shown in Fig. 6.5. The dielectric sphere is then made to excite at a higher-order resonance mode in order to increase the directivity of the radiation. These modes in the dielectric resonator depend on the shape of the dielectric resonator. Hence, instead of just using a spherical dielectric resonator for pattern reconfigurability [206], 3D SMGC based shapes can be used for these dielectric resonators. Additional surface modifications in the 3D SMGC based dielectrics can be optimized to provide better control over the frequency of operation, main lobe direction, or gain.

Author's Publications

Journals

1. Vignesh Shanmugam Bhaskar, E. L. Tan, and L. K. H. Holden, "Design of wide-band bowtie slot antenna using sectorially modified Gielis curves," *IEEE Antennas Wireless Propag. Lett.*, vol. 17, no. 12, pp. 2237-2240, 2018.
2. Vignesh Shanmugam Bhaskar and E. L. Tan, "Center-Fed Four-Way Power Divider with Wideband Harmonic Suppression" (To be submitted.)
3. Vignesh Shanmugam Bhaskar and E. L. Tan, "Same-Sense Circularly Polarized Grid-Slotted Patch Antenna with Wide Axial Ratio Bandwidth" (submitted to *IEEE Transactions on Antennas and Propagation* - under revision)

Conferences

1. Vignesh Shanmugam Bhaskar, "Compact combined antenna with slit for monopolar input pulse," 2017 IEEE International Symposium on Antennas and Propagation & USNC/URSI Notional Radio Science Meeting, pp. 699-700, 2017.
2. Vignesh Shanmugam Bhaskar, E. L. Tan, K. H. H. Li, and M. S. Tse, "Super shaped CPW-fed Monopole Antenna With Parasitic Strips for Unidirectional Pattern," 2018 IEEE international Symposium on Antennas and Propagation & USNC/URSI Notional Radio Science Meeting, pp. 1931-1932, 2018.
3. Vignesh Shanmugam Bhaskar, E. L. Tan, K. H. H. Li and M. S. Tse, "Exponentially Tapered TEM Horn Antenna With Dielectric Support for Monopolar Input Pulses," 2018 IEEE International Symposium on Antennas and Propagation & USNC/URSI Notional Radio Science Meeting, pp. 1755-1756, 2018.
4. Vignesh Shanmugam Bhaskar, E. L. Tan, K. H. H. Li and M. S. Tse, "Broadband CPW-fed Slot Antenna Using Super formula," 2018 IEEE international Symposium on Antennas and Propagation & USNC/URSI National Radio Science Meeting, pp. 1277-1278, 2018.

5. Vignesh Shanmugam Bhaskar, E. L. Tan and K. H. H. Li, "1 to 4 Way Wideband Power Divider using Substrate integrated Waveguide and Modified Wilkinson Structures," 2018 Joint IEEE EMC & APEMC Symposium, pp. 554-557, 2018.
6. Vignesh Shanmugam Bhaskar, E. L. Tan and K. H. H. Li, "Conical surface mount short horns for circular polarized patch antennas," European Conference on Antenna Propagation, pp. 1-3, 2018.
7. Vignesh Shanmugam Bhaskar, E. L. Tan, K. H. H. Li and M. S. Tse, "Design of Planar Mushroom-shaped wideband Monopole Antennas Using Gielis Curves," 2018 IEEE Asia-Pacific Conference on Antennas and Propagation (APCAP), pp. 231-232, 2018.
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9. Vignesh Shanmugam Bhaskar, E. L. Tan, and K. H. H. Li, "Three-dimensional bandstop frequency-selective structures based on Gielis-shaped loop resonator," 2019 IEEE International Symposium on Antennas and Propagation & USNC/URSI Notional Radio Science Meeting, pp. 1191-1192, 2019.
10. Vignesh Shanmugam Bhaskar, E. L. Tan, and K. H. H. Li, "Periodic Leaky-Wave Antenna with Modified Gielis-Shaped Patch Elements," 2019 IEEE International Symposium on Antennas and Propagation & USNC/URSI Notional Radio Science Meeting, pp. 135-136, 2019.
11. Vignesh Shanmugam Bhaskar and E. L. Tan, "Bidirectional linearly polarized grid-slotted patch antenna with Gielis-shaped patch," *Asia-Pacific Microwave Conference*, pp. 1542-1544, 2020.

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