Design and analysis of integrated array antennas for millimeter-wave applications

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Design and Analysis of Integrated Array Antennas for Millimeter-Wave Applications

Zhang Bing

School of Electrical and Electronic Engineering

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It feels like yesterday when I was enrolled by the School of Electrical and Electronic Engineering, Nanyang Technological. I have had so many pleasant memories in such a beautiful and sentimental island. It is hard to forget when I was strolling on campus, enjoying the gorgeous view of sunset and the gentle touches of breeze. It is hard to forget when I was working alone late at night in the laboratory with McDonald and Starbucks as my company. It is hard to forget the tremendous joy when I received the hardcopy of my first publication. All the precious sweetness and bitterness are hard to forget.

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Summary

The increasing demand for high bit rate wireless communication prompted people resort to the millimeter-wave (mmWave) spectrum, in which the free licensed 60-GHz band is of great application prospect. A single chip radio becomes possible for the small wavelength of millimeter electromagnetic (EM) waves. The antenna plays an important role because its independent properties can affect the radio as a whole. Nowadays, antenna designs are shifting from the design of a conventional discrete device to antenna-on-chip (AoC) and antenna-in-package (AiP). The AoC usually has low radiation efficiency because of the substrate's high conductivity, while the AiP features good performance since it is an independently designed device from the integrated circuit (IC) and it also has a wide selection of antenna prototypes and substrate materials.

Various antennas can be utilized in mmWave AiPs, among which the grid array antenna (GAA) is a capable candidate because of its high gain, broad impedance bandwidth and simple structure. Materials of capabilities to realize multilayer structures are suitable for AiP applications like the liquid crystal polymer (LCP) and low temperature co-fired ceramic (LTCC). The latter is used as the antenna substrates and packaging materials in this thesis.

The thesis is devoted to provide a physical insight into the characteristics of mmWave GAAs, develop a synthesis method for mmWave GAA design, and provide solutions for mmWave AiP applications using GAA as the antenna prototype. There are four aspects of the work: analysis, synthesis, design and packaging.

An introduction is given on the background of 60 GHz radios, the development of research work on GAAs and the LTCC process. A review of fundamentals of planar array antennas follows the introduction. An extensive analysis of microstrip GAAs with different number of radiating elements on a variety of substrates is conducted. A set of closed-form formula is developed for the synthesis of GAA design. A particular design example verifies the effectiveness of the synthesis method. A linearly-polarized (LP) GAA for 60-GHz AiP application is designed and packaged, which can be easily converted into either single balanced or dual balanced antennas. A circularly-polarized (CP) GAA is also designed and packaged, in which sub grid arrays are sequentially rotated and excited with 90° phase increment to radiate CP electromagnetic (EM) waves.

Chapter 1 Introduction

1.1 Background and Motivation

1.1.1 Millimeter-Wave Radio and Antenna-in-Package

In the past two decades, the ever increasing demand on high bit rate wireless communication has gone beyond the existing wireless channel capacities. Under such motivation, great effort has been taken to the research and exploration of the millimeter-wave (mmWave) spectrum for band resources. Besides the unexploited huge bandwidth, the mmWave spectrum also features many other application potentials: enabling densely packed networks that cover from short to medium ranges; leveraging frequency reuse to its paroxysm while increasing the security of each link; integrating active and passive devices into a single chip radio; exploiting quasi-unlimited and unique electromagnetic signatures for detection, diagnostic or imaging [1].

Compared with lower frequency electromagnetic (EM) waves that can propagate several kilometers, penetrate through walls and benefit from reflection and diffraction properties, the free-space loss (FSL) of mmWave is quite high. The large attenuation for mmWave precludes its application for long-range communications, but provides a spatial isolation which is preferable for frequency reuse in an indoor environment, reduces co-channel interference and provides extra safety for secure short range point-to-point links. Within the mmWave spectrum, the oxygen absorbing factor first peaks at 60 GHz. Thus the FSL (in the order of 10-15 dB/km in the 7 GHz band) is more severe than other

frequencies that usually used for mobile communication, which enables it as a qualified band for the high bit rate wireless communication such as home video streaming, high speed local area network, synchronization of mobile phones with computers and instantaneous file transfer using wireless universal series bus (USB).

As shown in Fig. 1.1, in 2000 Japan first issued the 60-GHz regulation for unlicensed utilization 59-66 GHz [2]. In 2004, the United States allocated the unlicensed 7 GHz band from 57-64 GHz [3]. Canada's regulation for the 60-GHz band is harmonized with the United States [4]. Australia allocated 3.5 GHz band from 59.4-62.9 GHz for the unlicensed use [5]. In Europe, the recommended bandwidth is from 57-66 GHz [6]. In China, the allocated band is from 59-64 GHz [7]. The IEEE 802.15.3c standard also dedicates the 57-66 GHz band with four 2.16-GHz channels for unlicensed usage [8].

Fig. 1.1: The 60-GHz standards and band allocations.

Since the wavelength of 60-GHz EM waves is in the order of millimeter, a single chip radio which is an assembly of several microwave monolithic integrated circuits and antennas becomes possible. Reynolds demonstrated the first 60-GHz fully integrated receiver and transmitter chipset in a 0.13- μm silicon-germanium (SiGe) technology [9]. Razavi demonstrated a 60-GHz transceiver chip in a 0.13- μm complimentary metal-oxide-semiconductor (CMOS) in 2006 [10]. Mitomo built a 60-GHz receiver chip in a 90-nm CMOS in 2007 [11]. Building-block circuits in a 65-nm CMOS were designed and characterized by Varonen in 2008 [12]. Marcu integrated the 60-GHz transceiver with the baseband circuitry in a 90-nm CMOS in 2009 [13]. A 60-GHz on-off-keying (OOK) receiver with an on-chip antenna was designed in a 90-nm CMOS by Kang in 2010 [14]. Vecchi demonstrated a 60-GHz wideband receiver on 65-nm CMOS in 2011 [15].

The antenna is an important component in a radio because of its independent properties that affect the radio as a whole. Antenna designs for the 60-GHz highly integrated radio are shifting from the conventional discrete devices to antenna-on-chip (AoC) and antenna-in-package (AiP) solutions [16]. As shown in Fig. 1.2 (a) the AoC solution is to integrate antennas with other front-end circuits on the same chip in mainstream silicon technologies such as SiGe or CMOS. Since substrate characteristics and processes are limited, only a few antenna prototypes could be adopted. The heating effect brought about by the low resistivity of the substrate as well as the surface wave loss from substrate's high permittivity are usually accountable for the low antenna efficiency, which is no more than 20%. The AiP solution (shown in Fig. 1.2 (b)) combines antennas with a highly integrated radio die into a standard surface mounted chip-scale package. Interconnection can be realized by wire-bonding or flip-chip technique. In the AiP solution, antennas and integrated circuits (IC) are designed and

implemented separately through different processes, thus wide selections of antenna prototypes and substrates are available which results in $> 80\%$ antenna efficiency. Liquid crystal polymers (LCP) and low temperature co-fired ceramics (LTCC) are popular for AiP applications. LCPs feature thinner thickness ($25\ \mu\text{m}$), finer conductor width (a few micrometers) and lower lamination temperature (285°C), but the relative permittivity is quite limited ($\epsilon_r = 3 \pm 0.2$) plus the slightly large loss tangent ($\tan \delta = 0.002\text{-}0.004$). The large variety of permittivity ($\epsilon_r = 4 - 60$) of LTCCs is very attractive for 60-GHz AiP applications. The dielectric loading of the LTCC substrates can help effectively minimize the antenna size, and existing processes are fine enough to satisfy the requirement for 60-GHz antenna fabrications.

Fig. 1.2: Antenna solutions for mmWave radio: (a) AoC and (b) AiP.

Different kinds of antennas were adopted on a wide selection of substrates for 60-GHz applications as shown in Fig. 1.3. On LTCC substrates, they were usually seen as cavity

antennas [17], dipoles [18]-[20], microstrip patch antennas [21]-[30], slot antennas [31]-[41], Yagi antennas [42]-[44], parabolic antenna [45] and rod antenna [46]. Martínez-Vázquez designed a cavity antenna fed by substrate integrated waveguide (SIW) on DuPont 943 green tape [17]. Dipoles were widely adopted because of their simple structure. For example Chang designed a V-band array of two folded dipoles with an IC carrier [18]. Keusgen integrated a dipole array on hybrid substrates into a package [19]. Sun presented a folded dipole array in a thin cavity-down ball grid array chip-scale package for the highly integrated 60-GHz radio [20]. Moreover, designs of microstrip patch antennas, slot antennas and Yagi antennas were also preferred for the simple design. A typical design of microstrip patch array was from Lamminen, who designed a patch array on Ferro A6M LTCC with either T-junctions or Wilkinson power dividers as the feeding network. An embedded cavity was applied to improve the radiation efficiency [22]. Xu demonstrated a slot array of > 20 dBi gain on 20-layer green tapes of LTCC substrates. No considerations for packaging could be seen in Xu's design. The tremendous substrate thickness hindered its industrial application [38]. Wai proposed an ultra-compact LTCC package with a Yagi antenna integrated, however the antenna's gain was limited [44]. A high-gain parabolic antenna was designed by Sekit. The antenna took fully the advantage of LTCC's capability to realize multilayer structure, and then was integrated with a MMIC into a radio [45]. A multi-beam rod antenna was designed by Huang to overcome the multipath fading in wireless communications, in which the extruding rods made the whole package quite bulky [46]. When it came to early 2000s', Zhang and Sun in the Micro Radio Group of Nanyang Technological University found grid array antennas (GAA) as competitive candidates for mmWave applications with their high gain, wide impedance bandwidth (BW) and the elimination of the feeding network, and then brought it to the stage of 60-GHz AiP applications [47]-[54].

However, they also found that the current synchronization of GAAs deteriorated dramatically with the increasing number of grids without a solution.

* For substrate classifications: ‘PCB’ includes Duroid, alumina, FR4 and other substrates from different vendors that compatible with the printing circuit board process; ‘others’ includes silica, glass, quartz, ceramic, foam, polymer, resin and MEMS based antennas.

**For antenna classifications: ‘microstrip patch’ also includes microstrip patch array; ‘dipole & monopole’ also includes dipole or monopole array; ‘others’ includes lens, planar inverted-F antennas (PIFA), inverted-F antennas (IFA), cavities, horns and waveguide antennas

Fig. 1.3: Statistics of antenna prototypes for 60-GHz application.

1.1.2 History of Grid Array Antennas

The GAA was invented by Kraus in 1964 [55] (Fig. 1.4). It was first designed as a non-resonant antenna with a back angle-fire beam. The antenna is composed of several interconnected rectangular elements, referred to as grid, whose long sides work as transmission line while short sides work both as transmission line and radiating element with the same mechanism of half-wave dipole. The length of its short side equals half wavelength and the long side twice as long. When bending short sides of each grid outward of the ground, the radiation can be enhanced. Two equal-amplitude side lobes

appear when feeding the antenna from the center instead of feeding it at the edge. Two years later, Kraus [56] was granted a patent, in which three methods were used to elongate the length of short sides for enhanced radiation: bending short sides of each grid outward of the ground, adding pins to every joint of adjacent grids and bending the ground plane toward the antenna. Moreover it was found that the antenna features a scanning beam controlled by the input signal's frequency.

Fig. 1.4: The backward angle-fire array antenna from Kraus [55].

In 1974, a GAA-based non-resonant antenna named as chain antenna (Fig. 1.5) was proposed by Tiuri [57]. In this work, he discovered that if the characteristic impedance of the connecting lines between each grid is half of that grid, EM waves could propagate along the chain without reflection. In the same year, a patent was granted [58]. In 1981 Conti proposed the wire grid microstrip antenna (Fig. 1.6) for mono-pulse radar that adopted four resonant GAAs with tapered aperture to control the side-lobe level [59].

Fig. 1.5: The chain antenna from Tiuri [57].

Fig. 1.6: The wire-grid microstrip antenna from Conti [59].

In 1995 Nakano proposed a GAA that was fed at the center by a differential pair as shown in Fig. 1.7 (a) [60]. The GAA has two layers of radiating arrays placed one above another with 90° rotation, which enables the GAA with two orthogonally linearly-polarized (LP) radiations. In 1997, meandering technology was applied to elongate the

effective length of short sides of the grids for higher gain [61] (Fig. 1.7 (b)). The first GAA to radiate circularly-polarized (CP) EM waves was realized by imposing C-shaped elements (Fig. 1.7 (c)), functioning as circularizer, above the grid array to change the polarization from linear to circular [62]. Based on the CP GAA, cross-mesh array antennas were designed that featured dual polarizations [63]-[65] (Fig. 1.7 (d)). The method of moments (MoM) was also developed by Nakano to analyze his formerly designed GAAs [66]-[68].

Fig. 1.7: Nakano's GAAs: (a) center-fed GAA [60], (b) meandering GAA [61], (c) CP GAA with C-shaped elements [62] and (d) cross-mesh array antenna [63].

Fig. 1.8: Palmer's GAA [69].

Fig. 1.9: Chen's GAA: (a) microstrip GAA with a rear-space structure [70] and (b) genetic algorithm optimized microstrip GAA [71].

Fig. 1.10: Zhang and Sun's packaged GAA: (a) top view and (b) bottom view [47].

GAAs were modeled from the perspective of impedance by Palmer in 1997, meanwhile he came up with a 9-grid GAA of 15% bandwidth and 13 dBi gain as shown in Fig. 1.8 [69]. Chen proposed two GAAs in Fig. 1.9 [70], [71]. One is a 2.4 GHz microstrip GAA with a rear-space configuration (Fig. 1.9 (a)), the other is a 2.4 GHz microstrip GAA with elliptical radiating element and meandering feeding line (Fig. 1.9 (b)) which is optimized by the genetic algorithm. When it came to the early 2000's, Zhang and Sun in the Micro Radio Group of Nanyang Technological University began to revive a series of antennas for 60-GHz AoC or AiP applications, among which they discovered that GAAs with their distinctive characteristics feature capable candidates for 60-GHz AiP applications (Fig. 1.10) [37], [44], [72]-[74].

However, no more than 30 papers on GAAs could be found within IEEE Xplore until recently, indicating the lack of knowledge of this antenna in the academia. Former designs of GAAs were mostly dedicated to low frequency (< 20 GHz) usage. Though Zhang and Sun brought the GAA to the stage of 60-GHz applications, their works were mostly implementation oriented. Characteristics of GAAs like resonance, input impedance, radiation resistance, gain, radiation patterns et al. have not been studied either qualitatively or quantitatively to provide a physical insight. Moreover, the lack of a systematic synthesis method is also the bottleneck for GAAs' wide-spread applications.

1.1.3 Low Temperature Co-Fired Ceramics

In this thesis, LTCCs are used as antenna substrates and packaging materials. The origin of LTCCs could be found from the RCA Corporation in late 1950s, and the bases of nowadays process (green tape fabrication, via forming and multilayer lamination) were developed at this time [75]-[77]. Thereafter, IBM was taking the lead in the process, whose circuit boards for commercialized computers in early 1980s were good examples [78]-[80]. Since the multilayer circuit board was co-fired at a high temperature of 1600°C , it is named as high temperature co-fired ceramic (HTCC) to differentiate from the later developed LTCC (co-firing temperature $< 1000^{\circ}\text{C}$). The earliest commercialized LTCC products could be seen in early 1990s when a few manufacturers developed multilayer boards using the process [81]-[84]. Though the LTCC technology has existed for more than 20 years, great interest in its application to fabricate ceramic based components and modules arose only in the early 2000s.

LTCC tape system includes rolls of glass/ceramic dielectric tapes with shrinkage-matched metal pastes. Different from thick film technology in which the sequential printing, drying and firing of each layer are processed sequentially, the LTCC technology processes all the different layers in parallel, resulting in optimum yields and cost effectiveness for the independence of each process upon the previous. Each layer can be inspected before stacking, which allows for very high yields. After all the layers are processed, they are stacked, laminated and co-fired at temperatures between 800° – 900° to form a high density, fully integrated device [85].

The advantages of LTCCs enable them as good options for 60-GHz AiP applications. They have a wide variety of dielectric constant, ranging from 4.7 to 60 as shown in Table 1.1, which functions as effective dielectric loading to minimize dimensions of the

package. The loss tangent of LTCCs is quite low (in the order of 0.001) which ensures minimum power dissipation by the antenna substrate. Their capabilities to realize multilayer structure facilitate the antenna's integration with other devices into a package. The cost effectiveness makes industrial mass production possible. However, LTCCs also have their disadvantages like the low thermal conductivity and unpredictable co-fired shrinkage. With the improvement of the technology, the co-fired horizontal shrinkage now could be well controlled. Though the co-fired vertical shrinkage is still not predictable, it could be effectively controlled within an acceptable range from 5% to 10%.

Table 1.1: Material specifications of LTCCs from relevant companies.

LTCC suppliers	Products	Dielectric constant (ϵ)	Loss tangent ($\tan \delta$)
CeraTec	Ceramtape GC	7.3-8.5(@1MHz)	0.002 (@1MHz)
Dupont	943 Low Loss Green Tape	7.4(@40GHz)	0.002 (<40GHz)
	951 Low Loss Green Tape	7.8(@3GHz)	0.003 (<3GHz)
Electro-Science Lab	41020-70C	7-8	0.005 (@ 1 MHz)
Ferro	A6S Tape	5.7-6.1(from 1GHz to 100GHz)	<0.002(from 1GHz to 100GHz)
	A6M Tape	5.7-6.1(from 1GHz to 100GHz)	<0.002(from 1GHz to 100GHz)
Heraeus	CT 700	7.5-7.9 (@1kHz)	0.0021(@1kHz)
	CT 701	7.5-7.9 (@1MHz)	0.003(@1MHz)
	CT 702	7.5-7.9 (@1kHz)	0.003(@1kHz)
	CT 703	7.5-7.9 (@1kHz)	0.003(@1kHz)
	CT 765	65 (@1MHz)	<0.002(@1MHz)
	CT 2000	9.1 (@2.5GHz)	<0.002(@2.5GHz)
	CT 800	5.3 (@20GHz)	<0.008(@20GHz)
	AHT03-003	Test based (can work up to 100GHz)	Test based(can work up to 100GHz)
	AHT98-029	Test based (can work up to 100GHz)	Test based (can work up to 100GHz)
	HL 800	6.8 (@2.5GHz)	<0.005(@2.5GHz)
	HL 2000	7.3 (@2.5GHz)	<0.0026(@2.5GHz)
Kyocera	G55	5.7	0.00125 (@10 GHz)
	GL600	9.5	0.003 (@ 10 GHz)
MARUWA	HA-995	9.7	-
Matsushita Kotobuki	MKE-100	7.8	-
Murata	BAS	6.1	0.003 (@ 5GHz)
	CZG	25.0	0.0015 (@ 5 GHz)
	-	60.0	0.0015 (@ 5 GHz)
NEC glass	MLS-25M	4.7	0.003 (@ 2.4 GHz)
	MLS-1000	8.0	0.002 (@ 2.4 GHz)

NEC Vacuum glass	MLS-41	19.0	0.002 (@ 2.5 GHz)
	MLS-61	8.1	0.007 (@ 2.4 GHz)
	GCS78	7.8	<0.003 (@ 1 MHz)
	GCS71	7.1	<0.003 (@ 1 MHz)
	GCS60	6.0	<0.003 (@ 1 MHz)
NGK	GC-11	7.9	0.005 (@ 3 GHz)
Niko	NL-Ag II	7.8	<0.003 (@ 1 MHz)
	NL-Ag III	7.1	<0.003 (@ 1 MHz)
Sumitomo Metal Electronics Device	LFC	7.7	-

The fabrication process of a multilayer component using LTCCs is shown in Fig. 1.11 [86], [87]. First, green tapes are blanked and then punched in order to form via holes. Secondly, via holes are metalized to form electrical connections between layers. Next, electrical conductors and passive circuits are patterned onto each layer by screen printing or photo imaging. The final steps are layer lamination, firing and device dicing.

Fig.1.11: LTCC fabrication process.

1.2 Objectives

Objectives of the research presented in this thesis include four aspects: analysis, synthesis, design and packaging. The analysis is conducted by design experiments of GAAs with different number of grids on various substrates. Based on the analysis, a systematic synthesis method is proposed to help shortening the design circle. According to the proposed synthesis method, a versatile model of the GAA for 60-GHz AiP applications using the LTCC as antenna substrates and packaging materials is introduced. Deriving from the versatile model, a LP and CP GAAs are designed, packaged, fabricated and tested.

1.3 Major Contributions of the Thesis

The work of this thesis coincides with the rapid development of 60-GHz single chip radios, in which antennas are key components. In the scope of 60-GHz AiP analysis and design, contributions of the work can be concluded as follows.

The first contribution is on the analysis of GAA's characteristics. Former work mostly focused on the design and optimization of the GAA's structure. The understanding of the behavior of GAA's characteristics like the resonant frequency, impedance BW et al. was still insufficient. In this thesis, through a series of design experiments of GAAs with different number of grids on various substrates, GAA's characteristics at the 60-GHz band are thoroughly analyzed for the first time. The analysis also verifies the GAA's radiation mechanism by pattern synthesis for the first time.

The second contribution is the proposed synthesis method. For the lack of knowledge of GAA's characteristics, the design procedures of a GAA used to be trial-and-error based, which is time-consuming. A systematic synthesis method is proposed that can help to shorten the design circle effectively.

The third contribution is on design and packaging. A versatile GAA prototype adopting four sub grid arrays is proposed. Deriving from the prototype, a series of GAAs can be developed as the LP GAA, CP GAA, single balanced GAA and dual balanced GAAs. The LP and CP GAAs are converted into packages. The optimization of the transition between different signal traces are conducted from the perspective of parallel-plate mode distribution and transmission coefficient, which provides a much straightforward interpretation of the transition's function on the GAA's performance.

1.4 Organization of the Thesis

The thesis includes six chapters. Chapter 1 gives an introduction of the overall thesis, including the background, motivation and contributions of the work. In this chapter, the information about research on the 60-GHz band and single-chip radio is introduced. A literature survey reviews the history of GAAs. The material characteristics and fabrication process of LTCCs are also involved.

Chapter 2 reviews the fundamentals of planar array antennas, which lays the foundation for the analysis and synthesis of GAA in the following chapters. Characteristics like the array factor, beamwidth and directivity are evaluated.

Chapter 3 analyzes the performance of GAAs with different number of radiating elements on a wide selection of substrates. By distinguishing the GAA's behavior at non-

resonance, series resonance and parallel resonance, it confines the following discussion of GAAs at parallel resonance. After a comprehensive study of the resonant frequency, impedance BW, half-power beamwidth, gain, 3-dB gain BW and radiation patterns, a synthesis method is proposed. A design example is given to verify the synthesis.

In Chapter 4, a LP GAA for 60-GHz application is designed and packaged. Replacement of the single-feed large radiating array by the multiple-feed sub grid arrays results in better synchronized current and application flexibilities, by which either a LP, CP, single balanced or dual balanced GAAs can be realized. A parametric study demonstrates the effect of the grid's dimensions on GAA's performance. The LP GAA is packaged into an AiP module. The transition between different signal traces in the package is optimized in terms of parallel mode distribution and transmission coefficient.

Chapter 5 presents a 60-GHz CP GAA. By sequential rotating the sub grid arrays in the LP GAA and feed each with 90° phase increment, CP EM waves can be generated. The CP GAA is also packaged. Its performance is analyzed against the feeding phase and amplitude errors, showing its robust fault-tolerance from the feeding network.

Chapter 6 concludes the thesis and gives recommendations for future work.

1.5 Summary

This chapter gives an introduction of the whole thesis. It starts with the background and motivation under which the thesis is carried on. After confining the focus of the thesis on the design of AiPs for mmWave applications, the history of GAA is briefly reviewed. The LTCC materials and processes that are used to realize the proposed GAAs are

introduced. Objectives and major contributions of the thesis follow the background and motivation. Finally the organization of the thesis is presented.

Chapter 2 Fundamentals of Planar Array Antenna

In certain applications, desired antenna characteristics can be satisfied by a single radiating element. However, a single radiating element is not always capable enough to meet some technical requirements for high gain, narrow and/or steerable beams, pattern nulls in certain directions and controlled side lobe level. The 60-GHz band is targeted for short range wireless communications, in which the considerable atmospheric attenuation requires high gain of RF front-end antennas. This can be accomplished by forming an assembly of radiating elements in a geometrical configuration that is referred to as an array. For example, the grid array antenna (GAA) is one kind of planar array antenna that can be considered as an assembly of half-wave dipoles. By constructing an antenna array, high directivity and focused beam can be achieved to compensate the large atmospheric attenuation in a 60-GHz link.

The array is characterized by element type, geometry, and the required excitations. The elements are usually chosen, if possible, to be identical. The excited field of the array is determined by the vector addition of the fields generated by the individual element (neglecting mutual coupling). To achieve high directivity, the array is designed to ensure that the field generated by each element interferes constructively in the desired regions and destructively in the remaining space.

The elements of an array can be distributed spatially to form a linear, planar or volume array. A planar array consists of elements distributed on a plane. Planar arrays composed by uniform elements with fixed separations are the most commonly utilized array antennas. The GAA in this thesis could be categorized as a planar array of uniform half-

wave dipoles as radiating elements, and the separation between radiating element is fixed by the same dimensions of each grid. In this chapter, fundamentals of planar array antenna, like the array factor, beamwidth and directivity will be discussed as the foundation for the analysis and synthesis of GAAs in Chapter 3 [88].

2.1 Array Factor

To derive the array factor of a planar array, let us refer to Fig. 2.1. As shown in Fig. 2.1

(a), if M elements are initially placed along the x -axis, the array factor can be written as:

$$AF = \sum_{m=1}^M I_{m1} e^{j(m-1)(kd_x \sin\theta \cos\phi + \beta_x)} \quad (2.1)$$

where I_{m1} is the excitation coefficient of each element.

(a)

(b)

Fig. 2.1: Linear and planar array geometries: (a) linear array and (b) planar array [88].

The spacing and progressive phase shifts between the elements along the x -axis are presented by d_x and β_x respectively. As shown in Fig. 2.1 (b), if N such arrays are placed next to each other in the y -direction with a distance d_y and a progressive phase β_y , a rectangular array will be formed. The array factor of the planar array can be expressed as:

$$AF = \sum_{n=1}^N I_{1n} \left[\sum_{m=1}^M I_{m1} e^{j(m-1)(kd_x \sin\theta \cos\phi + \beta_x)} \right] e^{j(n-1)(kd_y \sin\theta \sin\phi + \beta_y)} \quad (2.2)$$

or

$$AF = S_{xm} S_{yn} \quad (2.3)$$

where

$$S_{xm} = \sum_{m=1}^M I_{m1} e^{j(m-1)(kd_x \sin\theta \cos\phi + \beta_x)} \quad (2.4)$$

$$S_{yn} = \sum_{n=1}^N I_{1n} e^{j(n-1)(kd_y \sin \theta \sin \phi + \beta_y)} \quad (2.5)$$

Equation (2.3) indicates that the pattern of a rectangular array is the product of the array factors of the arrays in the x - and y -directions.

If the amplitudes of excitation in the y -direction are proportional to those along the x , the amplitude of the (m, n) the element can be written as:

$$I_{mn} = I_{m1} I_{1n} \quad (2.6)$$

If in addition the amplitude excitation of the entire array is uniform ($I_{mn} = I_0$), (2.2) can be expressed as:

$$AF = I_0 \sum_{m=1}^M e^{j(m-1)(kd_x \sin \theta \cos \phi + \beta_x)} \sum_{n=1}^N e^{j(n-1)(kd_y \sin \theta \sin \phi + \beta_y)} \quad (2.7)$$

The normalized form of (2.7) can also be written as:

$$AF_n(\theta, \phi) = \left[\frac{1}{M} \frac{\sin\left(\frac{M}{2}\psi_x\right)}{\sin\left(\frac{\psi_x}{2}\right)} \right] \left[\frac{1}{N} \frac{\sin\left(\frac{N}{2}\psi_y\right)}{\sin\left(\frac{\psi_y}{2}\right)} \right] \quad (2.8)$$

where

$$\psi_x = kd_x \sin \theta \cos \phi + \beta_x \quad (2.9)$$

$$\psi_y = kd_y \sin \theta \sin \phi + \beta_y \quad (2.10)$$

2.2 Beamwidth

The task of finding the beamwidth of non-uniform amplitude planar array is quite formidable. Instead, a very simple procedure will be outlined which can be used to compute these parameters for large arrays whose maximum is not scanned too far off

broadside. The method [89] utilizes results of a uniform linear array and the beam broadening factor of the amplitude distribution.

Fig. 2.2: Half-power beamwidth for a conical main beam oriented toward $\theta = \theta_0$, $\varphi = \varphi_0$ [88].

The maximum of the conical main beam of the array is assumed to be directed toward θ_0 , φ_0 as shown in Fig. 2.2. To define a beamwidth, two planes are chosen. One is the elevation plane defined by the angle $\varphi = \varphi_0$ and the other is a plane that is perpendicular to it. The corresponding half-power beamwidth of each is designated, respectively, by Θ_h and Ψ_h .

For a large array, with its maximum near broadside, the elevation plane half-power beamwidth Θ_h is given approximately by [89]

$$\Theta_h = \sqrt{\frac{1}{\cos^2 \theta_0 (\Theta_{x0}^{-2} \cos^2 \phi_0 + \Theta_{y0}^{-2} \sin^2 \phi_0)}} \quad (2.11)$$

where Θ_{x0} presents the half-power beamwidth of a M -element broadside linear array. Similarly, Θ_{y0} presents the half-power beamwidth of a N -element broadside array.

The value of Θ_{x0} and Θ_{y0} can be obtained by using previous results. For a uniform distribution, for example, the values of Θ_{x0} and Θ_{y0} can be obtained by using, respectively, the lengths $(L_x + d_x)/\lambda$ and $(L_y + d_y)/\lambda$. For a Tschebyscheff distribution, the values of Θ_{x0} and Θ_{y0} are obtained by multiplying each uniform distribution value by the beam broadening factor

$$f = 1 + 0.636 \left\{ \frac{2}{R_0} \cosh \left[\sqrt{(\cosh^{-1} R_0)^2 - \pi^2} \right] \right\}^2 \quad (2.12)$$

The same concept can be used to obtain the beamwidth of other distributions as long as their corresponding beam broadening factors are available.

For a square array ($M = N$, $\Theta_{x0} = \Theta_{y0}$), (2.80) reduces to

$$\Theta_h = \Theta_{x0} \sec \theta_0 = \Theta_{y0} \sec \theta_0 \quad (2.13)$$

Equation (2.13) indicates that for $\theta_0 > 0$ the beamwidth increases proportionally to $\sec \theta_0 = 1/\cos \theta_0$. The broadening of the beamwidth by $\sec \theta_0$, as θ_0 increases, is consistent with the reduction by $\cos \theta_0$ of the projected area of the array in the pointing direction.

The half-power beamwidth Ψ_h , in the plane that is perpendicular to the elevation $\varphi = \varphi_0$, is given by [89]

$$\Psi_h = \sqrt{\frac{1}{\Theta_{x0}^{-2} \sin^2 \phi_0 + \Theta_{y0}^{-2} \cos^2 \phi_0}} \quad (2.14)$$

and it does not depend on θ_0 . For a square array, (2.14) reduces to

$$\Psi_h = \Theta_{x0} = \Theta_{y0} \quad (2.15)$$

The values of Θ_{x0} and Θ_{y0} are the same as in (2.11) and (2.13).

For a planar array, it is useful to define a beam solid angel Ω_A by

$$\Omega_A = \Theta_h \Psi_h \quad (2.16)$$

Using (2.11) and (2.14), (2.16) can be expressed as:

$$\Omega_A = \frac{\Theta_{x0} \Theta_{y0} \sec \theta_0}{\left(\sin^2 \phi_0 + \frac{\Theta_{y0}^2}{\Theta_{x0}^2} \cos^2 \phi_0 \right)^{1/2} \left(\sin^2 \phi_0 + \frac{\Theta_{x0}^2}{\Theta_{y0}^2} \cos^2 \phi_0 \right)^{1/2}} \quad (2.17)$$

2.3 Directivity

The directivity of the array factor $AF(\theta, \phi)$ whose major beam is pointing in the $\theta = \theta_0$ and $\phi = \phi_0$ direction, can be expressed as:

$$D_0 = \frac{4\pi [AF(\theta_0, \phi_0)] [AF(\theta_0, \phi_0)]^* \big|_{max}}{\int_0^{2\pi} \int_0^\pi [AF(\theta, \phi)] [AF(\theta, \phi)]^* \sin \theta d\theta d\phi} \quad (2.18)$$

A novel method has been introduced [90] for integrating the terms of the directivity expression for isotropic and conical patterns. As in the case of the beamwidth, the task of evaluating (2.18) for non-uniform amplitude distribution is formidable. Instead, a very simple procedure will be outlined to compute the directivity of a planar array using data from linear arrays. It should be pointed out that the directivity of an array with bidirectional characteristics (two-sided pattern in free space) would be half the directivity of the same array with unidirectional (one-sided pattern) elements (e.g., dipoles over ground plane).

For large planar arrays, which are nearly broadside, the directivity reduces to [89]

$$D_0 = \pi \cos \theta_0 D_x D_y \quad (2.19)$$

where D_x and D_y are the directivities of broadside linear arrays each, respectively, of length and number of elements L_x , M and L_y , N . The factor $\cos\theta_0$ accounts for the decrease of the values, D_x and D_y , can be obtained by using

$$D_0 = \frac{2R_0^2}{1 + (R_0^2 - 1)f \frac{\lambda}{(L + d)}} \quad (2.20)$$

with the approximate beam broadening factor f . For Techebyscheff arrays, D_x and D_y can be obtained using (2.12).

For most practical amplitude distributions, the directivity of (2.19) is related to the beam solid angle of the same array by

$$D_0 \cong \frac{\pi^2}{\Omega_A(\text{rads}^2)} \quad (2.21)$$

where Ω_A is expressed in square radians or square degrees.

2.4 Summary

In this chapter, fundamentals of planar array antennas are introduced. Characteristics like the array factor, beamwidth and directivity are discussed, which lays the theoretical foundation for analysis and synthesis in the following chapter.

Chapter 3 Analysis and Synthesis of Microstrip Grid Array Antenna

This chapter presents a comprehensive study of microstrip (MS) grid array antennas (GAA) on a variety of substrates at the 60-GHz band. Characteristics of MS GAAs like resonant frequency, impedance bandwidth (BW), gain etc. are analyzed against the substrate permittivity, thickness as well as the number of radiating elements. A synthesis method is proposed based on the analysis.

3.1 Revisit of the Mechanism of Microstrip Grid Array Antenna

A typical MS GAA is shown in Fig. 3.1. The grid array radiator is on the top surface of a substrate of relative permittivity ϵ_r , loss tangent δ , length a , width b , and thickness h . It is fed at a single point by a pin of the diameter p through an aperture of the diameter d on the ground plane on the bottom surface of the substrate. The grid array radiator consists of rectangular grids whose long sides of the length l and the width w_l and short sides of the length s and the width w_s . In the operation of the antenna, as pointed out by Kraus [62], proper choice of frequency, lengths l and s , and feed point permits the grid array structure as either a non-resonant or a resonant antenna. This chapter concentrates on the latter type that requires approximately $l = \lambda_g$ and $s = \lambda_g/2$, where λ_g is the guided wavelength at the center frequency of operation [35]. The radiation is essentially from the short sides with the long sides acting mainly as guiding or transmission lines.

The MS GAA is designed on a substrate of $\epsilon_r = 4.5$, $\tan \delta = 0.002$, $a = 25$ mm, $b = 25$ mm, and $h = 0.254$ mm with grid dimensions of $l = 2.84$ mm, $w_l = 0.1$ mm, $s = 1.42$ mm, and $w_s = 0.1$ mm as well as the feeding pin diameter $p = 0.125$ mm and the diameter of aperture on ground plane $d = 0.25$ mm. We simulate the performance with the three-dimensional finite-element full-wave solver Ansoft HFSS. Fig. 3.2 (a) shows the simulated input impedance of the antenna as a function of frequency. It is seen that the antenna exhibits a series resonance (where the imaginary part of the input impedance is zero with a real part less than 50) at 57 GHz and a parallel resonance (where the imaginary part of the input impedance is zero with a real part larger than 50) at 60 GHz, respectively. Fig. 3.2 (b) shows the simulated $|S_{11}|$ of the antenna as a function of frequency. It is evident that the antenna matches poor at 57 GHz but well at 60 GHz to a 50Ω source. At 52 GHz the magnitude $|S_{11}|$ is -8 dB indicating an acceptable matching, so it is chosen to examine the behaviour of the antenna in the non-resonant condition. Fig. 3.2 (c) shows the peak realized gain of the antenna as a function of frequency. It can be seen that the maximum gain of 14 dBi appears at the parallel resonance 60 GHz with a 3-dB gain BW of 5 GHz.

Fig. 3.1: Configuration of a typical MS GAA.

(a)

(b)

(c)

Fig. 3.2: Simulated performance of the MS GAA in Fig. 3.1: (a) input impedance, (b) $|S_{11}|$ and (c) peak realized gain.

Fig. 3.3 (a) shows the current distribution of the MS GAA at the parallel resonance 60 GHz. It is interesting to note that the current on each short side is indeed in phase, while the current on each long side changes the phase at its middle to form a full wavelength current. Consequently, it is the current on the short side that contributes to the far-field radiation more dominantly than that on the long side. The far-field radiation patterns in both E- (parallel with the short side of the grid) and H-planes (parallel with the long side of the grid) of the antenna at the parallel resonance 60 GHz are shown in Figs. 3.3 (b) and (c), respectively. As expected, the main beam of radiation is in the broadside direction. The side-lobe level is quite low. The stronger co-polarized component (defined as the electric field vector that lies parallel with the dominant electric field distribution among the antenna's aperture) is due to the existence of the same phase of the currents on all the short sides, while the weaker cross-polarized component (defined as the electric vector that lies orthogonally with the dominant electric field distribution among the antenna's aperture) is due to the presence of anti-phase current couplets on all the

long sides. The phase transition of the currents on the long sides accounts for the nulls on the cross-polarized patterns.

Fig. 3.3: Simulated performance of the MS GAA in Fig. 3.1 at parallel resonance 60 GHz: (a) current distribution, (b) normalized E-plane radiation patterns and (c) normalized H-plane radiation patterns.

Fig. 3.4 (a) shows the current distribution of the antenna at series resonance 57 GHz. Note that the phase of currents on each short side is no longer the same which indicates that not all the short sides can work as effective radiating elements simultaneously, resulting in lower gain in Fig. 3.2 (c). Besides, strong anti-phase current can be detected near the feeding point which is another reason for lower gain at the series resonance. The

current also tends to point away from the antenna, leading to the far-field radiation patterns shown in Figs. 3.4 (b) and (c). The main beam of radiation in the H-plane is slightly shifted from the broadside direction and the undesired cross-polarized radiation is greatly enhanced by in-phase currents on the long side.

Fig. 3.4: Simulated performance of the MS GAA in Fig. 3.1 at series resonance 57 GHz: (a) current distribution, (b) normalized E-plane radiation patterns and (c) normalized H-plane radiation patterns.

Fig. 3.5 (a) shows the current distribution of the antenna at non-resonant frequency 52 GHz. It is seen that a large portion of the grids does not provide effective radiation for the scarcity of currents on them. The radiation patterns in both E- and H-planes of the

antenna as shown in Figs. 3.5 (b) and (c) are squinted at 52 GHz. The polarization purity becomes degraded.

Fig. 3.5: Simulated performance of the MS GAA in Fig. 3.1 at non-resonance 52 GHz: (a) current distribution, (b) normalized E-plane radiation patterns and (c) normalized H-plane radiation patterns.

The revisit has confirmed the basic principle of operation of MS GAAs. The radiation is controlled by the grid geometries. The resonant frequency depends on the grid dimensions. The resonant characteristics of the whole array structure determine the overall frequency response. More importantly, the revisit has revealed that the MS GAA performs better at parallel resonance than at series resonance for 60-GHz applications.

3.2 Analysis of Microstrip Grid Array Antenna

The array theory [62], the transmission line and S-parameter method [59], the method of moments (MoM) [66], [68], [91] and the finite-difference time-domain (FDTD) method, [64], [67] have been employed to analyze MS GAAs on substrates of lower permittivity at lower frequencies. In this section, it presents an extensive analysis of MS GAAs on a variety of substrates at the 60-GHz band using Ansoft HFSS, which is a full-wave finite element solver. MS GAAs of 4, 9, and 16 grids are analyzed. The MS GAA of 9 grids is shown in Fig. 3.1, which has 16 short sides (or 16 radiating elements). MS GAAs of 4 and 16 grids have 9 and 25 radiating elements, respectively. The substrates are 0.254 mm ($0.0508\lambda_0$), 0.381 mm ($0.0762\lambda_0$) and 0.508 mm ($0.1016\lambda_0$) thick, respectively. The substrates have the same planar dimensions $a = b = 25$ mm ($5\lambda_0 \times 5\lambda_0$). The grids and the ground planes are 0.017 mm thick copper. As relative permittivity ϵ_r and loss tangent δ for the majority of marketed substrates are unknown at 60 GHz, it assumes $\epsilon_r = 2.2$ and $\tan \delta = 0.0009$ for Rogers 5880, $\epsilon_r = 3.66$ and $\tan \delta = 0.004$ for Rogers 4350 and fused silica, $\epsilon_r = 4.5$ and $\tan \delta = 0.002$ for Rogers TMM4, $\epsilon_r = 6$ and $\tan \delta = 0.0023$ for Rogers TMM6 and Ferro A6 low temperature co-fired ceramic (LTCC), $\epsilon_r = 9.2$ and $\tan \delta = 0.0022$ for Rogers TMM10 and Kyocera GL950 LTCC, and $\epsilon_r = 10.2$ and $\tan \delta = 0.003$ for Rogers 3210 in the analysis. It also fixes the feed placement as well as the characteristic impedances to be 100Ω for the long sides and 100Ω for the short sides of the MS GAAs throughout the analysis.

3.2.1 Resonant Frequency

The resonant frequency is at which the parallel resonance occurs to the MS GAA. Fig. 3.6 shows the simulated resonant frequency for the three different thicknesses as a function of relative permittivity with the number of radiating elements as a parameter. As the substrate thickness has been taken into account in calculating the effective relative permittivity to design the MS GAA, it can be noted that the resonant frequency changes slightly with the substrate thickness or the number of radiating elements. For instance, the resonant frequency changes within 5% for the most cases with a fixed relative permittivity as the substrate thickness h varies from 0.254, 0.381, and 0.508 mm or the number of radiating elements n varies among 9, 16 and 25. Further note that the resonant frequency generally decreases with an increase in relative permittivity.

(a)

(b)

(c)

Fig. 3.6: Resonant frequency as a function of relative permittivity ϵ_r : (a) $h = 0.254$ mm, (b) $h = 0.381$ mm and (c) $h = 0.508$ mm.

3.2.2 Impedance Bandwidth

The impedance BW of a MS GAA is typically from 6 % to 12 % for $|S_{11}| \leq -10$ dB [59].

Fig. 3.7 shows the simulated impedance BW for the three different thicknesses as a function of relative permittivity with the number of radiating elements as a parameter. As with all MS structures, the substrate thickness and relative permittivity has a strong

influence on the impedance BW. Generally, a thicker substrate with a lower relative permittivity offers the MS GAA with a wider impedance BW. Taking the MS GAA of 9 radiating elements on the substrate of $\epsilon_r = 2.2$ and $h = 0.1016\lambda_0$ as an example, it achieves 36 % impedance BW. Similar to other resonant array antennas, the impedance BW of the MS GAA depends on the number of the radiating elements. Simply stated, for a given substrate size and controlled by the gain-BW relationship, the more radiating elements are, the narrower the impedance BW is.

(a)

(b)

(c)

Fig. 3.7: Impedance bandwidth as a function of relative permittivity ϵ_r : (a) $h = 0.254$ mm, (b) $h = 0.381$ mm and (c) $h = 0.508$ mm.

3.2.3 Half-Power Beamwidth

The half-power beamwidth of a MS GAA can be wider or narrower, depending mainly on the number of radiating elements. Fig. 3.8 shows the simulated half-power beamwidth across the main radiation lobe in the E-plane for the three different thicknesses as a function of relative permittivity with the number of radiating elements as a parameter. The half-power beamwidth in the H-plane is very similar. As expected, the half-power beamwidth decreases when the number of radiating elements. The half-power beamwidth is sensitive to the substrate relative permittivity but is not sensitive to the substrate thickness. The higher the substrate relative permittivity is, the wider the half-power beamwidth is, due to the excitation of stronger surface waves.

(a)

(b)

(c)

Fig. 3.8: Half-power beamwidth as a function of relative permittivity ϵ_r : (a) $h = 0.254$ mm, (b) $h = 0.381$ mm and (c) $h = 0.508$ mm.

3.2.4 Gain

A MS GAA is a high-gain antenna [56], [59], [69]. Fig. 3.9 shows the simulated peak realized gain in the broadside direction for the three different thicknesses as a function of relative permittivity with the number of radiating elements as a parameter. Note that the gain decreases with relative permittivity or thickness but increases with the number of radiating elements. The gain increment with the number of radiating elements will be limited by the attenuation of the MS lines that form the array. The gain decrement due to a higher permittivity or a thicker substrate can be attributed mainly to the excitation of stronger surface waves. The larger attenuation of the MS lines due to a higher permittivity and the more feeding loss for a thicker substrate also reduce the gain. Furthermore, the control over phase synchronization for more radiating elements on a thicker substrate of a higher permittivity becomes more difficult for a high gain.

(a)

(b)

(c)

Fig. 3.9: Peak realized gain as a function of relative permittivity ϵ_r : (a) $h = 0.254$ mm, (b) $h = 0.381$ mm and (c) $h = 0.508$ mm.

3.2.5 3-dB Gain Bandwidth

In addition to the impedance BW, a number of BWs can be defined under the radiation pattern characteristics (i. e., pattern shape, sidelobe level, gain, cross polarization, etc.). It is found that the most stringent BW criterion is seen to be the 3-dB gain BW [59]. Fig. 3.10 shows the simulated 3-dB gain BW for the three different thicknesses as a function of relative permittivity with the number of radiating elements as a parameter. Note that the 3-dB gain BW of the MS GAA decreases with the number of the radiating elements, which is similar to other resonant array antennas. The 3-dB gain BW is sensitive to the substrate thickness. The thicker the substrate is the wider the 3-dB gain BW is.

(a)

(b)

(c)

Fig. 3.10: 3-dB gain bandwidth as a function of relative permittivity ϵ_r : (a) $h = 0.254$ mm, (b) $h = 0.381$ mm and (c) $h = 0.508$ mm.

3.2.6 Radiation Patterns

In earlier literatures, the radiation mechanism of MS GAAs was considered equal with a planar array of half-wave dipoles, however it was not verified. In this thesis, the MS GAA is modeled as a uniform planar array as shown in Fig. 3.11. The co-polarized radiation patterns from simulation and calculation are compared, verifying the hypothesis of GAA's radiation mechanism for the first time.

Fig. 3.11: Array factor model of a MS GAA with 16 radiating elements.

The far-field co-polarized radiation pattern of a microstrip half-wave dipole on its E-plane could be expressed as [92]:

$$E_{\theta} = \frac{j\omega I_0 L \sin \theta e^{j\omega[t-(r/c)]}}{4\pi\epsilon_0 c^2 r} = j \frac{I_0 \beta L}{4\pi\epsilon_0 c r} \sin \theta e^{j\omega[t-(r/c)]} \quad (3.1)$$

which is plotted in Fig. 3.12.

Fig 3.12: Normalized E-plane radiation pattern of a microstrip half-wave dipole.

The array factor (AF) of uniform planar array with isotropic sources in Fig. 3.11 can be expressed as in (2.8) with

$$dx = dy = \sqrt{2}s = \frac{\sqrt{2}}{2}l \quad (3.2)$$

which is plotted in Fig. 3.13 with n as the number of radiating elements.

(a)

(b)

(c)

Fig 3.13: Normalized AFs of uniform planar arrays: (a) $n = 9$, (b) $n = 16$ and (c) $n = 25$.

Applying

$$E(\text{total}) = [E(\text{single element})] \times [\text{Array factor}] \quad (3.3)$$

radiation patterns of MS GAAs with different number of radiating elements can be obtained. Fig. 3.14 compares the calculated and simulated radiation patterns of MS GAAs with different number of radiating elements, verifying the hypothesis that the MS

GAA could be modeled as a planar uniform array of MS half-wave dipoles from the perspective of radiation.

(a)

(b)

(c)

Fig. 3.14: Normalized E-plane radiation patterns of MS GAAs: (a) $n = 9$, (b) $n = 16$ and (c) $n = 25$.

3.3 Synthesis of Microstrip Grid Array Antenna

The synthesis of MS GAAs has been attempted [59], [62], [69]; however, no direct synthesis method or design procedure has been developed. In this section, we describe the synthesis and outline the design steps required to implement a MS GAA. By specifying the required gain and the substrate's characteristics like permittivity and thickness, dimensions of a microstrip GAA can be directly calculated by the series of equations provided in this section.

In synthesizing a MS GAA to satisfy the specific performance requirements, the choice of the substrate must be made first to know the relative permittivity ϵ_r , loss tangent δ , and thickness h . Then, the grid lengths l and s for the MS GAA to operate at the required resonant frequency f_o can be calculated as

$$l = \lambda_g = \lambda_0 / \sqrt{\epsilon_{eff}} = c / f_o \sqrt{\epsilon_{eff}} \quad (3.4)$$

and

$$s = \lambda_g / 2 = l / 2 \quad (3.5)$$

for phase synchronization where c is the velocity of light in free space and ϵ_{eff} is the substrate effective permittivity.

The choice of the grid widths w_l and w_s is more involved. It is governed by the impedance, transmission, and radiation properties. As previously stated, the long sides act mainly as transmission lines. The currents on them generate the undesired cross-polarized radiation. Hence, the width w_l should be narrower to reduce it. The narrower width w_l implies the higher characteristic impedance as the characteristic impedance is approximately inversely proportional to the width w_l . The narrower width w_l also causes the higher transmission loss. As the transmission loss increases rapidly when the characteristic impedance is above 250Ω [93], the width w_l can be chosen to satisfy

$$Z_{0l}(w_l) \leq 250 \Omega \quad (3.6)$$

where Z_{0l} represents the characteristic impedance of the long sides.

The short sides act as both radiating elements and transmission lines. The currents on them produce the desired co-polarized radiation. The currents can be uniformly distributed or tapered by varying the width w_s . To find the proper width w_s , the radiation resistance of a short side is evaluated. Fig. 3.15 shows the simulated radiation resistance at the parallel resonance 60 GHz for the three different thicknesses and the same loss tangent $\tan \delta = 0.002$ as a function of the width w_s with the relative permittivity as a parameter. Generally, the radiation resistance decreases noticeably with the width for $w_s < 0.3$ mm but slowly with the width for $w_s > 0.3$ mm. The radiation resistance is higher for lower relative permittivity and is lower for thicker substrate. The radiation resistance varies from 100 to 850Ω and can be considered as the loads to the transmission lines of

both long and short sides. The application of matching condition at the input terminal to the source yields the required width w_s .

(a)

(b)

(c)

Fig. 3.15: Radiation resistance as a function of the width of the short side w_s : (a) $h = 0.254$ mm, (b) $h = 0.381$ mm and (c) $h = 0.508$ mm.

Based on the formula for the radiation resistance of a resonant half-wavelength MS line fed at one end [94] and simulated data, we can get the formula to calculate the radiation resistance of a short side by curve fitting. It applies when the number of radiating element $n \leq 25$, and can be expressed as:

$$R_r = \epsilon_{eff} \cdot Z_{0s}^2 \cdot \xi / (120 \cdot I) \quad (3.7)$$

where Z_{0s} represents the characteristic impedance of the short side and

$$I = \left(\frac{2\pi}{\lambda_0} h \right)^2 \times \left\{ 1.3 - 4 / (3\epsilon_r) + 0.53 / \epsilon_r^2 - [0.08856 - 0.08856 / \epsilon_r + 0.03795 / \epsilon_r^2] \times \left(\frac{\pi}{\lambda_0} w_s \right)^2 \right\} - [0.248714 - 0.370371 / \epsilon_r + 0.159887 / \epsilon_r^2] \times (1 / \epsilon_{eff}) \quad (3.8)$$

with our empirical factor

$$\begin{aligned} \xi = & -0.10283 + 0.76642h + 1.96385w_s - 5.63524hw_s + 5.54382h^2w_s \\ & - 2.23237w_s^2 + 14.30744hw_s^2 - 17.3777h^2w_s^2 \end{aligned} \quad (3.9)$$

for $0.0375 \leq h/\lambda_0 \leq 0.0625$,

$$\xi = \left[-0.10283 + 0.76642h - \left(163971 - 14.30744h + 17.3777h^2 - 6.09502e^{\frac{-h}{0.08608}} \right) w_s \right] \times (1.06527 + 0.00644\epsilon_r - 0.01467\epsilon_r^2) \quad (3.10)$$

for $0.0625 \leq h/\lambda_0 \leq 0.0875$, and

$$\xi = \left[-0.10283 + 0.76642h + \left(\frac{0.59266 + 0.63381w_s}{+6.05902e^{\frac{-h}{0.08608}} - 61.28545w_s e^{\frac{-h}{0.04904}}} \right) w_s \right] \times (2.51875 - 0.5172\epsilon_r + 0.02859\epsilon_r^2) \quad (3.11)$$

for $0.0875 \leq h/\lambda_0 \leq 0.12$. It is needed to stress that (3.8) is derived from asymptotic expansions that not only makes it simple to calculate the radiation power integral but also lends insight into the dependence of the radiation resistance upon the line width and substrate permittivity [94]. Nevertheless, some loss in accuracy is encountered and overcome by the introduced empirical factor.

Next, the substrate length a and width b can be determined from the number of grids required by the gain, when the number of radiating elements $n \leq 25$. The gain in dBi for the MS GAA can be estimated as

$$\text{Gain(dBi)} = \frac{g_1(n) + g_2(n)\epsilon_r + g_3(n)\epsilon_r^2 + 10\log n}{\epsilon_r^3 h} \quad (3.12)$$

where

$$\begin{cases} g_1(n) = 216 - 0.52n + 0.1n^2 \\ g_2(n) = -143.2 - 1.13n - 0.058n^2 \\ g_3(n) = 30.3 + 0.74n \end{cases} \quad (3.13)$$

for $0.0375 \leq h/\lambda_0 \leq 0.0625$,

$$\begin{cases} g_1(n) = -505 + 72.62n - 2.31n^2 \\ g_2(n) = 133.57 - 26.52n + 0.83n^2 \\ g_3(n) = 17.43 + 1.77n - 0.038n^2 \end{cases} \quad (3.14)$$

for $0.0625 \leq h/\lambda_0 \leq 0.0875$, and

$$\begin{cases} g_1(n) = -362 + 22n - 0.59n^2 \\ g_2(n) = 141 - 10n + 0.24n^2 \\ g_3(n) = 9 + 1.46n - 0.024n^2 \end{cases} \quad (3.15)$$

for $0.0875 \leq h/\lambda_0 \leq 0.12$.

Finally, the feed point must be located, typically at the joint of the long and short sides near the geometric centre of the antenna. The diameter p of the feed pin and the diameter d of the aperture on the ground plane should be complied with the feeding cable or connector.

In the following, the actual design procedure is demonstrated for an example of the MS GAA to operate at 60 GHz with the maximum gain of ≥ 13.5 dBi and the impedance BW of 7 GHz [37], [39], [42], [44], [45], [95], [96]. The Ferro A6S LTCC is chosen to be the substrate, which has a relative permittivity of 5.9, a loss tangent of 0.002. The substrate is 0.385 mm ($0.077\lambda_0$) thick to avoid the excitation of the TE_1 mode surface wave. For the sake of low cost and easy fabrication, the width and thickness of the grids are kept uniform as 0.15 mm and 0.01 mm, respectively. The grids are made of gold. At 60 GHz, the substrate effective permittivity is $\epsilon_{eff} = 4.3$ and the guided wavelength is $\lambda_g = 2.42$ mm. The grid lengths are $l = 2.5$ mm $\approx \lambda_g$ and $s = 1.365$ mm $\approx \lambda_g/2$. The characteristic impedances for the grid long and short sides are equal $Z_{0l} = Z_{0s} = 102 \Omega$ for the width $w_l = w_s = 0.15$ mm. The radiation resistance of a short side is $R_r = 250 \Omega$.

Given the specified gain value of 13.5 dBi, one can find the required number of grids to be at least 14, which leads to an estimation of the length and width of the substrate as $a = 11.5$ mm and $b = 5$ mm, respectively. Considering the requirement for singulation in LTCC, we enlarge the length $a = 13.5$ mm and the width $b = 8$ mm. The diameter of the feed pin is $p = 0.15$ mm and the diameter of the aperture on the ground plane is $d = 0.3$ mm. Fig. 3.16 shows the photograph of the fabricated MS GAA.

Fig. 3.16: Photograph of the fabricated 60-GHz MS GAA.

Fig. 3.17 shows the measured performance of the MS GAA. It is evident from Fig. 3.17 (a) that the measured impedance BW is 8.7 GHz from 56.3 to 65 GHz, which covers the whole band of 57-64 GHz. Figs. 3.17 (b) and (c) compare the simulated with the measured radiation patterns of the MS GAA. They reveal that both E- and H-plane patterns have the main beam of radiations in the broadside directions with small cross polarization components. Note that the measured and simulated co-polarized patterns at 60 GHz agree very well.

(a)

Fig. 3.17: Measured performance of the MS GAA: (a) $|S_{11}|$, normalized radiation patterns at 60 GHz: (b) E-plane and (c) H-plane.

3.4 Summary

In this chapter, the performance of MS GAAs is analyzed and a synthesis method is proposed. By investigating GAAs' performance at non-resonance, series resonance and parallel resonance, it is found that for 60-GHz applications GAAs feature preferable characteristics at parallel resonance. Through a comprehensive study of MS GAAs' resonant frequency, impedance BW, half-power beamwidth, gain, 3-dB gain BW and radiation patterns with different number of radiating elements on a wide selection of substrates. Based on the analysis, a synthesis method is proposed. A design example of the GAA on Ferro A6S LTCC verifies the synthesis method.

Chapter 4 Design of the Linearly-Polarized Grid Array Antenna for 60-GHz Antenna-in-Package Application

The design of Low Temperature Co-fired Ceramic (LTCC)-based linearly-polarized (LP) grid array antenna (GAA) is introduced in this chapter, which includes the design of radiating array, integration of the radiating array with the feeding network, design and optimization of the package.

4.1 Design of the 60-GHz Linearly-Polarized Grid Array Antenna

Ferro A6M LTCC design rules of the conductor width, the gap between two adjacent conductors, and via diameter ≥ 0.1 mm for mass production are followed. The A6M ceramic type has the nominal dielectric constant of 5.9 and loss tangent of 0.002 and its metallic paste is gold with good conductivity after firing.

4.1.1 Design of the Radiating Array

To achieve highly directive antenna performance, usually a large number of radiating elements are utilized to construct an array. The GAA is one kind of many array antennas. Performance of the GAA depends on the current synchronization on the grid. However, when increasing the number of grids in a GAA, the directivity does not increase proportionally, which can be explained by the unsynchronized currents in Fig. 4.1 (a)

showing the simulated 61.5-GHz current distribution of the LP GAA with 32 grids. The length of long sides $l = 2.24 \text{ mm} \approx \lambda_g$ and the width $w_l = 0.12 \text{ mm}$ and short sides of the length $s = 1.12 \text{ mm} \approx \lambda_g/2$ and the width $w_s = 0.12 \text{ mm}$. Ferro A6M LTCC ($\epsilon_r = 5.9$, $\tan \delta = 0.002$) substrate with a size of 15 mm long and wide and 0.4 mm thick is the substrate. It is fed at the marked point by a pin of diameter 0.15 mm through an aperture of diameter 0.3 mm in the ground plane. Note immediately from the figure that the incorrect current distribution occurs on grids that are far away from the feeding point at 61.5 GHz which is the central frequency of the design band. This is because that the phase error brought about by each grid may add up to large phase distortion for the grids far away from the feeding point. However, the E- (parallel with the short side of the grid) and H-plane (parallel with the long side of the grid) radiations pattern as shown in Figs. 4.1 (d) and (e) are not distorted greatly. The phasing of the instantaneous currents becomes more unsynchronized at the lowest frequency of 57 GHz and the highest frequency of 66 GHz of the design band, giving rise to the poor patterns shown in Figs. 4.1 (b) and (c) to (f) and (g) with large side lobes, lower gain and narrower 3-dB gain bandwidth (BW).

(a)

(b)

(c)

(d)

(e)

Fig. 4.1: Simulated performance of the LP GAA without sub grid arrays: (a) current distribution at 61.5 GHz, normalized radiation patterns at: (b) 57 GHz E-plane, (c) 57 GHz H-plane, (d) 61.5 GHz E-plane, (e) 61.5 GHz H-plane, (f) 66 GHz E-plane and (g) 66 GHz H-plane.

To achieve more synchronized currents on the GAA with a large number of grids, one can use the dual-feed technique [97]. In this thesis, we explore to form a large GAA with sub grid arrays and multiple feeds. For the small number of grids in the sub grid array, the accumulated phase error by each grid would not be as severe as that in a large array, resulting in better synchronized current. Fig. 4.2 shows top and bottom views of the simulated currents on the sub grid array. The sub grid array structure is chosen on the consideration for easy matching and good axial ration of the CP GAA in Chapter 5. It is designed on the top surface of a square LTCC substrate (length = 7 mm, width = 7 mm, and thickness = 0.4 mm) and is fed at a single point by a pin having the diameter of 0.15 mm through an aperture having a diameter of 0.3 mm on the ground plane on the bottom surface of the substrate. Dimensions of each grid are the same as Fig. 4.1. It is interesting to note that the currents on each short side are indeed in phase, while the currents on each long side experience a phase reverse in its middle to form a full wavelength current loop. Consequently, it is the currents on the short side that contributes to the far-field

radiation more dominantly than that the one from the long side. Better current synchronization is achieved when comparing with Fig. 4.1 (a).

Fig. 4.2: The sub grid array: (a) top view with the simulated current distribution and (b) bottom view.

Fig. 4.3 (a) shows the simulated $|S_{11}|$ of the LP sub grid array antenna versus frequency. It is evident from the figure that the impedance BW for $|S_{11}| \leq -10$ dB is from 51 to 64.8 GHz. Fig. 4.3 (b) shows the peak realized gain of the antenna versus frequency. It can be seen that the maximum gain of 12.6 dBi occurs at 62 GHz with a 3-dB gain BW of 9.8 GHz. The far-field radiation patterns in both E- and H-planes are shown in Figs. 4.3 (c) to (h), respectively. As expected, the main beam of radiation is definitely in the broadside direction in the 57 – 66 GHz band. The sidelobe level is lower than 15 dB. The stronger co-polarized E-field component is due to the existence of the in-phase current flowing on all the short sides of the element, while the weaker cross-polarized one (especially in the broadside direction) is due to the presence of out-of-phase current coupling to all the long sides. The phase transition of the currents on the long sides accounts for the nulls of the cross-polarized patterns [59].

(a)

(b)

(c)

(d)

(e)

(f)

Fig. 4.3: Simulated performance of the sub grid array: (a) $|S_{11}|$, (b) peak realized gain, normalized radiation patterns at: (c) 57 GHz E-plane, (d) 57 GHz H-plane, (e) 61.5 GHz E-plane, (f) 61.5 GHz H-plane, (g) 66 GHz E-plane and (h) 66 GHz H-plane.

The formation of a large GAA with sub grid arrays and multiple feeds also introduces sufficient application flexibility. For example, an LP GAA can be realized when sub grid arrays in the upper and lower quadrants are fed out-of-phase simultaneously, which will be introduced in the following part. Fig. 4.4 (a) shows the simulated current distribution at 61.5 GHz of the LP GAA of 32 grids formed by 4 sub grid arrays and 4 feeds. The grid and feed dimensions are the same as those given in Fig. 4.2. The gaps between the sub grid arrays are kept the same at 0.5 mm. It is clear from the figure that the current distribution is more satisfactory than that shown in Fig. 4.1 (a). The improvement is due to the fact that one feed excites 8 grids rather than 32 grids, leading to a smaller accumulated phase error. Although there are still incorrect current probably caused by the coupling between the sub grid arrays and multiple feeds, the radiation patterns as shown in Figs. 4.4 (b) to (g) are improved at 57, 61.5, and 66 GHz. The maximum radiation is in the broadside direction of the GAA in the 57 – 66 GHz band.

(a)

(b)

(c)

Fig. 4.4: Simulated performance of the LP GAA with sub grid arrays: (a) current distribution at 61.5 GHz, normalized radiation patterns at: (b) 57 GHz E-plane, (c) 57 GHz H-plane, (d) 61.5 GHz E-plane, (e) 61.5 GHz H-plane, (f) 66 GHz E-plane and (g) 66 GHz H-plane.

When the adjacent sub grid arrays are arranged orthogonally and the 4 feeds are excited with a 90° -phase increment clockwise or counterclockwise, the horizontal and vertical electric field components can combine into either right-handed (RH) or left-handed (LH) circularly-polarized (CP) radio waves to realize a CP GAA, which will be introduced in detail in Chapter 5. Fig. 4.5 (a) shows the simulated current distribution at

61.5 GHz of the CP GAA of 32 grids formed by 4 sub grid arrays and 4 feeds. Figs. 4.5 (b) to (g) show the radiation patterns of the CP GAA at 57, 61.5, and 66 GHz, respectively. As can be seen, the regular patterns are obtained with the maximum radiation in the broadside direction at the three frequencies.

(a)

(b)

(c)

Fig. 4.5: Simulated performance of the CP GAA with sub grid arrays: (a) current distribution at 61.5 GHz, normalized radiation patterns at: (b) 57 GHz $\phi = 0^\circ$, (c) 57 GHz $\phi = 90^\circ$, (d) 61.5 GHz $\phi = 0^\circ$, (e) 61.5 GHz $\phi = 90^\circ$, (f) 66 GHz $\phi = 0^\circ$ and (g) 66 GHz $\phi = 90^\circ$.

Moreover, when the sub grid arrays in the upper quadrants are driven simultaneously with 0° phase and those in lower quadrants with 180° phase in Fig. 4.4 (a), a single balanced antenna is realized which is compatible with a system of balanced output. While if the sub grid arrays in the left quadrants and in the right quadrants are

respectively driven differentially (that is to say, the sub grid arrays in the left quadrants are driven by a differential pair and those in the right quadrants are driven by the other differential pair) two balanced antennas with one for transmission and the other for reception are realized. Fig 4.6 (a) shows the simulated current distribution of the single balanced GAA at 61.5 GHz. Figs. 4.6 (b) to (g) show the simulated radiation patterns of the single balanced LP GAA with 4 sub grid arrays and 4 feeds at 57, 61.5 and 66 GHz. Fig. 4.7 (a) shows the simulated current distribution when one of the dual balanced GAAs is excited at 61.5 GHz. Figs. 4.4 (b) to (g) show the simulated radiation patterns of one of the dual balanced LP GAAs at 57, 61.5 and 66 GHz. As expected, the beamwidth becomes wider in the H-plane due to less number of radiating elements (or short sides) in the plane. The main lobe of radiation tilts slightly at 57 and 66 GHz.

(a)

(b)

(c)

(d)

(e)

Fig. 4.6: Simulated performance of the single balanced LP GAA: (a) current distribution at 61.5 GHz, normalized radiation patterns at: (b) 57 GHz E-plane, (c) 57 GHz H-plane, (d) 61.5 GHz E-plane, (e) 61.5 GHz H-plane, (f) 66 GHz E-plane and (g) 66 GHz H-plane.

(b)

(c)

(d)

(e)

Fig. 4.7: Simulated performance of the dual balanced GAAs: (a) current distribution at 61.5 GHz, normalized radiation patterns at: (b) 57 GHz E-plane, (c) 57 GHz H-plane, (d) 61.5 GHz E-plane, (e) 61.5 GHz H-plane, (f) 66 GHz E-plane and (g) 66 GHz H-plane.

As pointed out by Conti, et al. that the GAA is an interconnected array concept rather than an assembly of individually-matched radiators, the radiation BW is more relevant and restrictive than impedance BW for the GAA [59]. The radiation BW over which the maximum radiation remains in the broadside direction has been shown to be sufficient for the GAAs with 4 sub grid arrays and 4 feed points. The most stringent radiation BW is the 3-dB gain BW. Fig. 4.8 shows the simulated peak realized gain as a function of frequency. It is seen that the GAA of Fig. 4.1 (a) has the narrowest 3-dB gain BW and cannot cover the whole 60-GHz band from 57 to 66 GHz. The 3-dB gain BW of the single unbalanced LP GAA in Fig. 4.4 (a) is wider enough for the whole 60-GHz band, so is the single balanced LP GAA. Both the maximum gain values are enhanced by about 2 dB as compared with the GAA in Fig. 4.1 (a) because of improved current distribution. As expected, the maximum gain values of the single CP and dual balanced LP GAAs are about 3 dB lower. The 3-dB gain BWs are still enough to cover the whole 60-GHz band.

Fig. 4.8: Simulated peak realized gain.

4.1.2 Integration of the Radiating Array with the Feeding Network

Fig.4.9 (a) shows the exploded view of the LP GAA, in which the radiating array of four sub grid arrays is integrated with a quarter-wave matched T-junction feeding network. It has a size of $15 \times 15 \times 0.6 \text{ mm}^3$. The ceramic superstrate (sup) is 0.1 mm thick, which helps preventing the radiating elements from being scratched. The ceramic substrates 1 and 2 (sub₁ and sub₂) are 0.4 and 0.1 mm, respectively. The gold radiating array (rad), ground plane (gnd) and feeding network (fn) are all 0.01 mm thick. The superstrate does not bring about obvious frequency shift or gain decrement to the antenna for its moderate permittivity and low thickness, thus dimensions of the grid (Fig. 4.9 (b)) are kept the same as Fig. 4.4 (a). The sub grid arrays are fed out-of-phase in the upper and lower quadrants simultaneously by a quarter-wave matched T-junction microstrip feeding network (Fig. 4.9 (c)) on the bottom surface of sub₂. The fn has its input at Port 1 and outputs at Ports 2, 3, 4 and 5. The four output ports are connected to four sub grid arrays in quadrants 1, 2, 3 and 4 respectively by four vias of diameter 0.15 mm and length 0.5

mm through circular openings of diameters 0.3 mm on the gnd. The required phase at a feed point is realized by controlling the length of the microstrip line. The patterned gnd layer has many $0.4 \text{ mm} \times 0.4 \text{ mm}$ openings, which help to release different stress from ceramic and metallic layers to avoid warpage and micro fractures. They are deliberately made not directly under the short sides of grids or above the feeding network to reduce gain penalty and to minimize the detrimental coupling effect between the rad and the fn [97].

Fig. 4.9: The LP GAA: (a) exploded view, (b) radiating array and (c) feeding network.

4.1.3 Parametric Study

The grid dimensions and the gaps between the sub grid arrays are determined from a parametric study. The initial length of the grid short side s is half guided wavelength. The initial length of the grid long side l is twice as long as s . The widths of the grid short and long side w_s and w_l are governed by impedance, transmission and radiation properties [69].

For easy fabrication in large quantities, they are kept the same width with an initial value of 0.12 mm. Fig. 4.10 shows the simulated $|S_{11}|$ and peak realized gain as a function of frequency from 50 to 70 GHz with a varying length s and a fixed width $w_s = 0.12$ mm. It is seen from Fig. 4.10 (a) that both the lower and upper frequency edges are controlled by the length s . They shift upward for a shorter s and downward for a longer s . It is evident from Fig. 4.10 (b) that the maximum gain is slightly affected for the length s varying from 1.10 to 1.15 mm. The dependence of the 3-dB gain BW on the length s is

similar to that of the impedance BW. It is also the length $s = 1.12$ mm that the 3-dB gain BW is satisfied.

(a)

(b)

Fig. 4.10: Simulated (a) $|S_{11}|$ and (b) peak realized gain of the LP GAA with length s as a parameter.

Figs. 4.11 (a) and (b) show the simulated $|S_{11}|$ and peak realized gain as a function of frequency with a varying width w_s and a fixed length $s = 1.12$ mm. It is seen that both the impedance and 3-dB gain BWs are slightly affected by the width w_s varying from 0.10 to 0.15 mm. The width $w_s = 0.12$ mm yields the best performance.

(a)

(b)

Fig. 4.11: Simulated (a) $|S_{11}|$ and (b) peak realized gain of the LP GAA with width w_s as a parameter.

The gaps between the sub grid arrays affect the performance of the GAA. To simplify the design, we make the same separation g for the gaps. Figs. 4.12 (a) and (b) show the simulated $|S_{11}|$ and peak realized gain as a function of frequency with a varying separation g and a fixed length $s = 1.12$ mm and a fixed width $w_s = 0.12$ mm. It is evident that both the impedance and 3-dB gain BWs are affected by the separation g varying

from 0.3 to 0.7 mm. The separation $g = 0.5$ mm yields satisfactory performance with the smallest antenna size.

(a)

(b)

Fig. 4.12: Simulated (a) $|S_{11}|$ and (b) peak realized gain of the LP GAA with separation g as a parameter.

4.1.4 Measured Performance

The LP GAA was fabricated by VTT Finland and tested by LEAT-CNRS, Université Nice Sophia Antipolis. The measurement is limited to 67 GHz. Fig. 4.13 shows the top and bottom view of the fabricated LP GAA. The radiating array is covered by a superstrate as shown in Fig. 4.13 (a) and the feeding network can be seen directly on the bottom. Measured results in Fig. 4.14 show that the impedance BW of the LP GAA is 57.5-62 GHz, the 3-dB gain BW is 57.5-65 GHz with the maximum gain 16.5 dBi at 60.7 GHz, vertical beams appear in broadside direction 57-64 GHz. Disagreement between the simulation and measurement is caused by the fabrication tolerance of LTCC process, in which it is usually seen 7 – 10 % vertical substrate shrinkage, a variation of substrate permittivity 5.7 – 5.9 and loss tangent 0.002 – 0.0025.

Fig. 4.13: Photographs of the fabricated LP GAA: (a) top view and (b) bottom view.

(a)

(b)

(c)

(d)

(e)

(f)

Fig. 4.14: Measured performance of the LP GAA: (a) $|S_{11}|$, (b) peak realized gain, normalized radiation patterns at: (c) 57 GHz E-plane, (d) 57 GHz H-plane, (e) 60 GHz E-plane, (f) 60 GHz H-plane, (g) 64 GHz E-plane and (h) 64 GHz H-plane.

4.2 Packaging of the 60-GHz Linearly-Polarized Grid Array Antenna

In this section, the formerly design LP GAA is transformed into a package. First we introduce the package layout. Then the package is optimized from the perspective of parallel mode distribution. Finally the AiP is fabricated and tested.

4.2.1 Design of the Package

The packaged LP GAA has dimensions of $15 \text{ mm} \times 15 \text{ mm} \times 1 \text{ mm}$. Fig. 4.15 (a) shows its exploded view. It consists of five ceramic and five metallic layers. All the metallic layers are made of 0.01 mm thick gold. The top ceramic layer (sup) is 0.1 mm thick. The radiating array (rad) is below the superstrate as shown in Fig. 4.15 (b). To minimize mutual coupling between the feeding network and other components like filters,

amplifiers etc., the microstrip feeding network in Fig. 4.9 (a) is replaced by a stripline quarter-wave matched T-junction feeding network. The metallic ground plane (gnd_1) is for the radiating elements and the stripline feeding network (sfn). The ceramic substrates (sub_2 and sub_3) for the stripline feeding network have an equal thickness of 0.2 mm. As shown in Fig. 4.15 (c), the stripline feeding network has its input at Port 1 and outputs at Ports 2, 3, 4 and 5. The four output ports are connected to four sub microstrip grid arrays in quadrants 1, 2, 3 and 4 respectively by four vias of length 0.6 mm through four circular openings of diameters 0.3 mm on gnd_1 . The fencing vias of length 0.4 mm are located around each feeding via to connect the first and second ground planes. The bottom ceramic layer (sub_4) is 0.1 mm thick. It provides the substrate for the ground-signal-ground (GSG) testing pad or the power, signal and ground traces to be flip-chip bonded with the 60-GHz radio die. The testing pad (tp) shown in Fig. 4.15 (d) are designed for a GSG probe having a 250 μm pitch. The two square pads of 0.4 mm wide, grounded by conducting vias, are located beside the 50- Ω microstrip line. The gap between a grounded pad and the microstrip line is 0.1 mm. The microstrip line is connected to the stripline feeding network by a via of length 0.3 mm through a circular opening of diameter 0.3 mm on the second ground plane (gnd_2).

Fig. 4.15: The packaged LP GAA: (a) exploded view, (b) radiating array, (c) feeding network and (d) testing pad.

4.2.2 Optimization of the Package

For the design of a multilayer structure, vertical transition of signal traces on different layers with low reflection and small insertion loss is essential [98]-[100]. In this section we optimize the vertical via transitions between a microstrip line and a stripline

(representing the transition from the testing pad to the feeding network) as well as between a stripline and a microstrip T-junction (representing the transition from the feeding network to the sub grid array) for the design of the AiP module.

A. Optimization of the Via Fence

The fence of vias is usually understood from the perspective of circuit according to which they provide return current paths. Hence, a fence of one to four vias is suggested. Fig. 4.16 illustrates the simulated electric field distribution between the gnd_1 and gnd_2 for the case of no fence vias at 60 GHz. It is seen that the undesirable parallel-plate mode is strongly excited, and then guided to the substrate edges where the mode undergoes reflection, radiation and diffraction. Fig. 4.17 shows the simulated E-field distribution between gnd_1 and gnd_2 for the case of five fences of two vias at 60 GHz (one fence at the input port of the feeding network and four fences at the output ports). As shown, a strong parallel-plate mode still exists, indicating the failure of integrating the fences of two vias. The fences of four vias become effective to suppress the parallel-plate mode as evident in Fig. 4.18. The parallel-plate mode is effectively suppressed by the fences of seven vias. Much less field leakage is observed from the gaps between the adjacent vias as shown in Fig. 4.19. It clearly displays that the stripline supports a TEM mode, on which the equal phase front of the guided wave is vertical to the ground planes gnd_1 and gnd_2 . The coax-via configuration formed by the fence of vias and the feed via also supports the TEM mode, but the equal phase front of the guided wave is parallel with the ground planes gnd_1 and gnd_2 . Therefore, when the vertical TEM mode reaches the end of the stripline and is guided through the coax-via to propagate in the horizontal TEM mode, reflection and diffraction as well as radiation are expected. Fences of more vias can ensure smoother field transition between the vertical and horizontal TEM modes, stronger suppression of the parallel-plate mode, as well as maintain an equal potential between

ground planes gnd_1 and gnd_2 . The effects of fences of vias on the performance of the packaged LP in terms of $|S_{11}|$ and peak realized gain are given in Fig. 4.20. As expected, the fences of seven vias produce the most satisfactory performance.

Fig. 4.16: E-field distribution for the case of no fence of vias in the packaged LP GAA: (a) top view and (b) zoomed side view.

Fig. 4.17: E-field distribution for the case of five fences of two vias in the packaged LP GAA: (a) top view and (b) zoomed side view.

(a)

(b)

Fig. 4.18: E-field distribution for the case of five fences of four vias in the packaged LP GAA: (a) top view and (b) zoomed side view.

(a)

Fig. 4.19: E-field distribution for the case of five fences of seven vias in the packaged LP GAA: (a) top view and (b) zoomed side view.

Fig. 4.20: Simulated (a) $|S_{11}|$ and (b) peak realized gain of the packaged LP GAA with the number of fencing vias as a parameter.

B. Optimization of the Openings on Ground Planes

Fig. 4.21 shows the HFSS model of the vertical via transition between the 50- Ω microstrip line of the testing pad and the 50- Ω stripline of the feeding network in a three-layer structure of planar dimensions 10 mm $\times$ 10 mm. Port 1 is at one end of the microstrip line, while Port 2 is at one end of the stripline. Both ports are de-embedded to avoid direct coupling in simulation. The radius of the feeding via is a . A circular opening of radius b is set around the feeding via on the common ground gnd_1 . There are fencing vias around the feeding via. Seven fencing vias of radius $c = 0.05$ mm are used. They are located every 45° at the distance of $b + c + 0.1$ mm around the feeding via to connect the ground planes gnd_1 and gnd_2 .

Fig. 4.21: HFSS Model of the via transition between microstrip and stripline.

The vertical via transition is optimized by setting a at 0.05 mm and varying b from 0.10 to 0.30 mm with a 0.05 mm step. Fig. 4.22 shows the simulated $|S_{11}|$ and $|S_{21}|$ of the model in the frequency range of 50 to 70 GHz with $t_1 = 0.1$ mm and $t_2 = 0.4$ mm. It is evident from the figure that once b is larger than 0.10 mm, low reflection and small insertion loss are achieved.

(a)

(b)

Fig. 4.22: Simulated (a) $|S_{11}|$ and (b) $|S_{21}|$ of the transition between microstrip and stripline with opening b as a parameter.

Fig. 4.23 shows the HFSS model of the vertical via transition between the 50- Ω stripline of the feeding network and the 100 Ω T-junction of the microstrip grid array antenna in a three-layer structure of planar dimensions 10 mm $\times$ 10 mm. The vertical via transition is optimized by fixing a and c at 0.05 mm, l at 2.24 mm, s at 1.12 mm, t_1 and t_2 at 0.4 mm and varying b from 0.10 to 0.30 mm with a 0.05 mm step. Fig. 4.24 shows the

simulated $|S_{11}|$, $|S_{21}|$ and $|S_{31}|$ of the model. It can be seen that reflection at Port 1 decreases as b increases. $|S_{21}|$ improves with b and so does $|S_{31}|$. Once b is larger than 0.10 mm, low reflection and small insertion loss are achieved. A larger b results in lower reflection and smaller insertion loss but may increase the gain penalty due to the more deformed ground plane near the feeding point. Thus, we choose $b = 0.15$ mm for our design.

Fig. 4.23: HFSS Model of the via transition between stripline and microstrip T-junction.

(a)

(b)

(c)

Fig. 4.24: Simulated (a) $|S_{11}|$, (b) $|S_{21}|$ and (c) $|S_{31}|$ of the transition between stripline and microstrip T-junction with the opening b as a parameter.

Fig. 4.25 shows the HFSS model of overall feeding network that divides the single input signal into four output signals to feed the four sub grid arrays. As the feeding network is symmetrical, only transmissions from Port 1 to Ports 2 and 5 are presented. It is seen that phase control is realized by tuning the length of a line section in an output branch. The difference between lengths from Port 1 to Port 5 and from Port 1 to Port 2 is

half guided wavelength at 61 GHz, resulting in 180° phase difference. The characteristic impedances are $35\ \Omega$ for the quarter-wave sections of all T-junctions and $50\ \Omega$ for the other stripline sections.

Fig. 4.25: HFSS Model of the overall feeding network and testing pad of the packaged LP GAA.

Fig. 4.26 shows the simulated $|S_{11}|$, $|S_{21}|$, $|S_{51}|$ and phase difference between S_{51} and S_{21} of the feeding network when b varies. It is seen from Fig. 4.26 (a) that $|S_{11}|$ is below -10 dB from 57 to 64 GHz when b varies from 0.15 to 0.25 mm but deteriorates for $b = 0.10$ and 0.30 mm. Knowing that we achieve a power division ratio of 4 (-6 dB), the insertion loss between Ports 2 and 1 can be estimated from Fig. 4.26 (b) as $0.5 - 1$ dB in the 57 - 64 GHz band, except for $b = 0.10$ mm. The insertion loss between Ports 5 and 1 estimated from Fig. 4.26 (c) is 0.5 dB larger than the insertion loss between Ports 2 and 1 because of a longer transmission path. The phase difference between Ports 5 and 2 is not much affected by the opening size as shown in Fig. 4.26 (d). It shows 200° at 57 GHz, 180° at 61 GHz, and 175° at 64 GHz. The larger phase error at 57 GHz will affect the radiation patterns at this frequency [101].

(a)

(b)

(c)

(d)

Fig. 4.26: Simulated (a) $|S_{11}|$, (b) $|S_{21}|$, (c) $|S_{51}|$ and (d) phase ($S_{21}-S_{51}$) of the feeding network and the testing pad of the packaged LP GAA with opening b as a parameter.

Fig. 4.27 shows the HFSS model of the whole packaged LP GAA. In simulation, a lumped port between the signal trace of the tp and gnd₂ is used to excite the antenna. Radiation boundaries surround the structure with horizontal separations of $\lambda_0/4$ from each edge and vertical separations of λ_0 from the top and bottom surface.

Fig. 4.27: HFSS model of the packaged LP GAA.

Fig. 4.28 shows the simulated $|S_{11}|$ and the peak realized gain of the packaged LP GAA when b varies. It is seen that only $b = 0.15$ mm or 0.20 mm provides the impedance bandwidth $57 - 64$ GHz. For $b = 0.15$ mm, the 3-dB gain BW is from 57 to 66.3 GHz with the maximum gain 16.3 dBi at 60 GHz. For $b = 0.20$ mm, the 3-dB gain BW shifts up to $57.5 - 67.5$ GHz and the maximum gain is decreased to 15.8 dBi at 61 GHz. So $b = 0.15$ mm is an optimum value from both the feeding and radiation considerations.

(a)

(b)

Fig. 4.28: Simulated (a) $|S_{11}|$ and (b) peak realized gain of the packaged LP GAA with the opening b as a parameter.

4.2.3 Measured Performance

The AiP module was fabricated by VTT Finland and tested by LEAT-CNRS, Université Nice Sophia Antipolis. The measurement is limited to 67 GHz. Fig. 4.29 shows photographs of the fabricated packaged LP GAAs with and without signal traces on the bottom surface.

(a)

(b)

(c)

Fig. 4.29: Photographs of the packaged LP GAAs: (a) top view, (b) bottom view and (c) bottom view with signal traces.

Figs. 4.30 (a) and (b) compare the simulated and measured $|S_{11}|$ and peak realized gain of the packaged LP GAA without signal traces on the bottom. Performance of the packaged LP GAA is not affected by the signal traces. It is seen that discrepancies exist between the measured and simulated results. We believe that they are caused by fabrication tolerances and uncertainties in probe contacts and positioning. The measured impedance BW of the antenna is from 50 to 65.6 GHz and the maximum gain is 15 dBi at 62 GHz with 3-dB gain BW from 58.5 to 67 GHz. The measured $|S_{11}|$ and peak realized gain of the AiP module with signal traces features no large difference. Figs. 4.30 (c) to (h) compare the simulated and measured 2D radiation patterns. They reveal that both E- and H-plane patterns have the main beam of radiation in the broadside. The simulated and measured co-polar patterns at 60 GHz agree very well and both have very low back radiations. The maximum simulated realized gain is 9.7 dBi at 55 GHz, 16.3 dBi at 60 GHz, and 12.8 dBi at 67 GHz and occurs at coordinates $(\phi, \theta) = (0^\circ, 90^\circ)$. The maximum measured gain is 8.6 dBi at 55 GHz, 14.6 dBi at 60 GHz, and 12.1 dBi at 67 GHz, and occurs at the same coordinates. For clearer illustration and comparison, Figs.

4.31 (a) to (f) show the simulated and measured 3D radiation patterns of the packaged LP GAA.

(a)

(b)

(c)

(d)

(e)

(f)

Fig. 4.30: Measured performance of the packaged LP GAA: (a) $|S_{11}|$, (b) peak realized gain, normalized 2D radiation patterns at: (c) 55 GHz E-plane, (d) 55 GHz H-plane, (e) 60 GHz E-plane, (f) 60 GHz H-plane, (g) 67 GHz E-plane and (h) 67 GHz H-plane.

Fig. 4.31: 3D radiation patterns of the packaged LP GAA: (a) simulated 55 GHz, (b) measured 55 GHz, (c) simulated 60 GHz, (d) measured 60 GHz, (e) simulated 67 GHz and (f) measured 67 GHz.

4.3 Summary

In this chapter, a LP GAA integrates quadruple sub grid arrays in Ferro A6M LTCC is designed for 60-GHz AiP applications. An emphasis is given to the optimization of via transitions of low-loss array feeding network in the design, which is based on stripline quarter-wave matched T-junctions for the multilayer AiP structure. The effects of the double-end-grounded fences of vias versus the performance of the antenna are examined. The AiP module of dimensions $15 \times 15 \times 1 \text{ mm}^3$ is fabricated and measured. The AiP achieves satisfactory performance: impedance BW from 50 to 65.6 GHz, broadside radiation from 55 to 67 GHz and 3-dB gain BW from 58.5 to 67 GHz with the maximum gain of 15 dBi at 62 GHz. Moreover, the GAA could be converted into single balanced or dual balanced antennas.

Chapter 5 Design of the Circularly-Polarized Grid Array Antenna for 60-GHz Antenna-in-Package Application

Circularly-polarized (CP) antennas find wide applications in wireless communications for the capability to suppress multipath fading and inter-symbol interference. This chapter introduces the design of a CP grid array antenna (GAA) with sub grid arrays, including design of the radiating array and integration of the radiating array with feeding network. Finally, the CP GAA is transformed into a package.

5.1 Design of the 60-GHz Circularly-Polarized Grid Array Antenna

The design methods of CP array antennas can be divided into three categories: (1) each radiating element is designed as a CP radiator [28], (2) linearly-polarized (LP) radiators are orthogonally deposited and fed with equal amplitudes and 90° phase increment, known as sequential rotation [102], and (3) LP radiators with superimposed circularizers [62]-[64]. The complexity of feeding networks for methods (1) and (2) increases with the number of radiating elements [28], which impedes the realization of large arrays. Method (3) was adopted by Kawano and Nakano for CP microstrip (MS) GAA design, in which the large separation between the radiator and circularizer resulted in bulky volume of the antenna. The design method of the CP GAA in this chapter is based on the sequential rotation. Taking the advantage of the GAA that does not require complicated feeding network, adopting four sub grid arrays as the radiating array and

integrating the radiating array with a quarter-wave matched T-junction feeding network, a CP GAA is realized.

5.1.1 Design of the Radiating Array

The CP array antenna is shown in Fig. 5.1. It is formed by four LP sub grid arrays in a 2×2 configuration on the Ferro A6M low temperature co-fired ceramic (LTCC, $\epsilon_r = 5.9$, $\tan \delta = 0.002$) substrate (size $15 \times 15 \times 0.4 \text{ mm}^3$). The length of the long sides of each grid l is twice as the length of the short sides s , $l = 2s = 2.24 \text{ mm}$. The width of the long side w_l equals that of the short sides w_s , $w_l = w_s = 0.12 \text{ mm}$. The LP sub grid arrays are sequentially rotated and excited with 0° , 90° , 180° and 270° phases. Fig. 5.1 (a) shows the simulated current distribution with spacing g between the adjacent sub grid arrays of 0.5 mm . It is clear that the currents on the short sides of the sub grid arrays in quadrants 1 and 3 produce the vertical E-field components, while the currents on the short sides of the sub grid arrays in quadrants 2 and 4 produce the horizontal E-field components. The horizontal and vertical E-field components have the same amplitude and a 90° phase difference, thus generating CP electromagnetic (EM) waves. Figs. 5.2 (a) and (b) show the simulated axial ratio (AR) and peak realized gain of the CP array antenna versus amplitude and phase errors as listed in Table 5.1. It is seen that the 3-dB AR bandwidth (BW) covers 50 - 70 GHz for the case of no amplitude and phase errors. As expected, the AR degrades as the amplitude and phase errors increase. Once the phase error is more than 20° , the AR exceeds 3 dB. However, the gain is not much affected. Simulated radiation patterns of the CP GAA in Case I can be seen in Fig. 4.5 (b) to (g), in which vertical beams in the broadside direction can be detected from 57 to 66 GHz.

Fig. 5.1: Sub grid arrays of the CP GAA: (a) top view with simulated current distribution and (b) bottom view with feeding phases.

(a)

(b)

Fig. 5.2: Simulated (a) AR and (b) peak realized gain of sub grid arrays in the CP GAA with different feeding phases and amplitudes.

Table 5.1: Feeding phases and amplitudes of the sub grid arrays in the CP GAA in Fig. 5.1.

	Phase I ($^{\circ}$)	Phase II ($^{\circ}$)	Phase III ($^{\circ}$)	Phase IV ($^{\circ}$)	Amp. I (dB)	Amp. II (dB)	Amp. III (dB)	Amp. IV (dB)
Case 1	θ	$\theta-90$	$\theta-180$	$\theta-270$	α	α	α	α
Case 2	θ	$\theta-90$	$\theta-180$	$\theta-270$	α	$\alpha-1$	$\alpha-2$	$\alpha-3$
Case 3	θ	$\theta-90$	$\theta-180$	$\theta-270$	α	$\alpha-2$	$\alpha-4$	$\alpha-6$
Case 4	θ	$\theta-100$	$\theta-200$	$\theta-300$	α	α	α	α
Case 5	θ	$\theta-110$	$\theta-220$	$\theta-330$	α	α	α	α

5.1.2 Integration of the Radiating Array with the Feeding Network

Fig.5.3 (a) shows the exploded view of the LP GAA, in which the radiating array of four sub grid arrays is integrated with a quarter-wave matched T-junction feeding network. It has a size of $15 \times 15 \times 0.6 \text{ mm}^3$. The ceramic superstrate (sup) is 0.1 mm thick, which helps preventing the radiating array from being scratched. The ceramic substrates 1 and 2 (sub₁ and sub₂) are 0.4 and 0.1 mm thick, respectively. The gold radiating array (rad), ground plane (gnd) and feeding network (fn) are all 0.01 mm thick. The superstrate does not bring about obvious frequency shift or gain decrement to the antenna for its moderate permittivity and thickness, thus dimensions of the grid (Fig. 5.3 (b)) are kept the same as Fig. 5.1 (a). The sub grid arrays are fed with 90° sequential phase rotation by a quarter-wave matched T-junction feeding network (Fig. 5.3 (c)) on the bottom surface of sub₂. The fn has its input at Port 1 and outputs at Ports 2, 3, 4 and 5. The four output ports are connected to four sub grid arrays in quadrants 1, 2, 3 and 4 respectively by four vias of diameter 0.15 mm and length 0.5 mm through circular openings of diameters 0.3 mm on the gnd. The required phase at a feed point is realized by

controlling the length of the microstrip line. The patterned gnd has many $0.4 \text{ mm} \times 0.4 \text{ mm}$ openings, which help to release different stress from ceramic and metallic layers to avoid warpage and micro fractures. They are deliberately made not directly under the short sides of grids or above the fn to reduce gain penalty and to minimize the detrimental coupling effect between the rad and the fn [97].

Fig. 5.3: The CP GAA: (a) exploded view, (b) radiating array and (c) feeding network.

5.1.3 Parametric Study

Influence of the grid dimensions and gaps between the sub grid arrays on the antenna's performance is analyzed by a parametric study. The initial length of the grid short sides s is half guided wavelength. The initial length of the grid long side l is twice as s . The widths of the grid short and long side w_s and w_l are governed by impedance, transmission and radiation properties [69]. For easy fabrication in large quantity, they are kept the same width with an initial value of 0.12 mm. The separations between the sub grid arrays are also kept the same with an initial value of 0.5 mm. Fig. 5.4 shows the simulated AR, $|S_{11}|$ and peak realized gain as a function of frequency from 50 to 70 GHz with a varying length s and a fixed width $w_s = 0.12$ mm. The 3-dB AR BW is only satisfactory (55.6 – 64 GHz) when $s = 1.12$ mm in Fig. 5.4 (a). It is seen from Fig. 5.4 (b) that both the lower and upper frequency edges are controlled by the length s . They shift upward for a shorter s and downward for a longer s . The impedance BWs all cover the 57 – 64 GHz range when s varying from 1.10 to 1.15 mm. It is evident from Fig. 5.4 (c) that the maximum gain is slightly affected when the length s varying from 1.10 to

1.15 mm. The dependence of the 3-dB gain BW on the length s is similar to that of the impedance BW. It is also the length $s = 1.12$ mm that the 3-dB gain BW is satisfied.

(a)

(b)

(c)

Fig. 5.4: Simulated (a) AR, (b) $|S_{11}|$ and (c) peak realized gain of the CP GAA with length s as a parameter.

Fig. 5.5 shows the simulated CP GAA's performance as a function of frequency with a varying width w_s and a fixed length $s = 1.12$ mm. It is seen that the 3-dB AR BW is much affected by the width w_s varying from 0.10 to 0.15 mm, however both the impedance and 3-dB gain BWs are slightly affected. The width $w_s = 0.12$ mm yields the best performance.

(a)

(b)

(c)

Fig. 5.5: Simulated (a) AR, (b) $|S_{11}|$ and (c) peak realized gain of the CP GAA with width w_s as a parameter.

The gap g between the sub grid arrays is another influential factor. To simplify the design, we make the same separation for the gaps. Fig. 5.6 shows the simulated AR, $|S_{11}|$ and peak realized gain as a function of frequency with a varying separation g and a fixed length $s = 1.12$ mm and a fixed width $w_s = 0.12$ mm. It is evident that the 3-dB AR, impedance and 3-dB gain BWs are affected by the separation g varying from 0.3 to

0.7 mm. The separation $g = 0.5$ mm yields satisfactory performance with the smallest antenna size.

(a)

(b)

(c)

Fig. 5.6: Simulated (a) AR, (b) $|S_{11}|$ and (c) peak realized gain of the CP GAA with separation g as a parameter.

5.2 Packaging of the 60-GHz Circularly-Polarized Grid Array Antenna

In this section, the CP GAA is transformed into a package to meet the requirement for antenna-in-package (AiP) applications. For packaging consideration, a stripline T-junction with delay line phase shifter is used as the feeding network. A metal cross is applied to decouple each sub grid array to control the AR. The formerly applied superstrate is removed. Dimensions of the grid are adjusted.

5.2.1 Design of the Package

The AiP module has dimensions of $15 \text{ mm} \times 15 \text{ mm} \times 0.9 \text{ mm}$. Fig. 5.7 (a) shows its exploded view. It consists of four ceramic and five metallic layers. All the metallic layers are gold of 0.01 mm thick. On top is the radiating array (rad) composed of four

orthogonally deposited sub grid arrays. The formerly used superstrate in the LP GAA in Chapter 4 is removed. A 0.1 mm wide metallic cross is adopted to ensure of the purity of LP radiation from each sub grid array for a better AR performance [49] as shown in Fig. 5.7 (b). The metal cross does not have much influence on the antenna's gain and radiation patterns. Dimensions of the grid are slightly adjusted as $l = 2s = 2.36$ mm, $w_l = w_s = 0.15$ mm. The ceramic substrate layer (sub₁) is 0.4 mm thick. The metallic ground plane (gnd₁) is for the rad and the T-junction stripline feeding network (sfn). The ceramic substrates (sub₂ and sub₃) for the sfn have an equal thickness of 0.2 mm. As shown in Fig. 5.7 (c), the sfn has its input at Port 1 and outputs at Ports 2, 3, 4 and 5. Phase shift is realized by delay line between the input and output ports. The meander section at Port 4 in Fig. 5.3 (c) is redesigned to avoid tight coupling between the two close meander arms in the sfn. The four output ports are connected to four sub grid arrays in quadrants 1, 2, 3 and 4 respectively by four vias of length 0.6 mm through four circular openings on the gnd₁. The fencing vias of length 0.4 mm are located around each feeding via to connect the first and second ground planes. The bottom ceramic layer (sub₄) is 0.1 mm thick. It provides the substrate for the ground-signal-ground (GSG) testing pad or the power, signal and ground traces to be flip-chip bonded with the 60-GHz radio die. The testing pad (tp) shown in Fig. 5.7 (d) is designed the same as in the LP AiP module.

Fig. 5.7: The packaged CP GAA: (a) exploded view, (b) radiating array, (c) feeding network and (d) testing pad.

5.2.2 Optimization of the Package

When packaging the CP GAA in Fig. 5.3 into an antenna-in-package (AiP) module, it is found that the application of superstrate causes tight mutual coupling between sub grid arrays, which brings about great damage to the AR of the CP GAA. Thus the superstrate in Fig. 5.3 is removed and a metal cross is adopted to decouple the sub grid arrays. To

optimize the package, the effect of the metal cross on the CP AiP module's performance is analyzed, and the vertical transition between different signal traces is investigated from the perspective of parallel-plate mode propagation.

A. Optimization of the Metal Cross

The small separation between sub grid arrays in the CP GAA can give rise to mutual coupling, which becomes much severe when the CP GAA is transformed into a package. The coupling can affect the feeding phases and amplitudes from the output ports of the feeding network, which in turn influences the antenna's AR. To decouple the sub grid arrays, a metal cross is applied. The effect of the metal cross can be observed in Fig. 5.8. Compared with the 56.6 – 64 GHz 3-dB AR BW when the metal cross width $w_c = 0.10$ mm, the 3-dB AR BW is quite narrow when there is no metal cross as shown in Fig. 5.8 (a). The 3-dB AR BW also deteriorates with the increased width of the metal cross. The input impedance of the packaged CP GAA is not much affected in Fig. 5.8 (b). Though higher gain and wider 3-dB gain BW is achieved when there is no metal cross in Fig. 5.8 (c), the application of the $w_c = 0.10$ mm metal cross is the optimum considering the 3-dB AR, impedance, 3-dB gain BWs and the gain.

(a)

(b)

(c)

Fig. 5.8: Simulated (a) AR, (b) $|S_{11}|$ and (c) peak realized gain of the packaged CP GAA with metal cross width w_c as a parameter.

B. Optimization of the Via Fence

The transition between the rad, sfn and tp in Fig. 5.7 is realized by conductive feeding vias through openings on gnd_1 and gnd_2 , which is the same as the packaged LP GAA. The feeding via is surrounded by fencing vias. Fig. 5.9 illustrates the simulated electric field distribution between the gnd_1 and gnd_2 for the case of no fence vias at 60 GHz.

Similar with the packaged LP GAA, the undesirable parallel-plate mode is strongly excited, and then guided to the substrate edges where the mode undergoes reflection, radiation and diffraction. Fig. 5.10 shows the simulated field distribution between gnd_1 and gnd_2 for the case of five fences of two vias at 60 GHz, in which a strong parallel-plate mode still exists. The fences of four vias become effective to suppress the parallel-plate mode as evident in Fig. 5.11. The parallel-plate mode is effectively suppressed by the fences of seven vias. Much less field leakage is observed from the gaps between the adjacent vias as shown in Fig. 5.12. The effects of fences of vias on the performance of the AiP module in terms of AR, $|S_{11}|$ and peak realized gain are given in Fig. 5.13. As expected, the fences of seven vias produce the most satisfactory performance.

Fig. 5.9: E-field distribution for the case of no fence of vias in the packaged CP GAA: (a) top view and (b) zoomed side view.

Fig. 5.10: E-field distribution for the case of five fences of two vias in the packaged CP GAA: (a) top view and (b) zoomed side view.

Fig. 5.11: E-field distribution for the case of five fences of four vias in the packaged CP GAA: (a) top view and (b) zoomed side view.

Fig. 5.12: E-field distribution for the case of five fences of seven vias in the packaged CP GAA: (a) top view and (b) zoomed side view.

(a)

(b)

(c)

Fig. 5.13: Simulated (a) AR, (b) $|S_{11}|$ and (c) peak realized gain of the packaged CP GAA with opening b as a parameter.

C. Optimization of the Openings on Ground Planes

Fig. 5.14 shows the HFSS model of overall feeding network that divides the input signal into four output signals to feed the four sub grid arrays. The difference between lengths from Port 1 to Ports 2, 3, 4 and 5 is a quarter guided wavelength at 60 GHz, resulting in 90° phase difference sequentially. The characteristic impedances are 35Ω

for the quarter-wave sections of all T-junctions and $50\ \Omega$ for the other stripline sections. The transition between different layers of signal traces is realized by conductive feeding vias. The radius of the feeding via is a . A circular opening of radius b is set around the transition via on the common grounds gnd_1 and gnd_2 . There are seven fencing vias of radius $c = 0.05\ \text{mm}$ around the feeding via. They are located every 45° at the distance of $b + c + 0.1\ \text{mm}$ around the feeding via to connect gnd_1 and gnd_2 .

Fig. 5.14: HFSS Model of the overall feeding network and testing pad of the packaged CP GAA.

Fig. 5.15 shows the simulated $|S_{11}|$, $|S_{21}|$, $|S_{31}|$, $|S_{41}|$, $|S_{51}|$ and phase difference between output ports of the feeding network when b varies. It is seen from Fig. 5.15 (a) that $|S_{11}|$ is below $-10\ \text{dB}$ from 50 to 70 GHz when b varies from 0.15 to 0.25 mm but deteriorates for $b = 0.10$ and 0.30 mm. Knowing that we achieve a power division ratio of 4 ($-6\ \text{dB}$), the insertion loss between Ports 2 and 1 can be estimated from Fig. 5.15 (b)

as 0.5 – 2.5 dB in the 57 - 64 GHz band, except for $b = 0.10$ mm. The insertion loss between Ports 3 and 1 estimated from Fig. 5.15 (c) is a little larger than the insertion loss between Ports 2 and 1 in the 57- 64 GHz band. The insertion loss turns out to be even larger between Ports 4 and 1 in Fig. 5.15 (d) and improves between Ports 5 and 1 in Fig. 5.15 (e). The fluctuations of phase differences between the output ports of the feeding network improve with the opening size b . Nearly 90° phase difference between adjacent output ports of the feeding network at the center of the 57 – 64 GHz band can be observed from Fig. 5.15 (f) to (h), and the phase difference deteriorates at the upper and lower sides of the band.

(a)

(b)

(c)

(d)

(e)

(f)

(g)

(h)

Fig. 5.15: Simulated (a) $|S_{11}|$, (b) $|S_{21}|$, (c) $|S_{31}|$, (d) $|S_{41}|$, (e) $|S_{51}|$, (f) phase ($S_{21}-S_{31}$), (g) phase ($S_{31}-S_{41}$) and (h) phase ($S_{41}-S_{51}$) of the feeding network and testing pad of the packaged CP GAA with the opening b as a parameter.

Fig. 5.16 shows the simulated AR, $|S_{11}|$ and the peak realized gain of the packaged CP GAA when b varies. It is seen that only $b = 0.15$ mm provides the satisfactory 3-dB AR BW in Fig. 5.16 (a). The impedance BWs are wide enough when $b = 0.15$ or 0.20 mm in Fig. 5.16 (b). The maximum gain and the 3-dB gain BW is not much affected by b in Fig. 5.16 (c). The maximum gain is 15.59 dBi at 60 GHz and 3-dB gain BW 57.5 – 65.5 GHz when $b = 0.15$ mm. So $b = 0.15$ mm is an optimum value from both the feeding and radiation considerations.

(a)

(b)

(c)

Fig. 5.16: Simulated (a) AR, (b) $|S_{11}|$ and (c) peak realized gain of the packaged CP GAA with opening b as a parameter.

5.2.3 Measured Performance

The packaged CP GAA was fabricated by VTT Finland and tested by LEAT-CNRS, Université Nice Sophia Antipolis. The measurement is limited to 67 GHz. Fig. 5.17 shows the photographs of the fabricated packaged CP GAA.

(a)

(b)

Fig. 5.17: Photographs of the fabricated CP GAA: (a) top view and (b) bottom view.

Fig. 5.18 compares the measured and simulated performance of the packaged CP GAA. The measured 3-dB AR BW is 58 – 67 GHz in Fig. 5.18 (a). The measured impedance BW is 56 – 66 GHz in Fig. 5.18 (b). The maximum gain is 14.3 dBi at 62 GHz and the 3-dB gain BW is 59.5 – 67 GHz in Fig. 5.18 (c). Vertical beams in the broadside direction are observed from 57 to 66 GHz in Figs. 5.18 (d) to (i). Frequency shifts in the 3-dB AR, impedance, 3-dB gain and radiation BWs are mainly caused by fabrication inaccuracy. Ferro A6M LTCC material has the dielectric constant and loss tangent varied from 5.7 to 6.1 and 0.002 to 0.0025, respectively. The chosen LTCC process limits the minimum conductor width to 100 μm with a typical tolerance of 5%. The deviation of ceramic layer thickness in a fired LTCC substrate is typically a few percent. The fabrication tolerance analysis reveals that the dielectric constant reduced from 5.9 to 5.7 changes the input impedance and shifts up the 3-dB AR and 3-dB gain BWs. The line width variation from 0.15 mm by $\pm 5\%$ causes a reduction of the maximum simulated peak realized gain.

(a)

(b)

(c)

(d)

(e)

(f)

(g)

Fig. 5.18: Measured performance of the packaged CP GAA: (a) AR, (b) $|S_{11}|$, and (c) peak realized gain, radiation patterns at: (d) 57 GHz $\phi = 0^\circ$, (e) 57 GHz $\phi = 90^\circ$, (f) 60 GHz $\phi = 0^\circ$, (g) 60 GHz $\phi = 90^\circ$, (h) 66 GHz $\phi = 0^\circ$ and (i) 66 GHz $\phi = 90^\circ$

5.3 Summary

In this chapter, a CP GAA using 2×2 LP sub grid arrays in Ferro A6M LTCC for the 60-GHz AiP application is designed. The sub grid arrays are sequentially rotated and excited with 90° phase increment to radiate CP EM waves. The feeding network of the array antenna is based on stripline quarter-wave matched T-junctions. The fabricated AiP module has a size of $15 \times 15 \times 0.9 \text{ mm}^3$. It features good performance: impedance BW from 56 to 66 GHz, broadside radiation from 57 to 66 GHz, 3-dB AR BW from 58 to 67 GHz, 3-dB gain BW from 59.5 to 67 GHz with the maximum gain of 14.3 dBi at 62 GHz.

Chapter 6 Conclusions and Recommendations

6.1 Conclusions

In this thesis, grid array antennas (GAA) for millimeter-wave (mmWave) applications have been studied. It begins with an introduction of the mmWave antenna design, history of GAAs and the low temperature co-fired ceramics (LTCC) in Chapter 1. In Chapter 2, it goes into the fundamentals of planar array antennas. Then the main contributions of the thesis are presented in Chapters 3, 4 and 5, which can be considered as analysis, synthesis, design and packaging.

In the analysis part, GAA's characteristics are studied for the first time at the mmWave band. Characteristics of the GAA like the resonant frequency, impedance bandwidth (BW), half-power beamwidth, gain, 3-dB gain BW and radiation patterns are investigated against the number of radiating elements and substrates' characteristics. The analysis provides a physical insight and helps to form a better understanding of the GAA.

Based on the analysis, a synthesis method is proposed for the first time. Dimensions of the GAA can be calculated directly from the provided closed-form equations after the gain and substrate are specified. The synthesis method is effective in shortening the traditional circle of a GAA design which used to be a trial-and-error process, and also helps to promote the industrial application of GAAs.

In the design part, the problem of current synchronization of GAA with large radiating array is solved by dividing the radiating array into four sub grid arrays and feeding them

simultaneously with a T-junction. The division not only improves current synchronization but also offers more application flexibilities, by which circularly-polarized (CP), single balanced and dual balanced GAAs can be realized besides the linearly-polarized (LP) GAA.

The proposed LP and CP GAAs are finally made into an industrial-ready package. The package is optimized from the view of transitions between different layers of signal traces as well as the parallel mode distribution. Sensitivities of the GAAs to the phase and amplitude errors from the feeding network are also studied, indicating its robust fault tolerant capability which is highly valued by industry.

6.2 Recommendations for Future Research

The thesis analyzes the performance of GAAs on a wide selection of substrates in the 60-GHz band. Two examples of LP and CP GAAs on LTCC substrates with satisfactory performance are designed packaged. Further analysis, design and implementation should be carried out as a complementary extension of the thesis. The recommendations are listed as follows.

First, the problem of unsynchronized current of GAA with large radiating array is not perfectly solved but circumvented by dividing the radiating array into sub grid arrays. Performance of the GAA is improved because better synchronized current from the division, however the design complexity also increases with the division. Subsequent research, which is not presented in this thesis, reveals that unsynchronized currents might arise from the mutual coupling between radiating element. As for array antennas, the mutual coupling usually occurs through: space wave, surface wave and guided wave.

The space wave mutual coupling occurs simultaneously with radiation. It is determined by the radiation mechanism of each radiating element, the separation between each radiating element and the orientation of each radiating element to nearby ones. As a result, in most cases when the antenna's configuration is fixed, hardly anything can be done with the space wave mutual coupling. Surface wave and guided wave mutual coupling are much complicated. Especially in a grounded substrate slab, which is typical of a microstrip structure, EM waves can convert freely between the two propagation modes within certain incident angles. It is possible through complex mathematic procedures, like the method of moments (MoM), to distinguish the roles of mutual coupling through the different modes on the current synchronization of the GAA. The conclusion drawn from the MoM analysis would be very interesting and enlightening. Related measures can be developed accordingly for a totally current synchronized GAA with an undivided large radiating array and a single feed, by which the performance of the GAA will be improved significantly and much broader field of applications will be found.

Second, the division of the large radiating array into sub grid arrays brings application flexibilities, by which balanced GAAs can be realized as well. Chapters 4 and 5 focus on the design of unbalanced LP and CP GAAs. Further efforts are necessary for the design and implementation of the proposed single and dual balanced GAAs.

Third, the LP and CP GAAs are transformed into AiP modules to facilitate the integration with other devices into a radio. In the thesis, they are analyzed as a discrete component in a relatively clean EM ambience. It is believable that when further integration is considered, the antenna will reside in a much complicated EM environment. Interference between the antenna and other devices is supposed to occur which may influence their normal functions. Thus the problem of electromagnetic

compatibility (EMC) in a compact radio containing both passive and active components should be studied for the design of an optimum interface that insures efficient transition or isolation between different parts of the radio. Moreover, because of the low heat conductivity of LTCCs, further integration with active components of the antenna will have to consider the problem of heat radiation, which might require a properly design heat sink on the host board of the packaged radio.

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