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DESIGN AND ANALYSIS OF INTEGRATED MILLIMETER WAVE ANTENNA

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SUMMARY

The 24-GHz industrial, scientific and medical (ISM) band has been allocated by the European Commission and the Federal Communications Commission (FCC) for automotive short range radars applications, such as collision warning system, parking aid, adaptive cruise control, rear traffic crossing alert and blind spot detection. All of these systems require high-gain, low-profile and easily fabricated antennas. In the chip scale, comparing with antenna-on-chip (AOC), antenna-in-package (AIP) is a more promising candidate for 24-GHz single chip system, since AIP has a higher radiation efficiency and wider selection of antenna prototypes and substrate materials.

Among various reported antennas that are used for 24-GHz application, microstrip grid array antenna is a potential candidate because it has all the usual benefits of patch array antenna plus high radiation efficiency, high aperture efficiency and low cross-polarization. Several problems need to be solved to further push microstrip grid array antenna for 24-GHz and millimeter-wave applications. The first is that the mutual coupling mechanism between grid array radiators is unclear. The second is that the polarizations and feeding methods of microstrip grid array antennas are worth exploring. Besides, the radiation performance of large microstrip grid array antenna is limited by non-uniformed current distribution which results in high sidelobes and low gain. Finally, a thorough comparison between microstrip grid array antenna and patch array antennas is needed to further investigate the potential advantages and disadvantages of microstrip grid array antenna.

This thesis firstly presents an analysis of mutual coupling between sub grid arrays on electrically-thin substrate. It is found that the mutual coupling between sub grid arrays on electronically-thin substrate is mainly caused by dielectric polarization currents.

High isolation between sub grid arrays can be achieved by proper selection of the antenna array factor and size. The high isolation between sub grid arrays enlarges design flexibility of the antenna, which makes it easy to implement simultaneous transmission and reception (STAR) application, monopulse antenna and generating orbital angular momentum (OAM) modes. To validate the theory, a particular example of microstrip grid array antenna with 5 sub arrays is developed in RT/duroid 5880 substrate for STAR application with high measured isolation of 65.07 dB between transmitting and receiving antennas. Measured results from this configuration agree well with the simulated results.

Secondly, this thesis presents a multi-port microstrip grid array structure, which gives the flexibility to use it as a single antenna or multiple antennas. To realize such a structure, the key challenges are impedance matching and port isolation. It is shown that impedance matching can be achieved with stepped-impedance lines and port isolation can be enhanced from the defected ground plane. A particular example of 145 radiating elements with 4 ports is implemented for 24-GHz applications. Simulated and measured results are compared, which verifies the design, analysis, and versatility of the structure.

Thirdly, the microstrip grid array antennas excited by slots are designed. A 24 GHz microstrip grid array antenna excited by capacitive coupling slot is designed. Furthermore, a microstrip grid array antenna excited by coaxial-fed slot is proposed. A novel coaxial line to slot transition is introduced.

Fourthly, a dual-polarized microstrip mesh array antenna for 24-GHz applications is developed. The microstrip mesh array antenna enlarges the system polarization diversity with the capability of being utilized to support the simultaneous reception in two orthogonal polarizations, as well as transmission in either polarization. The

proposed design is composed of two grid arrays in two orthogonal directions, where matching stubs are specially designed to compensate the alteration of input impedance due to the direct coupling of the complementary array of another polarization.

Finally, a thorough and fair comparison between microstrip grid array antenna and patch array antenna is carried out. For basic elements, the performances of two-loop microstrip grid array antenna, conventional patch antenna and wideband patch antenna on the same substrate are compared. After that, a 145-element microstrip grid array antenna is compared with a bandwidth-enhanced patch array antenna. Both antennas are designed on single-layer RT/duroid 5880 substrate with size of $100\text{ mm} \times 100\text{ mm} \times 0.25\text{ mm}$. The discrepancies between the performances of these two antennas are analyzed with emphasis on the side lobe level and cross-polarization level.

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Chapter 1. INTRODUCTION

1.1 Background

In this section, background about 24-GHz antennas and history of grid array antenna are briefly introduced to give an overview of this research area.

1.1.1 24-GHz antennas

The 24-GHz ISM band with bandwidth of 250 MHz has drawn considerable attention since it was allocated by the European Commission and the FCC for automotive short range radars applications such as collision warning system, parking aid, adaptive cruise control, rear traffic crossing alert, and blind spot detection [1]-[2]. The small wavelength of 24-GHz band allows circuits and components to be realized in a compact size.

The advancement of silicon IC process opens up new opportunities to integrate the radar system in a single chip. Natarajan, *et al.* demonstrated the first fully integrated phased-array transmitter in a 0.18- μm CMOS in 2005, proving the feasibility of high-frequency integrated phased-array systems on silicon-based processes [3]. Oncu, *et al.* reported a new pulse generator consisting of a series of delay cells and edge combiners with waveform shaping for short-range radar in 90-nm CMOS in 2007 [4], laying the foundation for developing the key technology for one-chip short-range radar sensors. Yu, *et al.* firstly demonstrated an All-RF mm-wave phased array with full amplitude and phase control on each element in 0.13 μm CMOS in 2008 [5]. Cao, *et al.* integrated a 24-GHz transmitter with an on-chip dipole antenna and a synthesizer in 0.13 μm CMOS in 2008 [6]. A highly-integrated dual-band synthesizer prototype chip operating at 24 and 77 GHz was designed and implemented in a 0.18 μm BiCMOS technology by Jain, *et al.* in 2009 [7]. Jain, *et al.* also designed a dual-band transceiver

operating in the 22-29-GHz and 77-81-GHz in 0.18- μ m SiGe BiCMOS in 2009 [8]. All of the 24-GHz short-range radar systems require high-gain, low-profile and easy-fabrication antennas and antenna arrays.

Antennas are integrated with circuit system with antenna-on-chip (AoC) and antenna-in-package (AiP) solutions [9]. The AoC solution features the integration of the antennas or arrays together with front-end circuits on the same chip. However, AoC suffers from low radiation efficiency which is usually lower than 12% because low resistivity of the silicon substrate introduces heating effect, and high permittivity leads to strong surface wave power trapped in the substrate. The AiP solution that combines an antenna or array with a highly integrated radio die into a standard surface mounted chip-scale package device is a good solution to increase the radiation efficiency of the antenna and an important development in the miniaturization of wireless systems. Inter-connection between the radio die and the antenna can be realized by using either wire-bonding or flip-chip techniques. The radiation efficiency of the antenna with AiP solution can reach as high as 80%.

Many antennas and antenna arrays were proposed for 24-GHz short-range radar application. Dipole antennas were widely adopted due to their simple structure. For example, Padhi, *et al.* proposed a CPW fed slot dipole on GaAs substrate achieving gain of 2.1 dBi [10]. To achieve higher gain, Channabasappa, *et al.* designed a bowtie dipole array antenna achieving gain of about 13 dBi [11] and Alhalabi, *et al.* demonstrated an angled-dipole antenna achieving gain of 12.3 dBi and 45° phase scan [12]. However, T-junctions or Wilkinson couplers were used to feed the dipole array introducing additional conductive loss and undesired radiation from the feed network.

Patch antenna is a good candidate for 24-GHz automotive radar application due to their many advantages, such as low cost, low profile, and easy integration with active

circuits. Microstrip patch antennas and arrays were thus reported to apply in automotive radars with good electrical performance [13]–[21]. Shi, *et al.* proposed a patch array antenna with series feed on sapphire ($\epsilon_r=10.2$) wafer [13]. Slovic, *et al.* developed a patch array antenna consisting of two linear arrays of 24 serially connected patches [14]. Wincza, *et al.* demonstrated an electromagnetically coupled corner-series-fed antenna arrays on Arlon CuClad laminate [15]. A 24-GHz switch-antenna array frequency-modulated continuous-wave (FMCW) radar system with a single transmitting patch antenna and four receiving patch antennas was reported by Lee, *et al.* [16]. Han, *et al.* developed a bandwidth-enhanced corporate-fed microstrip array printed on a single-layer electrically-thin substrate [17]. Xu, *et al.* proposed a 24-GHz transmit/receive antenna module with patch antennas fed by a power dividing network [18]. The above mentioned microstrip patch antennas had their radiator and feeding network in the same layer. The radiation from the feed network interfered with the main radiation pattern, which was generally undesirable. To alleviate this effect, feeding networks and radiating antennas were located on different substrates, which avoided the interference of the feeding network with the main radiation pattern. For example, Gruszczynski, *et al.* reported a broadband patch array antenna fed by a parallel feeding network [19]. Mikulasek, *et al.* proposed a 2×2 microstrip patch antenna array fed by a substrate integrated waveguide (SIW) feeding network [20]. Similarly, Rossello, *et al.* proposed a 2×4 array of aperture-coupled microstrip patch antennas fed by a network designed using SIW technology [21]. Slomian, *et al.* developed an 8×4 microstrip patch antenna array with the stacked laminate structure separating the radiating elements and the feeding network [22]. However, these multi-layer structures were bulky, which violated the low-profile requirements of many automotive radar systems.

Many other types of antennas with low profile were developed for 24-GHz automotive radars application. For example, a high-gain compact Yagi–Uda antenna array was implemented with a 2-inch Teflon spherical lens achieving gain of 22 dBi [23]. Another high-gain Yagi-Uda antenna array with two elements was developed on both sides of a Rogers 5880 substrate which yielded a maximum gain of 12.5 dBi [24]. However, this type of antenna was electrically large and had limited gain. Slot antennas were also preferred for 24-GHz automotive radar application [25]–[27]. A typical design of slot array antenna was from Xu, *et al.*, who designed a co-planar waveguide center-fed substrate-integrated-waveguide slot antenna array [27]. The antenna was fabricated onto a single-layer Rogers 5880 and achieved maximum gain of 22.8 dBi with 32×4 slot elements. Waveguide slot array antenna is a good candidate for millimeter wave and automotive radar application [28]–[31] due to its high gain and high efficiency. A 45 linearly polarized hollow-waveguide 16×16 slot array antenna is proposed for point-to-point wireless communication in the 71–76 GHz and 81–86 GHz bands. The antenna array was fabricated by diffusion bonding of thin copper plates and achieved maximum gain of 32.9 dBi with 16×16 elements [30]. Many other antennas were designed for 24-GHz automotive radar application, such as Rotman lens antenna [32]–[33], horn antenna [34], fractal antenna array [35] and so on.

1.1.2 History of grid array antennas

Microstrip grid array antenna is a potential candidate for automotive radar systems with the requirements of high gain, low profile, and easy integration with circuit elements. Grid array antenna was firstly proposed by Kraus in 1964 as a linearly polarized travelling wave antenna, as shown in Fig. 1.1 [36]–[38]. It consists

of rectangular meshes whose long sides operate as transmission lines and short sides function as both transmission lines and radiators. The feed point was at the end of the antenna generating currents in the short sides with a phase progression. The mainbeam direction could be scanned or altered electrically by varying the feed frequency. Thus, the grid array antenna has significant advantage that it is not only a high gain antenna array but also is fed by a single feed point and has compact space volume.

Fig. 1.1. The first grid array antenna proposed by Kraus in 1964: (a) top view, (b) side view [38] and (c) 3D view.

The grid array antenna was firstly implemented in microstrip technology by Conti, *et al.* in 1981, as shown in Fig. 1.2 [39]. They proposed a transmission line model to analyze the radiation performance of the microstrip grid array antenna. It was found that the frequency response of the microstrip grid array antenna is susceptible to changes in effective dielectric constant of the microstrip transmission lines and the geometry of the rectangular loop. Microstrip grid array antenna has a weak cross-polarization due to cancellation of currents on the long sides and as the number of loops increases, the cross-polarization becomes weaker. Microstrip grid array antenna has a good bandwidth which is typically 6%-12% with most stringent bandwidth criteria of antenna gain while the least restrictive of cross polarization quality. Most importantly, Conti, *et al.* proposed a method to control the sidelobe level through adjusting the widths of the short side. A specific example of microstrip grid array antenna with four sub arrays was designed for monopulse radar application.

Fig. 1.2. The microstrip grid array antenna proposed by Conti, *et al.* [39].

Hildebrand provided a detailed numerical analysis of microstrip grid array antennas using the method of moments (MoM) in [40] and [41], as shown in Fig. 1.3. Nakano, *et al.* analyzed many types of microstrip grid array antenna with MOM since 1994 as

shown in Fig. 1.4 [43]-[50]. However, the MOM was based on the assumption that the ground plane of the antenna was infinite, which was not real in practical case. The numerical techniques such as the finite element method (FEM) [51] and finite-differential time domain method (FDTD) [52] etc, and many commercial software packages for full-wave simulation based on these numerical methods were developed in recent years, such as High-Frequency Structure Simulation (HFSS), and CST Microwave Studio and so on. It is more convenient for us to analyze and design the microstrip grid array antennas with these tools. Most of the simulations in this thesis are performed with the HFSS version 14.

Fig. 1.3. Hildebrand's microstrip grid array antenna analyzed with MOM in [41].

Fig. 1.4. Nakano's microstrip grid array antenna analyzed with MOM: (a) microstrip grid array antenna with meander line [43], (b) cross-mesh array antenna [47], and (c) a microstrip grid array antenna [50].

The synthesis of microstrip grid array antenna was firstly attempted by Palmer, *et al.* in 1997 as shown in Fig. 1.5 [53]. They found that it was easier to analyze and design a microstrip grid array antenna by seeing it as an array of microstrip fed patch or dipole elements rather than as a series of loops. The substrate thickness has a strong influence on the bandwidth of the antenna. A $200\ \Omega$ horizontal line impedance was chosen as the optimized value by taking the realizability conditions and the line loss into consideration. Vertical arms set the array radiation coefficients, with the first order approximation of the coefficient being proportional to width.

Fig. 1.5. Microstrip grid array antenna proposed by Palmer, *et al.* [53].

Optimization algorithms were adopted on microstrip grid array antenna to accelerate the antenna design and enhance the adaptability of system complexity [54]-[56]. Chen, *et al.* presented a microstrip grid array antenna using genetic algorithm (GA) achieving -10-dB input impedance and 3-dB gain bandwidth of 25% and 16.3%, respectively, with elliptical radiating elements [55]. Feng, *et al.* developed a high-gain and wide-band microstrip grid array antenna using GA achieving -10-dB input impedance and 3-dB gain bandwidth of 25.6% and 27.6%, respectively, with elliptical radiating elements as well as tapered ground plane [56].

In recent years, Zhang and Sun brought microstrip grid array antenna to a new stage by increasing its working frequency to 60 GHz and integrating it into a chip package [57]-[60]. Dual grid array antennas were designed in a chip package in low temperature co-fired ceramic (LTCC) technology and connected to the radio die through flip-chip bonding as shown in Fig. 1.7 [60]. One of the microstrip grid array antennas was single-endedly fed through ground-signal-ground (GSG) pads, and the other microstrip grid array antenna was differentially fed through ground-signal-ground-signal-ground (GSGSG) pads. Both antennas performed well in the 60-GHz band with maximum gain of 13.5 and 14.5 dBi, respectively.

An analysis and synthesis of microstrip grid array antenna at millimeter wave was described in [61] which gave a physical insight into the radiation behaviors and a comprehensive analysis of this type of antenna. It was found that a thin substrate provided sufficient gain for microstrip grid array antenna at millimeter-wave frequencies. The antenna gain could be enhanced by increasing the number of radiating elements but with a tradeoff between gain and bandwidth. It was found that unsynchronized currents occurred far from the feed point of large grid array and they

would cause large side lobes, low gain and narrow gain bandwidth for the grid array antenna.

(a)

(b)

Fig. 1.6. Wide-band microstrip grid array antennas optimized with GA: (a) microstrip grid array antenna with elliptical radiating elements and meander line in [55], and (b) microstrip grid array antenna with elliptical radiating elements and linearly tapered ground plane in [56].

Fig. 1.7. Integration of dual grid array antennas in a package [60].

A large microstrip grid array antenna with sub grid arrays and multiple feeds was proposed to achieve more synchronized currents as shown in Fig. 1.8 [62]. The sub grid arrays were fed by a single-ended signal source through a feeding network with one input and four output branches. The improved current distribution on grid array radiators was due to fewer loops to be fed by each feed point. Besides, the multiple feed points introduced high design flexibilities which enabled the antenna to be functioned as single balanced antenna and two balanced antennas [63]-[64]. These theoretical analysis and physical realization of microstrip grid array antenna at 60 GHz

paved the way for the applications of microstrip grid array antenna at higher millimeter frequencies such as 79 GHz [65], 122 GHz [66] and 145 GHz [67].

Fig. 1.8. Microstrip grid array antenna with subarrays and multiple feeds: (a) top view, and (b) bottom view [62].

1.2 Research motivation

Although there has been dramatic research progress in the microstrip grid array antenna analysis and design, some limitations are still to be solved to further push microstrip grid array antenna for 24-GHz and millimeter wave applications:

a) The mutual coupling mechanism between sub grid arrays is unclear. As mentioned above, the formation with sub grid arrays and multiple feeds is one of the most effective solutions adopted to improve the current synchronization of the large microstrip grid array antennas, due to the fact that the structure is compact as the original while the design flexibility is greatly enhanced [62]. Since sub grid arrays are used as antenna array elements, it is important to study the mutual coupling between two sub grid array elements. In general, mutual coupling occurs between two sub grid arrays excited by two coaxial probes and it is obtained by measuring S_{21} through the

input and output ports of the sub grid arrays. Although studies of the mutual coupling between microstrip patch antennas have been widely reported [68]-[70], no investigations have been done on the mutual coupling between sub grid arrays. Further studies are therefore needed in order to complete the characterization of this parameter.

b) The current distribution of single-endedly fed large microstrip grid array antenna is non-uniform. Most of the reported microstrip grid array antennas were excited by single feed. The single feed has an advantage of being simple; it however lacks flexibility and limits the use of grid array antenna to a certain extent. It becomes inefficient to excite a large microstrip grid array antenna because the current distribution gets worse. As a result, the antenna aperture efficiency deteriorates, so do the gain and gain bandwidth. To overcome the single-feed drawbacks, Nakano, et al. proposed a multi-feed scheme and showed the improvement in gain and gain bandwidth through numerical simulations of a large microstrip grid array antenna with one or four feeds [71]. However, the proposed antenna could only work when four feeds were excited, which still lacked design flexibilities.

c) The feeding structure of microstrip grid array antenna needs more exploration for millimeter-wave applications. As frequency increases, the feed pins become hard to fabricate and introduce large inductor which is undesired. Frei, *et al.* proposed a 79 GHz microstrip grid array antenna fed by a $100\ \Omega$ differential microstrip line on a single layer of Rogers RO3003 substrate [65]. Mi, *et al.* designed an E-band microstrip grid array antenna fed by a standard WR-10 waveguide with a carefully designed aperture in ground plane [72]. However, as an important excitation method for microstrip patch antenna, slot excitation has not been used for microstrip grid array antenna before. Slot feeding is easy to fabricate and low cost. Hence, it is worth exploring slot excitation on microstrip grid array antennas.

d) Most of the microstrip grid array antennas are designed as single linearly-polarized antenna. As an extension of grid array antenna, dual linearly-polarized mesh array antennas were firstly proposed by Nakano, *et al.* [46]-[47], which are realized in wire and operated below 13 GHz. However, the fabrication of wire grid at millimeter wave would be difficult. Therefore, microstrip line is a better candidate in the millimeter-wave design. Actually, the use of microstrip line would induce a minor orthogonal current component, which is perpendicular to the propagation direction of the microstrip line. The orthogonal current component might decrease the isolation between two polarizations. Moreover, the intersection of short sides introduces direct coupling from the complementary array of another polarization, which alter the input impedances at the input ports. These problems should be solved in microstrip mesh array antenna design.

e) Recent years, microstrip grid array antenna has been used in various applications such as 24 GHz [73]-[74], 60 GHz [57]-[64], 79 GHz [65], 120 GHz [66] and 145 GHz [67] systems, because it has all the usual benefits of the patch array antenna such as compact size, easy fabrication, conformable structure, light weight, and low cost. To further investigate the potential of microstrip grid array antenna, a thorough and fair comparison between microstrip grid array antenna and patch array antenna is necessary.

1.3 Major contributions

To overcome these limitations and solve these problems, this thesis conducts studies as follows.

Firstly, we present an analysis of mutual coupling between sub grid arrays on electrically-thin substrate. It is found that the mutual coupling between sub grid arrays

on electronically-thin substrate is mainly caused by dielectric polarization currents. High isolation between sub grid arrays can be achieved by proper selection of the antenna array factor and size. The high isolation between sub grid arrays enlarges design flexibility of the antenna, which make it easy to implement simultaneous transmission and reception (STAR) application, monopulse antenna and generating orbital angular momentum (OAM) modes.

Secondly, we develop for the first time a multi-port scheme and demonstrate the flexibility to configure the large microstrip grid array structure into one or multiple antennas. We show that the structure can be used for simultaneous transmission and reception. Consequently, the system does not need to use a circulator or a transmit/receive switch or two separated antennas anymore; the system cost and size are reduced. We also show that the structure can be used for multiple-input and multiple-output (MIMO) applications.

Thirdly, microstrip grid array antennas excited by slots are investigated. Coplanar waveguides (CPW) is used to feed the slot with its advantage of easy integration of active or passive components. Two kinds of CPW-fed slot excitations: the inductive coupling slot which is corresponding to a short-circuit slot and the capacitive coupling slot which is corresponding to an open-circuited excitation slot are investigated. According to the electric field distribution in the slot, the coupling mechanism between slot and grid array radiators can be figured out. What's more, the microstrip grid array antenna excited by coaxial-fed slot is also developed. A coaxial line to slot transition is introduced to better excite the slot. The antenna excited by coaxial-fed slot is easier for integration and measurement than that excited by CPW-fed slot.

Besides, single linearly-polarized microstrip grid array is extended to dual linearly-polarized mesh array antenna. The proposed design is composed of two grid arrays in

two orthogonal directions, where matching stubs are specially designed to compensate the alteration of input impedance due to the direct coupling of the complementary array of another polarization. The microstrip mesh array antenna enlarges the system polarization diversity with the capability of being utilized to support the simultaneous reception in two orthogonal polarizations, as well as transmission in either polarization.

Finally, a thorough and fair comparison between microstrip grid array antenna and patch array antenna is carried out for the first time. The radiation performances including impedance bandwidth, peak realized gain, gain bandwidth and radiation pattern are compared. The physical mechanism behind the high gain, low side lobes and low cross-polarization of microstrip grid array antennas are carefully investigated.

1.4 Organization of the thesis

Chapter 1 briefly introduces the reported 24 GHz single-chip radio and antenna-in-packaging technology. Some recent developments and current issues of antennas used for 24-GHz applications are described for comparison purposes. The emphasis is laid on grid array antennas, and their advantages and disadvantages are also reviewed and commented for convenient comparisons with the developed antennas in the latter chapter of this thesis.

Chapter 2 reviews the fundamental theory of surface wave in grounded dielectric substrate, which lays the foundation for the analysis and design of the microstrip grid array antenna in the following chapters. Boundary value method and ray optical method are studied to analyze the surface wave mode. The surface wave modes, cut-off frequency, wave number, incident angle of each mode are evaluated.

Chapter 3 analyzes the mutual coupling between sub grid arrays. We will investigate the mutual coupling between a pair of identical two-loop microstrip grid array

antennas designed at 24.15 GHz, because they are basic models which form various shapes of grid array radiators and can sufficiently represent all the properties of microstrip grid array antennas. In such cases, we analyze the influence of near field coupling, far field coupling and surface wave coupling of the antennas on RT/duroid 5880 substrate. We will present a compact microstrip grid array antenna with sub grid arrays for STAR application, monopulse radar and generating OAM modes. The proposed antenna has the usual benefits of a conventional microstrip grid array antenna plus high isolation between transmitting and receiving antennas, high gain and multi-function.

Chapter 4 presents a multiple port microstrip grid array structure to increase the synchronized currents. We compare the multi-port with the multi-feed schemes on the same large microstrip grid array structure, develop an impedance-matching technique based on stepped-impedance lines, and enhance the isolation between different ports with the defected ground plane. We confirm the feasibility of the multi-port scheme from experiments on a large microstrip grid array structure with 145 radiating elements and four ports at 24 GHz.

Chapter 5 proposes two novel microstrip grid array antennas by slots. The first one is excited by CPW-fed capacitive slot. A comparison between microstrip grid array antenna excited by capacitive and inductive slot is carried out. The second one is excited coaxial-fed slot. A coaxial line to slot transition is firstly introduced. The influence of feed position to the pattern balance is investigated.

Chapter 6 describes a single linearly-polarized microstrip grid array antenna and a dual linearly-polarized microstrip mesh array antenna at 24 GHz. The emphasis is focused on the microstrip mesh array antenna. The effect of the width of the short sides and matching techniques with stubs are investigated. Moreover, it is proved

numerically and experimentally that the proposed antenna can be scaled up to operate at 94-GHz.

Chapter 7 compares thoroughly the microstrip grid array antenna with patch array antenna on the similar substrate in order to give a further understanding of the advantages and disadvantages of microstrip grid array antennas. Their -10-dB impedance bandwidth, peak realized gain, 3-dB gain bandwidth and radiation patterns are all compared in this chapter.

Chapter 8 draws a summary of the research works presented in this thesis and gives recommendations for future work.

Chapter 2. SURFACE WAVE THEORY

The effect of the substrate on printed-circuit antennas is one of the most important factors that should be considered in antenna design. The dielectric substrates of antennas can support surface wave with a fundamental mode with no cutoff frequency. The contribution of surface wave to mutual coupling among array elements cannot be negligible. Many literatures about surface wave's effects on mutual couplings have been published in recent years. Ref. [75] employed reciprocity theorem and integral equation techniques to analyse the effect of surface waves for printed dipole and slot antennas on substrates. It was found that the surface wave excitation reduces the antenna gain of both the printed dipole and slot antennas. In [76], the data on resonant length, input impedance, bandwidth, surface wave power, dielectric loss, and mutual coupling for printed dipoles and microstrip patch antennas on different substrates were presented. It was reported that the surface wave in an array can be reduced by setting particular element spacing and excitations. The mutual coupling properties of printed dipoles arranged along broadside, collinear, and echelon directions were investigated in [77]. It was pointed out that on a thick substrate, printed dipoles have strong mutual coupling in collinear arrangement because of surface wave, and that the surface wave couplings are still significant when the array elements are several wavelengths apart.

2.1 Boundary-value method

In this section, the characteristics of surface waves propagating in the substrate will be analyzed by treating the structure as a boundary-value problem [78]. The numerical expression of TM^z modes and TE^z modes will be derived.

2.1.1 Transverse magnetic (TM) modes

Fig. 2.1. Geometry of a grounded dielectric slab.

Fig. 2.1 shows the geometry of a grounded dielectric slab waveguide. The dielectric slab, of thickness d and relative dielectric constant ϵ_r , is assumed to be of infinite extent in the y and z directions. The Helmholtz wave equation gives:

$$\left(\frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2} + \frac{\partial^2}{\partial z^2} + k^2 \right) E_z = 0. \quad (2.1)$$

The wave propagates in the $+z$ direction with a spatial dependence $e^{-j\beta z}$, and has no variation in the y direction ($\partial/\partial y = 0$).

$$E_z(x, y, z) = e_z(x) e^{-j\beta z} \quad (2.2)$$

So the wave equation (2.1) becomes:

$$\left(\frac{\partial^2}{\partial x^2} + k^2 - \beta^2 \right) e_z(x) = 0 \quad (2.3)$$

The dispersion relations in the substrate and air regions are represented as

$$\begin{aligned} k_c^2 &= \epsilon_r k_0^2 - \beta^2, \\ h^2 &= \beta^2 - k_0^2. \end{aligned} \quad (2.4)$$

where h^2 represents the exponentially decaying field for $x > d$ and β is the wave number in the propagation direction at $x = d$ interface.

For TM modes, applying boundary conditions, we can get the dispersion equation as

$$k_c d \tan(k_c d) = \varepsilon_r h d \quad (2.4)$$

$$(k_c d)^2 + (h d)^2 = (\varepsilon_r - 1)(k_0 d)^2. \quad (2.5)$$

TM₀ mode is the dominant mode of the dielectric slab waveguide with a zero cutoff frequency. The next TM mode, TM₁ mode occurs when the radius of the circle becomes greater than π . The cutoff frequency of the TM_n mode can be calculated by

$$f_c = \frac{nc}{2d\sqrt{\varepsilon_r - 1}}, \quad n = 0, 1, 2, \dots \quad (2.6)$$

2.1.2 Transverse electric (TE) modes

TE modes can also propagate in grounded dielectric slabs.

To satisfy the wave equation and boundary condition, we can get

$$-k_c d \cot(k_c d) = h d \quad (2.7)$$

$$(k_c d)^2 + (h d)^2 = (\varepsilon_r - 1)(k_0 d)^2 \quad (2.8)$$

Cut off frequency for TE modes is

$$f_c = \frac{(2n-1)c}{4d\sqrt{\varepsilon_r - 1}}, \quad n = 1, 2, 3, \dots \quad (2.9)$$

2.2 Ray-tracing method

In Section 2.1, the dielectric slab waveguide was analyzed as a boundary-value problem using modal techniques. In this section, the TM^z and TE^z modes in the dielectric slab waveguide are analyzed again by ray-tracing method, which can shed more physical insight into the propagation of surface wave [79].

Fig. 2.2 shows the side view of the slab waveguide and the coordinate system. It is assumed that the surface wave in the slab waveguide propagates in the z direction and bounce back and forth between its upper and lower interfaces in the x direction. Both the surface wave and structure are uniform in the y direction perpendicular to the xoz -plane.

Fig. 2.2. Side-view of a slab waveguide showing wave normals of the zig-zag waves corresponding to a guided mode.

Fig. 2.3. Reflecting plane wave representation.

Fig. 2.4. Phase wave-fronts.

In Fig. 2.3, the wave vectors A_I and B_I represent a plane wave which bounces back and forth within the slab. The horizontal wave vector components A_{Iz} and B_{Iz} are equal which implies a constant propagation of waves in the z direction. The vertical components A_{Ix} and B_{Ix} travel in the opposite sides with the same magnitude which are combined to form a standing wave. In order to get the same standing wave pattern along the z axis, the plane wave that makes one round trip, up and down across the slab, should have a phase shift equal to $2m\pi$, where m is an integer.

A one round trip phase shift includes both the phase change that is due to the distance traveled by the wave and the changes in the wave phase that are due to reflections from the upper and lower interface. k_d is the phase constant of the plane wave along wave vectors A_I and B_I , and the wave constant in the x direction is $k_d \cos \theta_i$. Therefore, the total phase shift of one round trip (up and down, including the phase changes due to reflection, is

$$2k_d d \cos \theta_i - \phi_1 - \phi_2 = 2m\pi \quad m=0,1,2,\dots \quad (2.10)$$

where ϕ_1 = phase of the reflection coefficient at slab-air interface

ϕ_2 = phase of the reflection coefficient at the ground

Assume the ground is PEC, then $\phi_2 = 0$.

2.2.1 Transverse magnetic (TM) modes

Fig. 2.5. Wave normal of TM mode wave refracted and reflected at the planar interface between air and the slab.

For the wave reflection at the interface between dielectric and the air space as shown in Fig. 2.5, the incident electric field E is related to the reflected electric field E_r by

$$\frac{E_r}{E} = -R = -\left[\frac{\sqrt{\epsilon_r} \cos \theta_r - \cos \theta_i}{\sqrt{\epsilon_r} \cos \theta_r + \cos \theta_i} \right] = \frac{\cos \theta_i - \sqrt{\epsilon_r} \cos \theta_r}{\cos \theta_i + \sqrt{\epsilon_r} \cos \theta_r} \quad (2.11)$$

where R is the reflection coefficient, θ_i is the incident angle and θ_r is the refraction angle.

For an incidence angle θ_i greater than the critical angle θ_c [$\theta_i > \theta_c = \sin^{-1}(\epsilon_r)^{1/2}$], equation (2.11) becomes

$$\begin{aligned} -R &= \frac{\cos \theta_i - \sqrt{\epsilon_r} \cos \theta_r}{\cos \theta_i + \sqrt{\epsilon_r} \cos \theta_r} \\ &= \frac{\cos \theta_i + j\sqrt{\epsilon_r} \sqrt{\epsilon_r \sin^2 \theta_i - 1}}{\cos \theta_i - j\sqrt{\epsilon_r} \sqrt{\epsilon_r \sin^2 \theta_i - 1}} \\ &= |R| \angle \phi = 1 \angle \phi \end{aligned} \quad (2.12)$$

where

$$\phi = 2 \tan^{-1} \left[\frac{\sqrt{\epsilon_r} \sqrt{\epsilon_r \sin^2 \theta_i - 1}}{\cos \theta_i} \right] \quad (2.12)$$

Therefore, the transcendental equation that relates the incident angle and propagation constant is derived

$$\tan(k_d d \cos \theta_i - m\pi) = \frac{\sqrt{\epsilon_r} \sqrt{\epsilon_r \sin^2 \theta_i - 1}}{\cos \theta_i} \quad m = 0, 1, 2 \dots \quad (2.13)$$

$$\text{Define } \phi_{TM} = \tan^{-1} \left(\frac{\sqrt{\epsilon_r} \sqrt{\epsilon_r \sin^2 \theta_i - 1}}{\cos \theta_i} \right) \quad (2.14)$$

Given $h=1\text{cm}$, $\epsilon_r=2.56$, $\mu_r=1$, $f=30\text{GHz}$, the transcendental equation (2.13) can be drawn as Fig. 2.6.

Fig. 2.6. Graphical solution for angles of incidence of TM_a^z modes in a dielectric slab with ground.

The left side of (2.13) is plotted as blue dash line for $m = 0, 2, 4$ in Fig. 2.6, and the intersections of these with that red curve representing the right side of (2.13) are solutions. There are three intersections that occur at $\theta_i = 81.4536^\circ$, 63.7465° and

44.7413° which represent, respectively, the modes TM_0^z , TM_2^z and TM_4^z . Because there is a ground on one side of the slab, TM odd modes do not exist in the slab.

2.2.2 Transverse electric (TE) modes

By a similar procedure to that of TM modes, it is shown that the transcendental equation for TE modes is

$$\tan\left(k_d d \cos\theta_i - \frac{2m-1}{2}\pi\right) = \frac{\sqrt{\varepsilon_r \sin^2 \theta_i - 1}}{\sqrt{\varepsilon_r} \cos\theta_i} \quad m = 0, 1, 2, \dots \quad (2.15)$$

$$\text{Define } \phi_{TE} = \tan^{-1}\left(\frac{\sqrt{\varepsilon_r \sin^2 \theta_i - 1}}{\sqrt{\varepsilon_r} \cos\theta_i}\right) \quad (2.16)$$

Given $h=1\text{cm}$, $\varepsilon_r=2.56$, $\mu_r=1$, $f=30\text{GHz}$, the TM_m^z and TE_m^z modes have been drawn in Fig. 2.7. The intersections of red curve and blue curve represent TM even modes and the intersections of green curve and black curve represent TE odd modes. There are TM modes: TM_0^z ($\theta_i = 81.4536^\circ$), TM_2^z ($\theta_i = 63.7456^\circ$), TM_4^z ($\theta_i = 44.7413^\circ$) and TE modes: TE_1^z ($\theta_i = 73.9605^\circ$), TE_2^z ($\theta_i = 56.815^\circ$) existing in the slab.

Fig. 2.7. Graphical solution for angles of incidence of TM_m^z and TE_m^z modes in a dielectric slab with ground.

2.2.3 The Goos-Hänchen shift

Fig. 2.8. Ray picture of total reflection at the interface between air and substrate.

The Goos-Hänchen shift occurs when the ray is reflected at the dielectric-air interface where the ray behaves differently with the wave normal [80]. When two plane waves are incident at two different angles, the propagation constants in the z direction for two plane waves are $\beta \pm \Delta\beta$. The complex amplitude $A(z)$ of the total incident wave at the interface $x=0$ is

$$\begin{aligned} A &= [\exp(j\Delta\beta z) + \exp(-j\Delta\beta z)] \exp(-j\beta z) \\ &= 2 \cos(\Delta\beta z) \cdot \exp(-j\beta z) \end{aligned} \quad (2.17)$$

where $\beta = k_d \sin \theta$.

With this, the amplitude $B(z)$ of the reflection wave at $x=0$ is

$$\begin{aligned} B &= \{\exp[j(\Delta\beta z - 2\Delta\phi)] + \exp[-j(\Delta\beta z - 2\Delta\phi)]\} \cdot \exp[-j(\beta z - 2\phi)] \\ &= \cos[\Delta\beta(z - 2z_s)] \cdot \exp[-j(\beta z - 2\phi)] \end{aligned} \quad (2.18)$$

where $z_s = \frac{d\phi}{d\beta}$.

From (2.13) and (2.15), we can get

$$\begin{aligned} \text{For TM mode, } k_0 z_s &= (\epsilon_r \sin^2 \theta - 1)^{-1/2} \tan \theta (\epsilon_r \sin^2 \theta - \cos^2 \theta)^{-1} \\ \text{For TE mode, } k_0 z_s &= (\epsilon_r \sin^2 \theta - 1)^{-1/2} \tan \theta \end{aligned} \quad (2.19)$$

In Fig. 11, the depth x_s penetrating into the air region is

$$x_s = z_s / \tan \theta \quad (2.20)$$

2.2.4 Effective waveguide thickness

Effective waveguide thickness d_e is a key factor in all the results. We shall always find that the surface wave field and power are both inversely proportional to d_e . Physically, this means that when the mode is well confined, the surface wave losses are high.

$$\begin{aligned} \text{For TM modes, } x_s &= \left(1 / \sqrt{\beta^2 - k_0^2}\right) \cdot (\varepsilon_r \sin^2 \theta - \cos^2 \theta)^{-1} \\ \text{For TE modes, } x_s &= \left(1 / \sqrt{\beta^2 - k_0^2}\right) \end{aligned} \quad (2.21)$$

For slabs with a ground plane on one side, we can write

$$d_e = d + x_s \quad (2.22)$$

The advantage of using the effective thickness is that we can relate d_e easily to the power per unit width, P , in all these modes by the formula

$$P = EHd_e / 4 \quad (2.23)$$

where E and H are the peak value of the transverse field amplitudes.

2.3 Surface wave in Rogers 5880 substrate at 24 GHz

In this section, the data on surface wave modes propagating at 24 GHz in Rogers 5880 ($\varepsilon_r=2.2$, $\tan\delta=0.0009$) substrates with thickness from 0.381 mm and 3.175 mm will be calculated with Matlab. Fig.2.9 shows the geometry of a grounded dielectric slab waveguide. The dielectric slab of thickness d and relative dielectric constant $\varepsilon_r=2.2$ is assumed to be of infinite extent in the y and z directions. When substrate thickness increases, TM_0 mode firstly occurs. TE_1 mode occurs when substrate

thickness d is larger than 2.85 mm. The propagation order for TM_n and TE_n modes is, $TM_0, TE_1, TM_2, TE_3 \dots$

Fig. 2.9. Side view of the dielectric substrate.

2.3.1 Substrate thickness of 0.381 mm

Only TM_0^z mode is supported in the substrate when the substrate thickness is 0.381 mm. Propagation constants k_c and h can be solved by transcendental equations (2.4) and (2.5), which are plotted in Fig. 2.10. The red line in Fig. 2.10 represents equation (2.4) and the black circle with radius of $\sqrt{\epsilon_r - 1}k_0 d$ represents equation (2.5). The intersection of these curves implies a solution of k_c and h . From Fig. 2.10, we can get $k_c = 505.76 \text{ rad/m}$ and $h = 548.47 \text{ Np/m}$.

Fig. 2.10. Graphical solution for attenuation and phase constants of surface modes when $d = 0.381 \text{ mm}$.

According to the ray-tracing method, the incident angle of TM_0 mode can be solved by transcendental equation (2.13), which is plotted in Fig. 2.11. The dash line represents the right part of equation (2.13) and the blue line represents the left part. The intersection of these two curves implies a solution of equation (2.13). The intersection of black line and blue line implies the solution of TE_0 mode. But in this model, the slab is covered with ground, so TE_0 mode does not exist. From Fig. 2.11, we can get the incident angle of TM_0 mode is 42.68° . It is noted that the critical angle for the substrate-air interface is equal to

$$\theta_c = \sin^{-1} \left(\sqrt{\varepsilon_0 / \varepsilon_r} \right) = \sin^{-1} \left(\sqrt{1/2.2} \right) = 42.392^\circ$$

The values of other parameters are shown in Table 2.1.

Fig. 2.11. Graphical solution for angles of incidence of surface modes when $d = 0.381$ mm.

Table 2.1. Parameters of TM_0 mode when thickness is 0.381 mm.

Parameter	Value
Cutoff frequency (f_c)	0 GHz
Wave number in the air (k_0)	503 rad/m
Wave number in the dielectric (k_d)	746.07 rad/m
Wave number in the z direction (β)	505.76 rad/m
Effective dielectric constant (ε_{eff})	1.011
Wave number in the x direction (k_c)	548.47 rad/m
Attenuation constant in the x direction (h)	52.772 Np/m
Critical angle (θ_c)	42.392°
Angle of incidence (θ_i)	42.68°

2.3.2 Substrate thickness of 3.175 mm

In this case, the substrate has a thickness of 3.175 mm where TE₁ mode occurs. TM₀ mode and TE₁ mode are simultaneously supported in the substrate. The red line in Fig. 2.12 represents equation (2.3). The intersection of dash line and red line implies the solution for k_c and h of TE₁ mode. The intersection of dash line and blue line implies the solution for k_c and h of TM₀ mode. From Fig. 2.12, we can get for TM₀ mode, the wave number in the x direction $k_c = 371.08$ rad/m and attenuation constant in the x direction $h = 407.32$ Np/m. For TE₁ mode, the wave number in the x direction $k_c = 544.22$ rad/m and attenuation constant in the x direction $h = 86.21$ Np/m.

Fig. 2.12. Graphical solution for attenuation and phase constants of surface modes when $d = 3.175$ mm.

In this case, the cutoff frequency of TE₁ mode is 21.5639 GHz. The intersection of red line and black line implies the solution for the incidence angle of TE₁ mode, while the intersection of blue line and dash line implies the solution for the incident angle of TM₀ mode. From Fig. 2.13, we find that the incident angles are 60.17° and 43.16° for TM₀ mode and TE₁ mode, respectively. The values of other parameters of TM₀ mode and TE₁ mode are shown in Table 2.2 and Table 2.3, respectively.

Fig. 2.13. Graphical solution for angles of incidence of surface modes when $d=3.175$ mm.

Table 2.2. Parameters of TM_0 mode when thickness is 3.175 mm.

Parameter	Value
Cutoff frequency (f_c)	0 GHz
Wave number in the air (k_0)	503 rad/m
Wave number in the dielectric (k_d)	746.07 rad/m
Wave number in the z direction (β)	647.23 rad/m
Effective dielectric constant (ϵ_{reff})	1.6557
Wave number in the x direction (k_c)	371.08 rad/m
Attenuation constant in the x direction (h)	407.32 Np/m
Critical angle (θ_c)	42.392°
Angle of incidence (θ_i)	60.1729°

Table 2.3. Parameters of TE_1 mode when thickness is 3.175 mm.

Parameter	Value
Cutoff frequency (f_c)	21.5639 GHz
Wave number in the air (k_0)	503 rad/m
Wave number in the dielectric (k_d)	746.07 rad/m
Wave number in the z direction (β)	510.33 rad/m
Effective dielectric constant (ϵ_{reff})	1.0294
Wave number in the x direction (k_c)	544.22 rad/m
Attenuation constant in the x direction (h)	86.212 Np/m
Critical angle (θ_c)	42.392°
Angle of incidence (θ_i)	43.1604°

2.3.3 Incident angle of surface wave as a function of substrate thickness

Fig.2.14 shows the incident angle changes with substrate thickness. The critical angle of the substrate is $\theta_c = \sin^{-1}(\sqrt{\epsilon_0/\epsilon_r}) = \sin^{-1}(\sqrt{1/2.2}) = 42.392^\circ$. As the

substrate thickness increases from 0mm, the incident angle TM_0 mode starts to increase from critical angle. When the substrate thickness increases to 2.85 mm, TE_1 mode can be supported in the substrate. The incident angle of TE_1 mode also starts from the critical angle and then continues to increase as the substrate thickness increases, but it is always smaller than that of TM_0 mode on the same d . According to the relation among incident angle, k_d and β , as incident angle increases, k_d decreases while β increases.

Fig. 2.14. Incident angle as a function of substrate thickness.

2.3.4 Propagation constant k_c as a function of substrate thickness

The propagation constant in the x direction k_c indicates the field variation in x direction. k_c is often graphically solved by transcendental equations (2.3) (2.4) (2.5) and (2.6). Fig. 2.15 shows that as thickness d increases from 0.127 mm to 6.37 mm, k_c decreases from 550.57 rad/m to 216.33 rad/m for TM_0 mode. When thickness increases to 2.85 mm, TE_1 mode occurs. The k_c of TE_1 mode decreases from 551 rad/m to 375.27 rad/m as substrate thickness d increasing from 2.85 mm to 6.37 mm. Note that the k_c of TE_1 mode is always larger than the k_c of TM_0 mode at the same thickness.

Fig. 2.15. Wave number in the x direction as a function of substrate thickness.

2.3.5 Propagation constant β as a function of substrate thickness

Propagation constant in the z direction β corresponds to the field variation in the z direction. From Fig. 2.16, we can see as substrate thickness increases from 0.127 mm to 6.37 mm, β of TM_0 mode increases from 503.47 rad/m to 714.01 rad/m. When the substrate thickness increases to 2.85 mm, TE_1 can be supported in the substrate and its β begins to increase from 503.1 rad/m. Fig. 2.17 and Fig. 2.18 show the H-field distribution of TM_0 mode excited by printed dipoles when thickness is 1.58 mm and 3.175 mm, respectively. Relatively homogeneous surface wave can be observed in the large distance away from the exciting antenna. Comparing these two figures, we can find that when thickness increases, β increases and then the wavelength decreases, so the distance between two peak decreases.

Fig. 2.16. Wave number in the z direction as a function of substrate thickness.

Fig. 2.17. H-field distribution of TM_0 mode propagating in the substrate with thickness of 1.58 mm and dielectric constant of 2.2.

Fig. 2.18. H-field distribution of TM_0 mode propagating in the substrate with thickness of 3.175 mm and dielectric constant of 2.2.

2.3.6 Attenuation constant in the x direction

The surface wave modes decay with distance from the boundary. In the air space, the field attenuates as e^{-hx} . As d increases from 0.127 mm to 6.37 mm, the attenuation constant increases from 21.866 Np/m to 506.76 Np/m for TM₀ mode and 2.47 Np/m to 403.46 Np/m for TE₁ mode. This attenuation is quite large. For example, when the substrate thickness is 0.127 mm, the TM₀ mode in $8.42\lambda_0$ decays to 10 percent of its value at the surface. When the substrate thickness is 6.37 mm, we should only go to $0.363\lambda_0$ from the surface before the TM₀ mode decays to 10 percent of its value at the surface. The field in the thin substrate is “loosely bound” and the field in thick substrate is “tightly bound”. It is noted that the attenuation constant in the x direction is larger for TM₀ mode than that for TE₁ mode at the same substrate thickness, which implies that the TM₀ mode decays faster in the x direction than TE₁ does.

Fig. 2.19. Attenuation constant as a function of substrate thickness.

2.3.7 Effective guide thickness

The effective waveguide thickness d_e is a key factor in all the results. Fig. 2.20 shows how d_e varies for different thickness and different modes. For the substrate d_e has a minimum where the surface wave power is largest. In this case, d_e of TM₀ mode

has a minimum value when $d = 2.85$ mm. This is where TE₁ mode begins to occur and TM₀ mode has the maximum power. For very thin substrate, d_e becomes very large and surface wave losses go to zero, which also indicates field is “loosely bound” to thin substrate.

Fig. 2.20. Effective guide thickness as a function of substrate thickness.

2.3.8 Surface wave power

In this section, the surface wave power excited by printed dipole is investigated. The normalized TM power is given by

$$P_{TM} = (3\lambda_0 \sin^2 \phi_{TM} \cos^2 \theta_i) / 4d_e \quad (2.24)$$

The normalized TE power is given by

$$P_{TE} = (3\lambda_0 \cos^2 \phi_{TE}) / 4d_e \quad (2.25)$$

$$\text{where } \phi_{TM} = \tan^{-1} \left(\frac{\sqrt{\epsilon_r} \sqrt{\epsilon_r \sin^2 \theta_i - 1}}{\cos \theta_i} \right) \text{ and } \phi_{TE} = \tan^{-1} \left(\frac{\sqrt{\epsilon_r \sin^2 \theta_i - 1}}{\sqrt{\epsilon_r} \cos \theta_i} \right)$$

All the power is normalized to the total power P_0 radiated by the dipole in free space, where

$$P_0 = \omega^2 \mu_0^{3/2} \epsilon_0^{1/2} I^2 l^2 / 12\pi \quad (2.26)$$

As shown in Fig. 2.21, The blue line means the power of TM_0 mode, red line means the power of TE_1 mode and black line means total surface wave power. The results show that, as the substrate thickness increases, the power of individual modes rises and falls, but the total power is a damped oscillation. TE modes reach its maximum value much more quickly than TM modes.

Fig. 2.21. Normalized surface-wave power for printed dipole on a grounded substrate.

2.4 Surface wave modes excited by printed dipoles on Roger 5880 substrates

2.4.1 Model of printed dipole on Rogers 5880 substrate in HFSS

As shown in Fig. 2.22, models of printed dipole operating at 24 GHz on Roger 5880 substrate with thickness of 0.381 mm and 3.175 mm are built in HFSS. The dipole is lying along the z direction and the length of the dipole is modified with different substrate thicknesses to ensure the resonance at 24 GHz. The distance between the substrate edge and the printed dipole is three wavelengths in free space allowing

surface wave to better propagate. Radiation boundaries are attached to the edges and corners of the substrate to reduce reflection and diffraction from the edges and corners.

Fig. 2.23 shows the printed dipole on Rogers 5880 substrate with thickness of 0.381 mm and TM_0 mode is supported in the substrate. The direction of propagation of TM_0 mode is along the direction of dipole. Fig. 2.24 and Fig. 2.25 show E-field distribution and H-field distribution, respectively, of TM_0 mode in 0.381 mm substrate on the xoz -plane.

Fig. 2.26. shows the printed dipole on Rogers 5880 substrate with thickness of 3.175 mm and TE_1 mode is supported in the substrate. Because HFSS is a multi-mode simulation software, the fields we see in 3.175 mm substrate actually are the combined action of TM_0 mode and TE_1 mode, while TE_1 mode power dominates the surface wave power. From Fig. 2.26, we can see the TE_1 mode propagates in the broadside direction of dipole. Fig. 2.27 and Fig. 2.28 show E-field distribution and H-field distribution, respectively, of TE_1 mode in 3.175 mm substrate on the xoz -plane. E-field and H-field distribution of TE_1 mode are also simulated in Matlab. Fig. 2.28 shows the simulated E-field distribution vectors of TE_1 mode in the xoy plane and Fig. 2.30 shows the simulated H-field distribution vectors of TE_1 mode in the xoz plane. The results in Matlab agree well with the results of HFSS. Fig. 2.31 to Fig. 2.32 show the TM_0 mode field distribution in the 3.175 mm substrate. TM_0 mode is still propagating along dipole direction in 3.175 mm substrate. E-field and H-field distribution of TM_0 mode are also simulated in Matlab. Fig. 2.33 shows the simulated E-field distribution vectors of TM_0 mode in the xoz plane and Fig. 2.34 shows the simulated H-field distribution vectors of TM_0 mode in the xoy plane. The results in Matlab agree well with the results of HFSS.

Fig. 2.22. Model of printed dipole operating at 24 GHz on Rogers 5880 substrate.

Fig. 2.23. E-field propagation of TM₀ mode excited by printed dipole on 0.38mm substrate.

Fig. 2.24. E-field of TM_0 mode excited by printed dipole on 0.38mm substrate in the xoz plane.

Fig. 2.25. H-field distribution of TM_0 mode when substrate thickness is 0.38mm in the xoz plane.

Fig. 2.26. E-field propagation of TE_1 mode excited by printed dipole on 3.175 mm substrate.

Fig. 2.27. E-field of TE_1 mode excited by printed dipole on 3.175 mm substrate in the xoy plane.

Fig. 2.28. H-field distribution of TE_1 mode when substrate thickness is 3.175 mm in the xoy plane.

Fig. 2.29. E-field distribution of TE_1 mode when substrate thickness is 0.76mm in the xoz plane.

Fig. 2.30. H-field distribution of TE_1 mode when substrate thickness is 0.76mm in the xoy plane.

Fig. 2.31. E-field of TM_0 mode excited by printed dipole on 3.175 mm substrate in the xoz plane.

Fig. 2.32. H-field of TM_0 mode excited by printed dipole on 3.175 mm substrate in the xoz plane.

Fig. 2.33. E-field distribution of TM_0 mode when substrate thickness is 3.175 mm in the xoz plane.

Fig. 2.34. H-field distribution of TM₀ mode when substrate thickness is 3.175 mm in the xoz plane.

2.4.2 Radiated power by dipole

The reciprocity theorem is applied for calculation of radiated power. The normalized wave impedance with respect to wave impedance in free space is defined as

$$\begin{aligned}\eta_1 &= \frac{\cos \theta_1}{\sqrt{\varepsilon_r} \cos \theta_0} && \text{for TM waves} \\ \eta_2 &= \frac{\cos \theta_0}{\sqrt{\varepsilon_r} \cos \theta_1} && \text{for TE waves}\end{aligned}\quad (2.26)$$

where θ_0 is the angle in the air and θ_1 the refracted angle, as calculated by Snell's law, in the substrate.

$$\begin{aligned}A_i &\equiv \cos^2 \psi + \eta_i^2 \sin^2 \psi \\ B_i &\equiv \cos^2 \psi + \left(\frac{\eta_i + \eta_i^{-1}}{2} \right)^2 \sin^2 \psi\end{aligned}\quad (2.27)$$

where $i=1$ for TE and $i=2$ for TM waves, and ψ is the electrical substrate thickness, given by

$$\psi = 2\pi\sqrt{\varepsilon_r}h\cos\theta_1/\lambda_0 \quad (2.28)$$

The power radiated down through the substrate is

$$P_{dipole}^{down} = \frac{3}{8} \int_0^1 \left[\frac{1}{B_1} + \frac{u^2}{B_2} \right] du \quad (2.29)$$

While the power radiated up into the air is

$$P_{dipole}^{up} = \frac{3}{8} \int_0^1 \left[\frac{A_1}{B_1} + u^2 \frac{A_2}{B_2} \right] du \quad (2.30)$$

where $u = \cos \theta_0$

Fig. 2.35 shows radiated power by dipole on grounded substrate with dielectric constant of 2.2 at 24 GHz and it is observed that the power coupled into surface-wave modes is always larger than the power radiated in the air, which indicates that surface wave power dominates the radiated power. It is also seen that the radiated dipole power exhibits minimum value when the thickness is around a quarter-dielectric wavelength.

Fig. 2.35. Radiated power by dipole on grounded substrate with dielectric constant of 2.2 at 24 GHz as a function of the substrate thickness normalized to the dielectric wavelength.

2.4.3 Surface wave reflection

In Section 2.2, the numerical model is considered to be infinite in the y and z directions. In Section 2.3, radiation boundary is attached on the edge of Rogers 5880 substrate to simulated the infinities in the y and z directions of the substrate. In practice, the substrate of antenna cannot be infinitely large and reflections from edge and corner must be taken into consideration. From the HFSS simulation, we can see that the surface waves in the substrate excited from a printed dipole are cylindrical waves instead of plane waves. The finite grounded dielectric substrate in cylindrical coordinate system is shown in Fig. 2.36. Electrically thin substrate with thickness less than $\lambda_0/10$ is considered in this section.

Fig. 2. 36. Configuration of finite grounded structure in cylindrical coordinate system.

Accounting for reflection, the solution under cylindrical coordinate to Helmholtz equation can be written as

$$\psi_1 = C_1 \cos(k_c x) \cos(n\phi) [H_n^{(2)}(k_\rho \rho) + \Gamma_m H_n^{(1)}(k_\rho \rho)] \quad (2.31)$$

where Γ_m is the reflection coefficient, and $\Gamma_m H_n^{(1)}(k_\rho \rho)$ represents the reflected wave, which arises from the discontinuity of the ground plane.

A closed-form expression for the reflection coefficient is proposed by [81] and [82] using equivalent principle. The reflection coefficient at the substrate edge R_{edge} is given by

$$R_{edge} = (1 - g_{edge}) / (1 + g_{edge}) \quad (2.32)$$

with

$$g_{edge} = \frac{2k_0^2}{3k_\rho} \sum_{m=0}^{\infty} \frac{T_m^2}{\sigma_m} \left/ \left\{ \frac{1}{2\varepsilon_r^2} \left[d + \frac{\sin(2k_c d)}{2k_c} \right] + \frac{\cos^2(k_c d)}{2|h|} \right\} \right. \quad (2.33)$$

where

$$\begin{aligned} T_m = & \frac{1}{k_\rho} \cos(k_c d) \exp(|h|d) \times \left[\frac{k_\rho - h}{k_0} \right]^v + \frac{d}{\varepsilon_r} \left(\frac{k_0 d}{2} \right)^v \bigg/ \Gamma(v+2) \\ & - h \cos(k_c d) \exp(hd) \left(\frac{k_0 d}{2} \right)^v \bigg/ \Gamma(v+2) \times \{1 - |h|d(v+1)/(v+2)\} \end{aligned} \quad (2.34)$$

where $v = 2m/3$ and $\sigma_0 = 2, \sigma_m = 1, m \neq 0$. k_ρ is the wave number in propagation direction, h is the attenuation constant in the x direction, k_c is the wave number in the x direction and k_0 is the wave number in the air.

The coefficient in the field expressions Γ_m at the edge ($\rho = b$) can be calculated using equation

$$\Gamma_m = R_{edge} H_m^{(2)}(k_\rho b) / H_m^{(1)}(k_\rho b) \quad (2.35)$$

Taking reflection into consideration, the TM modes field distribution under cylindrical coordinate can be rewritten as follows

$$E_\rho = -\frac{C_1}{j\omega\varepsilon_0\varepsilon_r} k_c \sin(k_c x) \cos(n\phi) \left[H_m^{(2)}(k_\rho \rho) + \Gamma_m H_m^{(1)}(k_\rho \rho) \right] \quad (2.36)$$

$$E_\phi = \frac{C_1}{j\omega\epsilon_0\epsilon_r\rho} nk_c \sin(k_c x) \sin(n\phi) [H_m^{(2)}(k_\rho\rho) + \Gamma_m H_m^{(1)}(k_\rho\rho)] \quad (2.37)$$

$$E_z = \frac{C_1 k_\rho^2}{j\omega\epsilon_0\epsilon_r} \cos(k_c x) \cos(n\phi) [H_m^{(2)}(k_\rho\rho) + \Gamma_m H_m^{(1)}(k_\rho\rho)] \quad (2.38)$$

$$H_\rho = -\frac{C_1 n}{\rho} \cos(k_c x) \sin(n\phi) [H_m^{(2)}(k_\rho\rho) + \Gamma_m H_m^{(1)}(k_\rho\rho)] \quad (2.39)$$

$$H_\phi = -C_1 \cos(k_c x) \cos(n\phi) [H_m^{(2)}(k_\rho\rho) + \Gamma_m H_m^{(1)}(k_\rho\rho)] \quad (2.40)$$

$$H_z = 0 \quad (2.41)$$

Consider the Rogers 5880 substrates with dielectric constant of 2.2 at $f = 24\text{GHz}$. Fig. 2.37 shows the reflection coefficient as a function of substrate thickness. Only TM_0 mode is supported in the substrate under this condition. As the substrate thickness increases from $0.01\lambda_0$ to $0.085\lambda_0$, the reflection coefficient changes from 0.33 to 0.40, which implies that the reflection wave is comparable to the incident wave, which cannot be neglected. As the substrate thickness increases, the TM_0 mode becomes strong and the reflected coefficient increases.

Fig. 2. 37. Reflection coefficient as a function of normalized substrate thickness.

2.5 Summary

Surface wave modes propagating at 24 GHz in Rogers 5880 substrate were analysed by boundary-value model solution approach and ray-tracing method. Data on propagation constant in transverse direction, propagation constant in propagation direction, attenuation constant, incident angle and effective thickness were presented for different substrate thickness from 0.127 mm to 6.37 mm. The TM_0 mode has no cutoff frequency and second high order mode TE_1 mode occurs when substrate thickness larger than 2.853 mm. It was shown that the propagation constant in transverse direction for TE_1 mode is larger than that for TM_0 mode at the same substrate thickness, while the propagation constant in propagation direction for TE_1 mode is smaller than that for TM_0 mode at the same substrate thickness. The attenuation constant in transverse direction is also an important parameter for surface wave. As substrate thickness increases, the attenuation constant decreases for both TM_0 mode and TE_1 mode, which implies that the field in the thin substrate is “loosely bound” and the field in thick substrate is “tightly bound”. In addition, the attenuation constant in transverse direction for TM_0 mode is larger than that for TE_1 mode at the same substrate thickness, which indicates that TM_0 mode decays faster in transverse direction in the air field than TE_1 does at the same substrate thickness. Furthermore, it is interesting to notice that the surface-wave power is inversely proportional to effective thickness of the substrate. In 24 GHz Rogers 5880 substrate, TM_0 mode achieves both minimal effective thickness of 4.87 mm and maximal power at substrate thickness of 2.85 mm.

The models of printed dipole operating at 24 GHz, lying on Rogers 5880 substrate with thickness of 0.381 mm and 3.175mm, were simulated in HFSS, respectively. It was shown that on the interface between air and the substrate, the TM_0 mode

propagates along the dipole direction and the TE_1 mode propagates in the broadside direction. The side views of field distribution for TM_0 mode and TE_1 mode in HFSS agreed well with that in Matlab. The radiation power by the dipole was also calculated by reciprocity theorem. It was observed that the power coupled into surface-wave modes is always larger than the power radiated in the air, which indicates that surface wave power dominates the radiated power. Finally, reflection of surface wave at the edge was taken into consideration. A numerical expression of reflection coefficient was proposed and the reflection coefficient for the cases of Rogers 5880 substrate with electrically thin thickness is calculated. As the substrate thickness increases, the TM_0 mode becomes strong and the reflected coefficient increases.

Chapter 3. STUDY OF MUTUAL COUPLING BETWEEN SUB GRID ARRAYS ON ELECTRICALLY-THIN SUBSTRATE

This chapter presents an analysis of mutual coupling between sub grid arrays on electrically-thin substrate. The coupling properties of two loop microstrip grid array antennas are investigated. An analysis of the relationship between array factor and mutual coupling is presented. A design of microstrip grid array antenna with sub arrays for STAR application is provided. Finally, a conclusion will be drawn.

3.1 Coupling between two-loop microstrip grid array antenna

In this section, the mutual coupling between a pair of two-loop microstrip grid array antennas is investigated with emphasis on near-field and far-field coupling. The physical source of the radiation and the far-field coupling is explored.

3.1.1 Near-field and far-field coupling

Two-loop microstrip grid array antennas are basic models which form various shapes of grid array radiators and can sufficiently represent all the properties of microstrip grid array antennas. In this section, the level of mutual coupling between a pair of two-loop microstrip grid array antennas is studied as a function of the distance between the feed points. Three distinct two-loop microstrip grid array antenna designs are compared. The first one is the classical antenna whose size of the dielectric substrate is 100 mm $\times$ 100mm. The substrate is RT/duroid 5880 with a thickness of $h = 0.381$ mm, dielectric constant of $\epsilon_r = 2.2$, and loss tangent of $\delta = 0.0009$. For the two-loop grid array radiator, the length and width of the short sides are $s_l = 4.53$ mm and $w_{sl} = 2.5$ mm, respectively, and the length and width of the long sides are $l_l = 9.06$ mm and $w_{ll} = 0.4$ mm, respectively. The antenna is designed at $f = 24.15$ GHz ($\lambda_0 = 12.42$ mm) and is fed by coaxial probe drilled through the ground plane with inner pin

connected to the grid array radiator and outer conductor connected to the ground plane. The input impedance match of the antenna can be modified by displacing the feed point position from the edge of the mesh junction. It is found that, as the displacement increases, the input impedance decreases. The second design is cookie antenna, shown in Fig. 2, where the dielectric exists only underneath the grid array radiator and just 0.1mm around it. The surface wave cannot propagate in this structure, so that it does not have any effect on the mutual coupling and the radiation pattern. Only the near- and far-field couplings are considered between these antennas. For the same resonant frequency, the cookie antenna has a slightly larger dimension due to different edge effects in first and second designs. We can get $s_2 = 4.62$ mm, $w_{s2} = 2.5$ mm, $l_2 = 9.24$ mm, and $w_{l2} = 0.4$ mm. For comparison, the microstrip grid array antenna located in the air at the same height h above the ground plane is investigated as the third design. For the resonant frequency of 24.15 GHz, the dimensions of the third design are $s_3 = 6.25$ mm, $w_{s3} = 3.5$ mm, $l_3 = 12.5$ mm, and $w_{l3} = 0.4$ mm.

Fig. 3.1. Two-loop classical microstrip grid array antenna. (a) Top view. (b) Side view.

Fig. 3.2. Two-loop cookie microstrip grid array antenna. (a) Top view. (b) Side view.

Fig. 3.3. Mutual coupling geometry of two-loop microstrip grid array antenna. When the antennas are (a) orientated in the E-plane. (b) orientated in the H-plane.

Fig. 3.4. Mutual coupling as a function the distance between a pair of the two-loop microstrip grid array antennas at the resonant frequency, when the offset is along (a) E-plane and (b) H-plane.

The coupling at the resonant frequency between two identical two-loop microstrip grid array antennas printed on RT/duroid 5880 substrate and located on the same ground plane is evaluated. The substrate thickness is 0.381mm, where only TM_0 surface wave mode is supported.

When the antennas are orientated in the E-plane, as shown in Fig. 3.4 (a), the $|S_{21}|$ for classic antennas is close to that for cookie antennas indicating that surface wave coupling is not important. Only near field and far field couplings are dominant in classic antenna with substrate thickness of 0.381mm. When the centre-to-centre distance between the two antenna arrays is smaller than $1.3\lambda_0$, the near field coupling dominates (decay of 12 dB per doubled spacing as shown by the theory curve). When the centre-to-centre distance between the two antenna arrays is larger than $1.3\lambda_0$, the far field coupling dominates (decay of 6 dB per doubled spacing as shown by the theory curve), which is due to the strong radiation in the direction of the x -axis. Some ripples are observed on $|S_{21}|$ when the distance between two antennas increases. This is because of the multipath effect of space wave. Some of the space waves are

transmitted in the air and some are transmitted on the boundary between dielectric substrate and air region. The addition or cancellation of the space wave energy from these two paths leads to the ripples of $|S_{21}|$. Eventually, the surface-wave coupling overwhelm far-field coupling and dominates at large distance.

The coupling between the antennas located in the air is not smaller than that for the classic antennas until the distance between the antennas is larger than $0.7\lambda_0$. When the distance between the antennas is larger than $2.5\lambda_0$, far field coupling dominates, and the coupling between two antennas suspended the air is about 20 dB lower than that for classic antennas.

For all three antennas, the decay along H-plane, as shown in Fig. 3.4 (b), is due mainly to the near-field coupling (decay of 12 dB per doubled spacing as shown by the theory curve).

Hence, the mutual coupling effect between two-loop microstrip grid arrays is mainly dominated by near-field coupling when the distance between them is in the range of $(0.5 - 1)\lambda_0$. The far-field coupling is dominant for the two-loop classic microstrip grid array antennas are orientated in the E-plane with a distance larger than $1.3\lambda_0$ and the surface-wave coupling dominates at very large distances.

3.1.2 Radiation pattern and far-field coupling

Fig. 3.5 depicts the gain of the classic and cookie microstrip grid array antenna, as well as the one suspended in the air, which are the same antennas shown in Section II. The gain in the horizontal plane (xoy) is plotted as a function of ϕ , where $\phi = 0^\circ$ corresponds to the direction of x axis. The gain in the vertical planes (xoz and yoz plane) is plotted as a function of θ , where $\theta = 0^\circ$ corresponds to the direction of the z axis.

A prominent radiation is observed for the classic antenna along the horizontal plane. It is found that the gain of the classic antenna in the direction of the x axis is for about 10 dB lower than in the z direction, while the gain in the direction of the y axis is about 30 dB lower than in the z direction. In contrast to this, the gain of the antenna suspended in the air is for about 30 dB lower in the horizontal direction than in the z direction. These results confirm the results shown in Fig. 3.4 that the strong coupling between the classic two-loop microstrip grid array antennas orientated in the E-plane is due to the strong radiation in this direction.

Fig. 3.5. Co-polarized gain (in dBi) of two-loop microstrip grid array antenna in (a) horizontal plane (xoy), (b) E-plane (xoz), and (c) H-plane (yoz).

Fig. 3.6. Qualitative sketch of currents and fields of two-loop classical microstrip grid array antenna.

The strong radiation of two-loop classic microstrip grid array antenna on horizontal plane can be explained by the currents on the grid array and the electric fields under the radiators. Fig. 3.6 sketches the line currents on the grid array and their images which account for the reflection of the ground plane. The current in the protrusion of the inner pin of the coaxial line and its image (i_{feed}), the E-field under the radiator (E) and the polarization vector in the substrate (P) are indicated in Fig. 3.6 as well. The electric field in the far field region only has z component since the tangential electric field is vanishing at the ground plane. The magnetic vector-potential determines the E-field in the far field region, which gives $E_z = -j\omega A_z$. A_z can only be generated by an electrical current source along the z direction. The only current of the antenna along the z direction is the current i_{feed} . For the microstrip grid array antenna located in the air excited with a single coaxial cable, only the current i_{feed} is accounted for the radiation on horizontal plane, generating an omnidirectional radiation pattern. The ripples on the horizontal radiation pattern of the microstrip grid array antenna located in the air are due to the finite ground plane, which causes the reflection and diffraction effect. If the electrical current source of microstrip grid array antenna in the z direction

were eliminated, for example microstrip grid array fed by a microstrip structure, the radiation pattern should have a null in the horizontal plane.

The strong electric field on the horizontal direction of the classic microstrip grid array antenna is mainly caused by the superposition effect of polarization currents in the dielectric. The microstrip grid array radiator can be viewed as an interconnected array of half-of-wavelength dipoles or patches. The length of the short sides of grid array is half of wavelength, so there is half-wavelength resonant field between the short sides and ground plane. The resonant field is close to standing wave with the E-field at both ends changed between maximum to minimum, while the E-field at the middle keeps minimal. In electrically thin substrate, the electric field is mainly perpendicular to the ground plane, except near the grid array edges. Note that these fields have no effect on the antenna suspended in the air. However, when the electric field is applied to the dielectric substrate of the classic microstrip grid array antenna, electric polarization occurs in the dielectric substrate with the polarization vector of $P = (1 - \varepsilon_r) \varepsilon_0 E$. Since the electric field is mainly in the z direction, the polarization vector has predominantly the z-direction component. The polarized dielectric molecules in the dielectric form electric dipoles which is resonant at the same frequency of the electric field, generating polarization currents with current density of $J_p = j\omega P$. The polarization currents are the sources of the strong magnetic vector-potential in the vertical direction A_z .

The polarized current in the substrate beneath the grid array antenna is shown in Fig. 3.7. For a single short side, the electric field in the dielectric is out-of-phased in two halves of the short side, so the polarized currents J_p are also out-of-phased as shown in Fig. 3.7. In the y direction, the radiation fields of these currents cancel out, leading to

weak radiation in the y direction. In the x direction, the radiated fields add up due to the half-of-wavelength distance between them, resulting in a strong radiation. The long sides do not contribute to the radiation due to the out-of-phase current on it.

Fig. 3.7. Current distribution on grid array and polarized current in the dielectric substrate.

The total radiation field of the array is not only determined by the radiation pattern of individual element, but also determined by the array factor.

The normalized array factor for a two-loop grid array in horizontal plane is

$$af_{it} = \frac{1 + 2 \cos(2\pi \sin \phi)}{3} \quad (3.1)$$

with the middle element as the reference point. The array factor shown in Fig. 3.8 has maximal value in the directions of x -axis ($\phi = 0^\circ$) and y -axis ($\phi = 90^\circ$). Since the radiation pattern of an individual element has a maximal value in the direction of x -axis and a null in the direction of y -axis, the total radiation pattern has a maximal value in the direction of x -axis and a null in the direction of y -axis, which agree well with the simulated horizontal radiation pattern shown in Fig. 3.5 (a).

Fig. 3.8. Array factor in horizontal plane of two-loop microstrip grid array antenna.

3.2 Array factor and coupling

Next, we will investigate the effect of sub grid array design on the mutual coupling between sub grid arrays. Sub grid array has an advantage of flexible element arrangement to form different array factor. When the array factor pattern and the radiating element pattern are multiplied together, the grating lobe and null match and cancel the high level of radiating in horizontal plane, leading to a low-level mutual coupling.

For example, we consider four-loop microstrip grid array antenna on the RT/duroid 5880 substrate with thickness of 0.381 mm, a length of 100 mm and a width of 100 mm. As shown in Fig. 3.9, the long sides of the meshes have length of $l = 9.28$ mm and width of $w_l = 0.4$ mm. The short sides of the meshes have length of $s = 4.64$ mm and width of $w_s = 2.5$ mm. The array factor of the four-loop microstrip grid array antenna is calculated as

$$AF = 1 + 2 \cos(2\pi \cos \phi) + 2 \cos\left(\sqrt{2}\pi \cos\left(\frac{\pi}{4} - \phi\right)\right) + 2 \cos\left(\sqrt{2}\pi \cos\left(\frac{3\pi}{4} - \phi\right)\right) \quad (3.2)$$

The normalized array factor is drawn in Fig. 3.10, showing split beam in the direction of x-axis ($\phi = 0^\circ$) and y-axis ($\phi = 90^\circ$), which results in low radiation level in horizontal plane in direction of x-axis and y-axis, as shown in Fig. 3.11 (a).

Fig. 3.9. Configuration of four-loop microstrip grid array antenna.

Fig. 3.10. Normalized array factor of four-loop microstrip grid array antenna in horizontal plane.

Fig. 3.12 shows the mutual coupling in E-plane and H-plane for four-loop microstrip grid array antenna. In the E-plane direction, the near field coupling dominates when the distance between the two antennas is smaller than $2\lambda_0$, and the far field coupling dominates when the distance between two antennas is greater than $2\lambda_0$. In the H-plane, near-field coupling dominates. The $|S_{21}|$ has lowest points at $g = 2.6\lambda_0$, this is due to

cancellation of surface wave and space wave. The mutual coupling between two-loop microstrip grid array antennas is also shown in Fig. 3.12 for comparison. The far-field coupling between two four-loop microstrip grid array antennas orientated in the E-plane is 30 dB smaller than that of two-loop grid array antennas, indicating that the array factor of four-loop microstrip grid array antenna suppresses the far-field radiation along x-axis direction in horizontal plane. The near-field coupling of four-loop microstrip grid array antenna is also significantly suppressed.

Fig. 3.11. Radiation patterns of four loop microstrip grid array antenna. (a) Horizontal plane. (b) E-plane and H-plane.

Fig. 3.12. $|S_{21}|$ of two-loop and four-loop microstrip grid array antenna orientated in the E-plane and H-plane.

The high isolation between sub grid arrays enlarges design flexibility of the whole antenna array, making it easy to implement various applications such as simultaneous transmission and reception (STAR) application, monopulse antenna and generating OAM modes. Dipole array and patch array with a large element spacing are typical structures adopted in these applications [83]-[91]. The larger the element spacing is, the weaker mutual coupling between array elements is and the more concentrated energy is near the beam axis. Large element spacing, however, leads to a bulky volume and low aperture efficiency. Microstrip grid array antenna with sub grid arrays is a good candidate for these applications, since sub grid arrays have high isolation between adjacent sub grid arrays with small clearance between edges and large distance between phase centers, which lead to high aperture efficiency and concentrated radiation pattern in a compact structure.

3.3 Design of microstrip grid array antenna for STAR applications

To demonstrate how the grid sub array can form large antennas for various applications, we provide a design of 24-GHz microstrip grid array antenna with 5 sub grid arrays for STAR application as shown in Fig. 3.13 where the layout and prototype views are provided. In this design, RT/Duroid 5880 is chosen as the substrate with a dielectric constant of $\epsilon_r = 2.2$, a loss tangent of $\delta = 0.0009$, a length of $a = 60$ mm, a width of $b = 60$ mm, and a thickness of $h = 0.381$ mm. The sub grid arrays 1, 2, 3, 4 and 5 on the top surface of the substrate are fed by port 1, 2, 3, 4 and 5, respectively, where sub grid arrays 1 is operated as receiving antenna and sub grid arrays 2, 3, 4 and 5 are operated as transmitting antennas. The receiving antenna consists of 11 radiating elements, which is arranged as an extension of the four-loop microstrip grid array to obtain high gain while weak radiation in the horizontal plane. The long side of the mesh of the receiving antenna has a length of $l_l = 9.4$ mm and width $w_{ll} = 1$ mm. The

short side has a length of $w_l = 4.7$ mm and width $w_{sl} = 2.5$ mm. The normalized radiation pattern of the receiving antenna in the horizontal plane is shown in Fig. 3.14(a). It can be seen that the gain in the horizontal plane is smaller than -20 dB, indicating weak radiation. Four transmitting antennas are designed complementary to the edge of the receiving antenna and are arranged symmetrically and closely around the receiving antenna. The long side of the mesh of the transmitting antenna has a length of $l_2 = 9.6$ mm and width $w_{l2} = 0.4$ mm. The short side has a length of $w_2 = 4.8$ mm and width $w_{s2} = 2.5$ mm. Fig. 3.14(b) shows the normalized radiation pattern of the transmitting antenna in the horizontal plane, which is smaller than -30 dB. The distance between port 2 and 3 is $g = 36.8$ mm and that between port 2 and 5 is $d = 38$ mm.

Fig. 3.15 shows the simulated and measured isolation between single-ended ports the unexcited ports opened. The simulated and measured isolation between port 2 and 3 at 24.15 GHz are 49.24 dB and 46.40 dB, respectively and the simulated and measured isolation between port 2 and 5 at 24.15 GHz are 64.43 dB and 59.96 dB, respectively. The simulated and measured isolation between port 1 and 2 at 24.15 GHz are 32.55 dB and 33.80 dB, respectively, which are sufficiently low for each sub grid arrays to be represented by a point source as in theoretical calculation.

When port 2, 3, 4 and 5 are excited with the same amplitude and with phase of 0° , 180° , 0° and 180° , the currents on the transmitting sub grid arrays are cancelled at the receiving ports leading to high isolation between transmitting and receiving ports. Fig. 3.16 shows the simulated and measured mixed-mode S-parameters of the designed microstrip grid array antenna for STAR application. A Rohde-Schwarz ZVA 67 four-port vector network analyzer and two identical ET industries' 180° hybrids are used in the measurement. Both the transmitting and receiving -10-dB impedance bandwidth

cover the 24-GHz band as shown in Fig. 3.16 (a). Fig. 3.16 (b) shows the simulated and measured isolation between transmitting and receiving ports. The isolation > 62 dB is measured over the operating bandwidth with 65.07 dB at 24.15 GHz. The discrepancy between simulated and measured isolation is mainly caused by the imbalances of the hybrids, the imperfect of the antenna fabrication, soldering difference and measurements nonidealities. To show the influence of the hybrids' imbalance only, the simulated S-parameters of the antenna arrays are linked to the actual hybrid and cable measured S-parameters for comparison. The used hybrids exhibit amplitude and phase imbalances less than ± 0.55 dB and $\pm 10^\circ$, respectively. As seen from Fig. 3.16 (b), the simulated isolation of the ideal system degrades up to 20 dB due to the hybrid imbalances, which indicates the significant impact of phase and amplitude imbalances the isolation between transmitting and receiving ports.

(a)

Fig. 3.13. Configuration of microstrip grid array antenna with 5 sub arrays. (a) Layout; (b) top view; (c) bottom view.

Fig. 3.14. Normalized radiation pattern in the horizontal plane at 24.15 GHz for (a) receiving antenna and (b) transmitting antenna.

Fig. 3.15. Simulated and measured isolation for the microstrip grid array antenna.

(a)

(b)

Fig. 3.16. Simulated and measured s-parameters of microstrip grid array antenna for STAR application. (a) $|S_{11}|$ and $|S_{22}|$; (b) $|S_{21}|$.

To further determine the effect of imbalances of hybrids on the isolation between transmitting and receiving antennas, a computational based sensitivity study is carried out. For simplicity, the amplitude and phase imbalances are assumed to be frequency independent, which is not practical but still can give some understanding of the effect of imbalances on isolation. Fig. 3.17 shows the impact of amplitude and phase imbalances on the isolation performance. The red dash line represents the simulated isolation for the ideal case (perfect hybrids), where a high isolation > 65 dB is achieved over the operating bandwidth. Since the antenna is not symmetrical about H-plane, slight amplitude imbalance of the two hybrids compensates the asymmetric

leading to an increase of the isolation. However, once the phase imbalance of the hybrids is introduced, the isolation is negatively impacted and decreases by 25 dB as seen in Fig. 3.17 (b). It can be clearly seen that phase imbalance has a larger negative effect on the isolation than the amplitude imbalance dose.

(a)

(b)

Fig. 3.17. Isolation performance with (a) ± 0.55 dB amplitude imbalance and 0° phase imbalance, and (b) $\pm 10^\circ$ phase imbalance and 0 dB amplitude imbalance.

Fig. 3.18 (a) and (b) show the peak realized gain for transmission and reception, respectively. The measured maximal peak realized gains are 21.03 dBi for transmission and 15.87 dBi for reception. The simulated maximal peak realized gains are 21.06 dBi for transmission and 15.09 dBi for reception. Fig. 3.19 shows the simulated and measured radiation patterns for transmission at 24.15 GHz. The measured transmitting E-plane radiation pattern has a high main beam in broadside direction which is due to the misalignment of the antenna phase center during the measurement. Both of simulated and measured transmitting radiation patterns have deep nulls in H-plane, leading to high isolation between transmission and reception. However, the measured main beams in H-plane are not symmetrical which may be caused by the unbalanced power distribution of the feeding network. This may also be the cause of the discrepancy between measured isolation and simulated isolation between transmission and reception.

Fig. 3.18. Simulated peak realized gain as a function of frequency of microstrip grid array antenna for STAR application. (a) Transmission; (b) reception.

Fig. 3.20 shows the simulated and measured radiation patterns for reception at 24.15 GHz, who have broadside main beam in both E-plane and H-plane as expected. The total efficiencies calculated from the measured peak realized gain and simulated

directivity are 90.12% and 88.48% at 24.15 GHz for the transmitting and receiving antennas, respectively. The total efficiencies calculated from the simulated peak realized gain and simulated directivity are 97.77% and 93.67% at 24.15 GHz for the transmitting and receiving antennas, respectively.

Fig. 3.19. Simulated and measured radiation patterns at 24.15 GHz of transmitted microstrip grid array antenna for STAR application. (a) E-plane; (b) H-plane.

Fig. 3.20. Simulated and measured radiation patterns at 24.15 GHz of received microstrip grid array antenna for STAR application. (a) E-plane; (b) H-plane.

3.4 Summary

The mutual coupling between sub grid arrays is studied in the chapter. It is demonstrated that the mutual coupling between sub grid arrays on electronically thin substrate is mainly caused by polarized currents in the dielectric substrate and can be reduced by proper array factors of each sub grid arrays. The large microstrip grid array antenna form by sub grid arrays with high isolation has large design flexibility to implement STAR application, monopulse antenna and generating OAM modes. A design example microstrip grid array antenna with 5 sub grid arrays on RT/duroid 5880 substrate for STAR application has been given. The proposed microstirp grid array antenna can achieve high measured isolation of 65.07 dB at 24.15 GHz between transmitting and receiving ports.

Chapter 4. A MULTI-PORT MICROSTRIP GRID ARRAY STRUCTURE

This chapter presents a multi-port microstrip grid array structure, which gives the flexibility to use it as a single antenna or multiple antennas. To realize such a structure, the key challenges are impedance matching and port isolation. It is shown that impedance matching can be made with stepped-impedance lines and port isolation can be enhanced from the defected ground plane. A particular example of 145 radiating elements with 7 ports is implemented in RT/duroid 5880 substrate for 24-GHz applications. Simulated and measured results are compared, which verifies the design, analysis, and versatility of the structure.

4.1 The multi-port microstrip grid array structure

Fig. 4.1 shows the microstrip grid array structure on a Rogers RT Duroid 5880 substrate with a dielectric constant of $\epsilon_r = 2.2$, a loss tangent of $\delta = 0.0009$, a length of $a = 100$ mm, a width of $b = 100$ mm, and a thickness of $h = 0.381$ mm. Note that there are 128 rectangular loops whose 145 long sides act mainly as transmission line elements and whose 145 short sides function essentially as radiating elements. The resonance of the grid array radiator requires the length

$$l = \frac{\lambda}{\sqrt{\epsilon_{eff}}} \quad (4.1)$$

where ϵ_{eff} is the effective dielectric constant that can be estimated as

$$\epsilon_{eff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left[1 + 12 \frac{h}{w_l} \right]^{-\frac{1}{2}} + 0.02(\epsilon_r - 1) \left[1 - \frac{w_l}{h} \right]^2 \quad (4.2)$$

Considering realizability from fabrication and reliability for use in automobiles, we prefer $Z_c = 125 \Omega$. Substituting (4.2) into (4.1), we get $w_l \approx 0.4$ mm, which leads to $\epsilon_{eff} \approx 1.7$ from (4.1) and $l \approx 9.6$ mm from (4.1).

Similarly, the resonance of the grid array radiator also requires the length s to be half of the length l , that is, $s \approx 4.9$ mm. It is more involved to determine the widths w_s because the width w_s sets the array radiation coefficients with the first order approximation of the coefficient being proportional to width [53]. For simplicity, we assign the same radiation coefficient for all radiating elements, leading to the same width w_s for all short sides. To enhance the radiation of co-polar components, the choice of w_s should be

$$w_s \leq \frac{\lambda}{4\sqrt{\epsilon_{eff}}} \approx 2.4 \text{ mm} \quad (4.3)$$

After that, we layout as many rectangular meshes as possible on the substrate to form the grid array radiator to achieve the directivity. There are 128 meshes in this example, so the number of short sides (or radiating elements) is 145. The directivity is estimated by

$$D = 5.16 + 10 \log 145 = 26.77 \text{ dB} \quad (4.4)$$

where 5.16 dB accounts for the directivity due to the ground plane and one radiating element. It indicates that the directivity > 25 dB is achievable. The simulations are conducted mainly to yield the optimized length l and width w_l for the long sides and length s and width w_s for the short sides of rectangular meshes. They are found to be $l = 9.68$ mm, $w_l = 0.4$ mm, $s = 4.84$ mm, and $w_s = 1.5$ mm, which are quite close to our calculated values. In what follows, we analyze the feeding, matching and isolation associated with the structure.

Fig. 4.1. Illustration of the microstrip grid array structure (F_1 and P_2 are at the same point).

4.1.1 Multi-port versus multi-feed schemes

For the large microstrip grid array structure shown in Fig. 4.1, Nakano et al. proposed the multi-feed scheme to improve the gain and gain bandwidth [71]. They suggested exciting the structure at feeds F_1 , F_2 , F_3 and F_4 . To achieve the synchronized currents on the radiating elements, the distance between the two feeds on the same principal plane should be an even multiple of $\lambda_g/2$ and all the feeds must be excited with the same amplitude and phase.

We propose the multiple-port scheme for the flexibility to configure the large microstrip grid array structure into one or multiple antennas. We suggest exciting the structure at ports P_1 , P_2 , P_3 and P_4 . The distance between every two ports should be an odd multiple of $\lambda_g/2$ and all the ports are excited with the same amplitude but paired phases as 180° , 0° , 180° and 0° for broadside radiation.

Fig. 4.2 compares the simulated radiation performances of the large microstrip grid array structure excited with the multi-feed and multi-port schemes. Note that the multi-

port scheme results in a higher peak realized gain, a wider 3-dB gain bandwidth, a lower cross polarization component in the H plane, and a higher radiation efficiency than the multi-feed scheme does because the multi-port scheme enjoys the differential advantage that helps to suppress surface wave. Also the multi-port scheme implies that the ports are connected directly to sources, while the multi-feed scheme requires a power divider to connect the feeds to a source. The insertion loss of the power divider reduces the gain of the structure.

(a)

(b)

Fig. 4.2. Simulated radiation performances: (a) peak realized gain, (b) radiation efficiency, (c) E-plane patterns at 24.15 GHz, and (d) H-plane patterns at 24.15 GHz.

4.1.2 Impedance-matching technique

The microstrip grid array structure is an interconnected array concept. The input impedance at a port of the microstrip grid array structure for the first-order approximation can be expressed as R/n where R is the radiation resistance of a radiating element and n is the number of radiating elements. Obviously, as n increases, the input impedance may decrease to a value much smaller than the source impedance, leading to a poor matching. To improve impedance matching, an effective technique based on stepped-impedance lines as shown in Fig. 4.3 is devised as follows.

Note that the microstrip grid array structure is symmetrical about the E plane. One only needs to consider the half structure. The impedance at port one is Z_{in} . The impedance at node 2 is Z_L , which is equal to Z_{in} because the microstrip grid array structure is resonant; all the node impedance must be essentially identical as they are all a half wavelength apart in the simple transmission line model.

Fig. 4.3. Microstrip grid array structure model with a stepped-impedance line.

A stepped-impedance line is used to match Z_L to Z_s . The impedance Z_s should be 2 times of the source impedance of 50Ω . The stepped-impedance line consists of two sections. For simplicity, we assign the width of the first section to be the same as the width of the long side of the loop. Hence, its characteristic impedance Z_l is known. The width of the second section should be less than $\lambda_g/4$ to suppress higher-order modes. The characteristic impedance of the second section is Z_2 , which should be smaller than Z_l . The lengths of the first and second sections are denoted as l_l and l_2 , respectively, and $l_l + l_2 = l$.

According to the theory of small reflection, the overall reflection coefficient

$$\Gamma \approx \Gamma_0 + \Gamma_1 e^{-2j\theta_1} + \Gamma_2 e^{-4j\theta_2} \quad (4.5)$$

where

$$\Gamma_0 = \frac{Z_l - Z_s}{Z_l + Z_s} \quad (4.6)$$

$$\Gamma_1 = \frac{Z_2 - Z_l}{Z_2 + Z_l} \quad (4.7)$$

$$\Gamma_2 = \frac{Z_L - Z_2}{Z_L + Z_2} \quad (4.8)$$

$$\theta_i = \frac{2\pi}{\lambda_g} l_i \quad (i=1,2) \quad (4.9)$$

The tapered line is set in one loop of the grid array radiator, so

$$\theta_1 + \theta_2 = 2\pi \quad (4.10)$$

$$0 \leq \theta_1 \leq 2\pi \quad (4.11)$$

$$0 \leq \theta_2 \leq 2\pi \quad (4.12)$$

If the width of the tapered line w_2 is larger than $\lambda_g/4$, there exists high order modes in the tapered line, lead to degradation of the radiation performance. Therefore, we choose $40 \leq Z_2$.

Fig. 4.4 shows the calculated reflection coefficient seen by port 1 with different Z_2 and l_2 . Note that the impedance matching can be achieved when Z_2 changes from 40Ω to 70Ω and l_2 are at discrete values. $Z_2 = 50 \Omega$ or $w_2 = 1.2 \text{ mm}$ and $l_2 = 2 \text{ mm}$ are chosen as the initial values for optimizing with the HFSS^{TM} simulator.

Fig. 4.4. Total reflection coefficient of the microstrip grid array model with a stepped-impedance line at 24.15 GHz.

Fig. 4.5 shows the simulated $|S_{11}|$ of the microstrip grid array structure excited at port 1 versus l_2 with a fixed $w_2 = 1.2$ mm. Note that $l_2 = 2.5$ mm satisfies the impedance matching and bandwidth requirements. Fig. 4.6 shows the simulated $|S_{11}|$ of the microstrip grid array structure excited at port 1 versus w_2 with a fixed $l_2 = 2.5$ mm. Note that as the w_2 increases from 0.4 to 1 mm, the $|S_{11}|$ value drops from -7 dB to -11 dB, the resonant frequency keeps unchanged. Consequently, in the optimization of the stepped impedance line, the resonant frequency can be adjusted with the length of l_2 and the matching is controlled by the width of w_2 . It should be pointed out that, due to the interconnection between ports in the proposed structure, adding a stepped-impedance line to a port affects the input impedances of other ports. However, it is found that this effect is not critical which can be alleviated by slightly modifying the dimensions of the stepped-impedance line.

Fig. 4.5. Simulated $|S_{11}|$ of the microstrip grid array structure excited at port 1 versus l_2 .

Fig. 4.6. Simulated $|S_{11}|$ of the microstrip grid array structure excited at port 1 versus w_2 .

4.1.3 Port isolation enhancement

For the multiple-port scheme, isolations between different ports are of great importance. High isolation between ports enables the multi-port microstrip grid array structure to be used as transmitting and receiving antennas simultaneously. This is in contrast to the multiple-feed scheme where isolations between different feeds are not a concern.

Fig. 4.7 shows the simulated single-ended S-parameters when port 1 and port 4 are excited and the other ports are opened. Note from $|S_{11}|$ and $|S_{44}|$ that both ports 1 and 4 are matched well. The isolation between ports 1 and 4 given by $-|S_{41}|$ is 19.19 dB at 24.15 GHz.

Fig. 4.8 shows the simulated differential-mode S-parameters when ports 1 and 2 are configured as balanced port 1 and ports 3 and 4 as balanced port 2. Note from $|S_{dd11}|$ and $|S_{dd22}|$ that both balanced ports are matched well. The isolation between the two balanced ports given by $-|S_{dd21}|$ is 17.18 dB at 24.15 GHz.

Fig. 4.7. Simulated single-ended S-parameters of the microstrip grid array structure when port 1 and port 4 are excited and port 2 and port 3 are opened.

Fig. 4.8. Simulated differential-mode S-parameters of the microstrip grid array structure when port 1 and port 2 are configured as balanced port 1 and port 3 and port 4 as balanced port 2.

There have been techniques to enhance isolation between antennas [92]. Little has been done to enhance isolation between ports of an antenna. It is found that the isolation between different ports can be greatly enhanced with the defected ground

plane. We cut 6 slots on the ground plane under the long sides of the loops near the geometrical center of the microstrip grid array structure. Each slot has a length of $\lambda_g = 9.68$ mm and a width of 2.94 mm. Our simulations show that the defected ground plane does not degrade the radiation performance but enhances the isolation significantly, for instance, the isolation between two single-ended ports as shown in Fig. 4.7 improves to 33.69 dB and between the two balanced ports as shown in Fig. 4.8 improves to 32.80 dB at 24.15 GHz. The current distributions of the multi-port microstrip grid array structure at 24.15 GHz on the ground plane and on the defected ground plane are displayed in Fig. 4.9 (a) and (b), respectively. It is shown that the defected ground plane weakens the coupling current between the ports, as a result, the isolation is enhanced.

(a)

Fig. 4.9. Current distribution at 24.15 GHz for multi-port microstrip grid array structure (a) without slot and (b) with slot.

4.2 Results and discussion

A particular example of 145 radiating elements with 4 ports is implemented in a RT/duroid 5880 substrate with a size of $100 \times 100 \times 0.381$ mm³ for 24-GHz applications. Fig. 4.10 shows the photographs of a fabricated sample. It is seen that a metallic plate is attached with the ground plane by four plastic screws to flatten the structure and four 2.92-mm SMA connectors are soldered to the four ports for measurements. The S -parameter measurements of the fabricated structure were carried out with a Rohde-Schwarz four-port vector network analyser. The radiation patterns were measured in an anechoic chamber. The gain was obtained with the comparison method. A comprehensive measurement campaign was conducted to fully characterize the multi-port microstrip grid array structure as a single antenna or multiple antennas. An

acceptable agreement has been achieved between all the measured and simulated results. In the following the results for the structure as a single antenna are briefly mentioned and as two antennas are discussed.

(a)

(b)

Fig. 4.10. Photographs of the fabricated multi-port microstrip grid array structure with 4 ports: (a) Top view and (b) Bottom view.

Fig. 4.11 shows the simulated and measured results for the structure excited at port 2 as an unbalanced antenna. The simulated and measured 10-dB input impedance bandwidths are 0.38 and 0.33 GHz, respectively. The measured maximum gain is 23.26 dBi at 24.18 GHz with the 3-dB gain bandwidth of 0.80 GHz from 23.74 to 24.54 GHz. The simulated maximum gain is 23.58 dBi at 24.15 GHz with the 3-dB gain bandwidth of 2.52 GHz from 22 to 24.52 GHz.

The total efficiency calculated from the measured peak realized gain and simulated directivity is 86.04% at 24.15 GHz and the total efficiency calculated from the simulated peak realized gain and simulated directivity is 91.91% at 24.15 GHz.

(a)

(b)

Fig. 4.11. Simulated and measured results for the structure excited at port 2: (a) $|S_{11}|$, (b) peak realized gain, (c) simulated radiation pattern at 24.15 GHz in E-plane, and (d) simulated radiation pattern at 24.15 GHz in H-plane.

Fig. 4.12 shows the simulated and measured results for the structure excited at ports 2 and 3 as a balanced antenna. The simulated and measured 10-dB input impedance bandwidths are 0.33 and 0.30 GHz, respectively. The measured maximum gain is 26.84 dBi at 24.17 GHz with the 3-dB gain bandwidth of 0.70 GHz from 23.82 to 24.52 GHz. The simulated maximum gain is 26.39 dBi at 24.15 GHz with the 3-dB gain bandwidth of 0.70 GHz from 23.80 GHz to 24.50 GHz.

For the structure, the large physical distance between ports 1 and 4 guarantees a high isolation between the two ports, which enables the structure to be used for simultaneous transmission and reception as well as for MIMO application. Fig. 4.13 shows the simulated and measured results when the structure is transmitting at port 1 and receiving at port 4. When the transmitting radiation performances are measured, the receiving port is terminated with 50 Ω load and vice versa. The measured -10-dB impedance bandwidth is 0.24 GHz from 24.17 to 24.41 GHz and 0.37 GHz from 24.12 GHz to 24.49 GHz for transmission and reception, respectively. The measured isolation between two ports (S_{14}) at 24.15 GHz is 34.24 dB, which

is about 10-dB higher than the typical isolation level of a waveguide-based circulator or a CMOS-based transmit/receive switch. The simulated and measured peak realized gains at transmitting port 1 and receiving port 4 have small differences in the 24-GHz band from 24.05 to 24.25 GHz. Thus, together with the adaptive leakage cancellation in the receiver, we believe that the structure can function as two unbalanced antennas with one for transmission and the other for reception for frequency modulated continuous wave radar and sensor applications in the band [92], [93]. The main beams of radiation in both simulated and measured E planes are deviated 2° to the left or right from the broadside direction. This is because the phase centers of the structure are near the ports rather than the geometrical center of the structure. As expected, the main beams of radiation in both simulated and measured H planes are in the broadside direction. The total efficiencies calculated from the measured peak realized gain and simulated directivity are 94.07% and 79.63% at 24.15 GHz for the transmitting and receiving antennas, respectively. The total efficiency calculated from the simulated peak realized gain and simulated directivity is 93.63% at 24.15 GHz for both transmitting and receiving antennas.

(a)

(b)

(c)

(d)

Fig. 4.12. Simulated and measured results for the structure excited at ports 2 and 3 as a balanced antenna: (a) $|S_{11}|$, (b) peak realized gain, (c) simulated radiation pattern at 24.15 GHz in E-plane, and (d) simulated radiation pattern at 24.15 GHz in H-plane.

(a)

(b)

(c)

(d)

Fig. 4.13. Simulated and measured results for the structure transmitting at port 1 and receiving at port 4: (a) S-parameters, (b) peak realized gain, (c) transmitting radiation pattern at 24.15 GHz in E-plane, (d) transmitting radiation pattern at 24.15 GHz in H-plane, (e) receiving radiation pattern at 24.15 GHz in E-plane and (f) receiving radiation pattern at 24.15 GHz in H-plane.

Popular metrics for MIMO antennas are the ratio of the mean effective gains and the envelope correlation coefficient. Mean effective gain of one antenna is defined as the ratio of the mean received to the mean incident power at that antenna. Envelope correlation coefficient describes the correlation between the radiation patterns of antennas in the MIMO system. Usually, the ratio should be smaller than 3 dB and the envelope correlation coefficient should be smaller than 0.5 to ensure good channel characteristics [94], [95]. The ratio of the mean effective gains for the structure at ports 1 and 4 is shown in Fig. 4.14(a), which is smaller than 1 dB in the 24-GHz band. The envelope correlation coefficient for the structure at ports 1 and 4 is shown in Fig. 4.14(b), which is smaller than -60 dB in the 24-GHz band.

For the structure, ports 1 and 2 can be configured as one balanced antenna for transmission and ports 3 and 4 as another balanced antenna for reception or as balanced antennas for a 2×2 MIMO system. Fig. 4.15 shows the simulated and

measured results. The simulated impedance bandwidths for both transmitting and receiving antennas are 0.40 GHz from 24.15 GHz to 24.55 GHz. The measured impedance bandwidths for the transmitting and receiving antennas are 0.52 GHz from 24.07 GHz to 24.59 GHz and 0.64 GHz from 24.12 GHz to 24.75 GHz, respectively. The simulated and measured isolations between balanced ports 1 and 2 at 24.15 GHz are 32.80 dB and 34.13 dB, respectively. The simulated and measured gains are different, due to the loads connected to the antenna are not exactly $50\ \Omega$ in the 24-GHz band. For both antennas, the desired broadside radiation patterns and acceptable cross-polarization components are observed. The total efficiencies calculated from the measured peak realized gain and simulated directivity are 83.61% and 83.67% at 24.15 GHz for the transmitting and receiving antennas, respectively. The total efficiency calculated from the simulated peak realized gain and simulated directivity is 92.09% at 24.15 GHz for both transmitting and receiving antennas.

(a)

Fig. 4.14. Simulated and measured results for the structure as 1×2 unbalanced MIMO antennas: (a) magnitude of ratios of mean effective gains and (b) envelope correlation.

Fig. 4.15. Simulated and measured results for the structure transmitting at balanced port 1 and receiving at balanced port 2: (a) S-parameters, (b) peak realized gain, (c) transmitting radiation pattern at 24.15 GHz in E-plane, (d) transmitting radiation pattern at 24.15 GHz in H-plane, (e) receiving radiation pattern at 24.15 GHz in E-plane and (f) receiving radiation pattern at 24.15 GHz in H-plane.

The ratio of the mean effective gains for the structure at balanced port 1 and 2 is shown in Fig. 4.16(a), which is smaller than 2.5 dB. The envelope correlation coefficient for the structure at balanced port 1 and 2 is shown in Fig. 4.16(b), which is smaller than -50 dB in the 24-GHz band. The difference between measured and simulated results is because the irregular solder joints lead to a slight difference

between two balanced ports. But the measured results still satisfy the criteria in [96], [97], indicating low correlation and sufficient power balance between two balanced ports.

(a)

(b)

Fig. 4.16. Simulated and measured results for the structure as 1×2 balanced MIMO antennas: (a) magnitude of ratios of mean effective gains and (b) envelope correlation.

4.3 Summary

A novel high-gain multi-port microstrip grid array structure of 145 elements has been proposed with large design flexibility, which enables the proposed structure easy to implement one or two unbalanced antennas as well as one or two balanced antennas. Stepped-impedance lines are specially designed on the long sides of the meshes near the feeding ports to achieve good matching. Isolations between different ports have been studied and a simple technique of cutting slots in the ground plane to perturb the electromagnetic coupling between ports so as to increase the isolation has been simulated. It has been shown that the slot in the ground plane can dramatically improve isolation without degrading the radiation performance of the antenna. A design example multi-port microstrip grid array structure with 4 ports on RT/duroid 5880 substrate has been given. The designed multi-port microstrip grid array structure has excellent performances when it is operated as one or two unbalanced antenna as well as one or two balanced antenna. More interestingly, the novel multi-port microstrip grid array structure has the unique capability to function as MIMO antenna with envelope correlations between radiation patterns of the two antenna elements lower than -40 dB over the 10-dB input impedance bandwidth of the proposed structure.

Chapter 5. MICROSTRIP GRID ARRAY ANTENNA EXCITED BY SLOT

The design of microstrip grid array antennas excited by slot are introduced in this chapter, which includes the design of microstrip grid array antenna excited by capacitive slot and microstrip grid array antenna excited by coaxial-fed slot.

5.1 A microstrip grid array antenna excited by capacitive coupling slot

Most of the microstrip grid array antennas in the literatures are fed by probe drilled through the ground plane. As an important excitation method for microstrip patch antenna, slot excitation has not been used for microstrip grid array antenna before. Hence, it is worth exploring slot excitation on microstrip grid array antennas. The slot is located in the middle of the ground plane with the long side along the H-plane of microstrip grid array antenna. Coplanar waveguide (CPW) is used to feed the slot with its advantage of easy integration of active or passive components. There are two kinds of CPW-fed slot excitation: the inductive coupling slot corresponding to a short-circuited slot and the capacitive coupling slot corresponding to an open-circuited slot [97]-[98], as shown in Fig. 5.1. In inductive coupling slot, the electric field is distributed all along the slot on both sides of the signal line. The slot is not directly under the grid array radiator, so the grid array radiators cannot be effectively excited. In capacitive coupling slot, the electric field is mainly concentrated in the gap width in the line region, and the electric field in the slot has the same direction with the short sides of grid array radiator over the slot. Thus, capacitive coupling slot can realize more efficient excitation.

(a) Inductive coupling slot

(b) Capacitively coupling slot

Fig. 5.1. CPW-fed slot excitation.

The model of microstrip grid array antenna excited by capacitive coupling slot is shown in Fig. 5.2. The whole dimension of the antenna is $70\text{mm} \times 70\text{mm} \times 0.787\text{mm}$. Rogers 5880 with dielectric constant of 2.2 and tangent loss of 0.0009 is chosen as the substrate. The antenna only has two metal layers. The grid array radiator of 50 loops is on the top surface of the substrate. The long sides of the loops are operated as transmission lines and the short sides of the loops serve as both radiators and transmission lines. The length and the width of the short sides are 4.9 mm and 2.5 mm, respectively, while they are 9.8 mm and 0.4 mm of the long sides. A ground plane with a slot in the center is on the bottom surface of the substrate. It is found that the capacitance of the input impedance increases as the slot width increases. The real part of input impedance increases as the slot length increases. To achieve good matching, the length of the slot s_l is chosen as 15 mm and the width of the slot s_w is chosen as 0.3 mm. As shown in Fig. 5.2 (b), 50- Ω CPW line is used to feed the slot. In order to mitigate the influence of CPW line to radiators, the length of the CPW line is chosen as 3mm, which is between a quarter and a half of guided wavelength. The width of the CPW signal line is 3mm and the width of gap is 0.1mm. A de-embedded lumped port is used to excite the CPW line.

(a) Microstrip grid array antenna.

(b) Zoom-in view of capacitive coupling slot.

Fig. 5.2. Microstrip grid array antenna excited by capacitive coupling slot.

Fig. 5.3 (a) and (b) show the simulated $|S_{11}|$ and peak realized gain as a function of frequency, respectively. The -10-dB impedance bandwidth is 2.95 GHz from 22 GHz to 24.95 GHz, which covers the 24-GHz band from 24 GHz to 24.25 GHz. The maximal peak realized gain is 18 dBi at 24.4 GHz and 3 dB gain bandwidth is 2.15 GHz from 23.03 GHz to 25.18 GHz. Fig. 5.3 (c) and (d) show the radiation pattern in E-plane and H-plane at 24 GHz, respectively. As expected, the antenna radiation patterns have the main beam of radiation in the broadside direction and weak cross-

polarization lower than 20dB. However, since the slot radiates into both sides of the ground plane, undesirably low front-to-back ratios are observed, which can be enlarged by adding a cavity behind the slot. Fig. 5.4 and Fig. 5.5 show the E-field distribution in the slot and current distribution on the grid array radiator, respectively. As expected, the electric field is mainly concentrated in the gap width in the line region, and the electric field in the slot has the same direction with the short sides of grid array radiator over the slot, which makes the uniform of the current distribution on the grid array radiators.

(a)

(b)

Fig. 5.3. Simulated performance of the microstrip grid array antenna excited by capacitive slot in Fig. 5.2: (a) $|S_{11}|$ as a function of frequency, (b) peak realized gain as a function of frequency, (c) radiation pattern in E-plane at 24 GHz and (d) radiation pattern in H-plane at 24 GHz.

Fig. 5.4. E-field distribution in the capacitive slot.

Fig. 5.5. Current distribution on the grid array radiator excited by capacitive slot.

Next, the microstrip grid array antenna excited by inductive coupling slot is also investigated. The substrate and the grid array radiator on the top surface of the substrate are the same as those in Fig. 5.2 (a). Fig. 5.6 shows the zoom-in view of the inductive coupling slot in the middle of the ground plane. It is found that as the length of the slot increases, the real part of the input impedance increases. As the width of the slot increases, the capacitance of the input impedance increases. The length of the slot s_l is 11 mm and the width of the slot s_w is 1.6 mm. A short CPW line whose length is 3 mm is used to feed the slot. The width of the CPW signal line is 2 mm and the width of gap is 0.1mm to achieve a 50- Ω characteristic impedance. A de-embedded lumped port is used to excite the CPW line.

Fig. 5.7 shows the simulated performance of the microstrip grid array antenna excited by inductive slot. The antenna achieves a -10-dB impedance bandwidth of 3.04 GHz from 22.96 to 26 GHz. However, it has a peak realized gain of 11 dBi and high sidelobes in the radiation patterns in both E-plane and H-plane, which demonstrates that the grid array radiator is not efficiently excited. Fig. 5.8 and Fig. 5.9 show the E-field distribution in the slot and the current distribution on the grid array radiator at 24

GHz. As expected, the E-field in the slot is distributed all along the slot on both sides of the signal line and weak on the signal line which is directly under the grid array radiator. Thus, the grid array radiator is not efficiently excited and has nonuniform current distribution, which results in the low gain and high side lobes in Fig. 5.7.

Fig. 5.6. Microstrip grid array antenna excited by inductive coupling slot.

(a) $|S_{11}|$

(b) Peak realized gain

Fig. 5.7. Simulated performance of the microstrip grid array antenna excited by inductive slot in Fig. 5.6.

Fig. 5.8. E-field distribution in the inductive slot.

Fig. 5.9. Current distribution on the grid array radiator excited by inductive slot.

5.2 A 24 GHz microstrip grid array antenna excited by coaxial-fed slot

In this section, a microstrip grid array antenna excited by coaxial-fed slot on Rogers 5880 substrate is reported. This work demonstrates that compared with capacitive slot, coaxial to slot transition is a more effective method to excite microstrip grid array antenna.

5.2.1 Microstrip grid array antenna excited by coaxial-fed slot

Fig. 5.10 shows the explored view of the microstrip grid array antenna excited by coaxial-fed slot. The whole dimension of the antenna is $70\text{mm} \times 70\text{mm} \times 1.041\text{mm}$. It consists of two dielectric and three metallic layers. All of the metallic layers are made up of 0.017mm thick copper and the dielectric layers are Rogers 5880 with a dielectric constant of 2.2 and tangent loss of 0.0009. The top layer is the grid array radiator composed of 50 rectangular loops. The long sides of the loops are operated as transmission lines and the short sides of the loops serve as both radiators and transmission lines. The widths of the short sides and long sides are chosen as $w_s = 2.5\text{mm}$ and $w_l = 0.4\text{mm}$, respectively. The resonance of the grid array radiator demands the length of the short sides $s = 4.9\text{mm} \approx \lambda_g/2$ and length of the long sides $l = 9.8\text{mm} \approx \lambda_g$. It is found in the simulation that the inductance introduced by the slot lead to the decrease of resonance frequency of grid array radiator. This effect can be alleviated by shortening the length of short sides near the slot $w_0 = 4.2\text{mm}$. The dielectric substrate layer (sub₁) with a thickness of 0.787mm is between the radiator and the metallic ground plane (gnd₁). A slot with a length of 10 mm and width of 0.7mm is located in the middle of the gnd₁ initially. To increase the front-to-back ratios of the antenna, a cavity is added below the slot. The cavity is realized by adding

a 0.254 mm dielectric layer (sub₂) and a metal layer (gnd₂) below the gnd₁. Fencing vias with length of 0.254 mm and diameter of 0.5 mm are located around the slot to connect the gnd₁ and gnd₂. The space between vias (from centre to centre) is 0.6mm. The fencing vias form a cavity with length of $c_l = 12$ mm, width of $c_w = 6$ mm and height of $h_l = 0.254$ mm. The slot is excited by coaxial probe with Teflon ($\epsilon_r = 2.1$) filling, inner diameter 0.5 mm and outer diameter 2 mm. A circular opening of diameter 2 mm is set on gnd₂. The inner pin of coaxial probe is connected to the centre of the slot on one side and the outer conductor is connected to gnd₂. Eleven vias are set to connect the other side of slot to gnd₂.

Fig. 5.10. Microstrip grid array antenna excited by slot: (a) explosion view, (b) top view.

The coaxial line supports a TEM mode, on which the electric field lines of the guided wave is parallel to the ground plane. The electric field lines in the slot are parallel to the short side on the ground plane. The TEM mode excited by the coaxial line under the slot can be smoothly transited into the slot. The short side of the grid array over the slot also supports an electric field parallel to the short sides, which has the same direction with electric field in the slot. So the fields in slot can effectively excite the grid array radiator.

Three kinds of modes are supported from the input port. They are cavity resonant mode, parallel plane mode and grid array mode.

The resonance frequency of cavity resonant mode with the cavity length of c_l , width of c_w and height of h , is given by

$$f_{mnr} = \frac{c}{2\pi\sqrt{\mu_r\epsilon_r}} \sqrt{\left(\frac{m\pi}{cl}\right)^2 + \left(\frac{n\pi}{cw}\right)^2 + \left(\frac{r\pi}{h}\right)^2}$$

where c is the light velocity in free space.

Gnd₁ and Gnd₂ form a parallel plane structure, which may support TEM mode, TM mode and TE mode. TEM mode can be supported without cutoff frequency. The substrate between gnd₁ and gnd₂ is 0.254 mm, so the cutoff frequencies for TM₁ mode and TE₁ mode are both $f_c = 398.15$ GHz. Thus only TEM mode can be supported in parallel plane structure at 24 GHz band.

The grid array mode is determined by the dimension of the grid array radiator. The slot introduces additional inductance to the grid array which leads to the decrease of resonant frequency of grid array mode.

In the following, we will discuss the effect of cavity size, grid array radiator size and the position of cavity and slot.

Fig. 5.11. Z_{11} of microstrip grid array antenna excited by coaxial-fed-slot for $c_w=4.5$ mm, 6, and 7.5mm: $c_f=12$, $l=9.8$ mm, $w_f=0.4$ mm, $s=4.9$ mm, $w_s=2.5$ mm, $s_f=10$ mm and $s_w=0.7$ mm.

Fig. 5.11 shows the effect of the c_w on Z_{11} . As can be seen, only when $c_w=6$ mm which is about half wavelength at 24 GHz, the antenna has a resonance around 24 GHz. So the cavity width c_w should be chosen around half a wavelength at the operating frequency.

Fig. 5.12. Z_{11} of microstrip grid array antenna excited by coaxial-fed-slot as a function of frequency for $c_f=9$, 12 and 15mm: $c_w=6$ mm, $l=9.8$ mm, $w_f=0.4$ mm, $s=4.9$ mm, $w_s=2.5$ mm, $s_f=10$ mm and $s_w=0.7$ mm.

The effect of c_l is examined. With reference to Fig. 5.11, the c_w is chosen as 6mm. From Fig. 5.12, we can clearly see that as c_l increases, the resonance frequency of cavity resonant mode decreases. The cavity resonant mode has a high Q factor, which

leads to a resonant impedance higher than 50 Ω . The grid array resonant mode becomes clear when the cavity resonant mode is near 24 GHz. Its resonant frequency is not affected by the change of c_l , and it is mainly controlled by the grid array size. The resonant frequency of grid array mode is around 23.5 GHz, because the slot in ground plane introduces additional inductance to the input port and has mutual inductance with grid array radiator, which decreases the resonance frequency of grid array mode. Adjusting the length of the short sides near the slot can efficiently compensate this inductance, as shown in Fig. 5.13. When w_0 is reduced to 4.2mm, the resonance frequency increases to 24 GHz and the 10 dB impedance bandwidth covers the 24 GHz band.

Fig. 5.13. Z_{11} of the modified microstrip grid array antenna as a function of frequency for $w_1=3.9, 4.2, 4.5$ and 4.9 mm: $c_w=6$ mm, $c_l=12$ mm, $l=9.8$ mm, $w_l=0.4$ mm, $s=4.9$ mm, $w_s=2.5$ mm, $s_l=10$ mm and $s_w=0.7$ mm.

The optimized parameters of the antenna are chosen as: $c_w=6$ mm, $c_l=12$ mm, $l=9.8$ mm, $w_l=0.4$ mm, $s=4.9$ mm, $w_s=2.5$ mm, $w_l=4.2$ mm, $s_l=10$ mm and $s_w=0.7$ mm. Fig. 5.14 (a) and (b) show the simulated $|S_{11}|$ and peak realized gain as a function of frequency, respectively. We can see that the input impedance bandwidth is 1.12 GHz from 23.76 GHz to 24.88 GHz. The 3 dB gain bandwidth is 1.68 GHz from 23.52

GHz to 25.2 GHz and the maximal peak realized gain is 20.57 dBi at 24.72 GHz. Fig. 5.15 (c) and (d) show the radiation pattern in E-plane and H-plane at 24 GHz, respectively. The front-to-back ratio of the radiation patterns is as high as 35.99dB at 24 GHz, which is due to the cavity added below the slot.

Fig. 5.14. Simulated performance of microstrip grid array antenna excited by coaxial-fed slot: (a) $|S_{11}|$, (b) peak realized gain, (c) radiation patterns at 24 GHz in Eplane, (d) radiation patterns at 24 GHz in H-plane.

5.2.2 Simultaneous transmission and reception application

(a)

(b)

Fig. 5.15. Microstrip grid array antenna excited by slot for simultaneous transmission and reception: (a) explosion view, (b) top view.

The microstrip grid array antenna excited by slot for simultaneous transmission and reception as shown in Fig. 5.15 can be functioned as simultaneously transmitted and

received antennas by adding two ports which are excited by coaxial pins at the two sides of port 1. The size of the grid array radiator, the cavity and the coaxial-fed-slot are the same as that chosen in Section 5.2.1. Port 1 is excited single-endly as a transmitting port. Port 2 and port 3 are excited in common mode with the same amplitude and phase, which are set as receiving port. The radiation pattern of the antenna will have a deep null in the broadside, leading to a high isolation between port 1 and the common mode excited by port 2 and 3. The input impedance of the common mode are calculated as

$$S_{cc11} = 0.5(S_{22} + S_{23} + S_{32} + S_{33}) \quad (5.1)$$

And the isolation between port 1 and the common mode are calculated as

$$S_{c1} = \frac{1}{\sqrt{2}}(S_{12} + S_{13}) \quad (5.2)$$

S_{c1} is mainly determined by the balance of the field distribution on both sides of the slot. The asymmetric of coaxial line to slot transition cause unbalance of the current distribution on both sides of the slot. It can be compensated by displacing the cavity, slot and coaxial line under the slot from the center of the ground plane. Fig. 5.16 shows the $|S_{c1}|$ as a function of the displacement p . With reference to the figure, the $|S_{c1}|$ decreases from -40.65 dB to -56.50 dB as p increases from -0.1mm to 0.1mm and increase from -56.50 dB to -44.62 dB as p increases from 0.1mm to 0.2mm, indicating that the $|S_{c1}|$ can be controlled by p effectively. With reference to the figure, the $|S_{c1}|$ of $p=0.1\text{mm}$ is close to that of the basic microstrip grid array antenna at 24 GHz. The surface current on the ground plane when $p = 0.2 \text{ mm}$ is shown in Fig. 5.17. As expected, the surface current is balanced on both sides of the slot. It was found that the other performances of the antenna, such as matching and gain are not affected by the

slight position change of the cavity and slot. Therefore, the slot-differential ports isolation problem can be solved without the need to re-tune other performance.

Fig. 5.16. $|S_{c1}|$ of microstrip grid array antenna for simultaneous transmission and reception as a function of frequency for $p = -0.1, 0$ and 0.2 mm.

Fig. 5.17. E-field distribution on the ground plane at 24 GHz when $p = 0.1$ mm.

(a)

(b)

(c)

(d)

Fig. 5.18. Simulated results of microstrip grid array antenna for simultaneous transmission and reception: (a) $|S_{11}|$ and $|S_{cc11}|$, (b) peak realized gain, (c) transmitted radiation pattern in E-plane at 24 GHz, (d) transmitted radiation pattern in H-plane at 24 GHz, (e) received radiation pattern in E-plane at 24 GHz and (f) received radiation pattern in H-plane at 24 GHz.

Fig. 5.18 shows the simulated results of microstrip grid array antenna for simultaneous transmission and reception. The transmitted and received input impedance bandwidths are 1.12 GHz from 23.72 GHz to 24.84 GHz, and 0.88 GHz from 23.72 GHz to 24.60 GHz. Both of them are covered the 24-GHz band. The transmitted 3-dB gain bandwidth is 1.68 GHz from 23.6 to 25.28 GHz with a maximum gain of 19.51 dBi at 24.8 GHz. The received 3-dB gain bandwidth is 2.08 GHz from 23.2 to 25.28 GHz with a maximum gain of 21.01 dBi at 24.24 GHz. The transmitted radiation patterns have broadside mainbeam and weak cross-polarization, while the received radiation patterns have a deep null in the broadsides direction, which result in high isolation between transmitting and receiving ports as expected.

5.3 Summary

The microstrip grid array antenna is usually fed by single-ended coaxial. The single-ended coaxial cable has an advantage of being simple, but it influences the symmetric of the radiation pattern slightly. As frequency increases, via may introduce inductors to the grid array, which makes it different to achieve good matching. In this chapter, we investigated the microstrip grid array antennas excited by slot. The microstrip grid array antenna excited by capacitive coupling slot has simple structure but suffers from low front-to-back ratios. The microstrip grid array antenna excited by coaxial-fed slot with a cavity added below the slot can radiate symmetrical radiation patterns which can be used for STAR application. However, additional layer of dielectric substrate for cavity is needed.

Chapter 6. DUAL-POLARIZED MICROSTRIP MESH ARRAY ANTENNA

In this chapter, a novel idea to design mesh array antenna through microstrip technology as an extension of microstrip grid array antenna. In Section 6.1, a 24-GHz linear-polarized microstrip grid array antenna is designed and the design rules are given as well. In Section 6.2, a 24-GHz dual-polarized microstrip mesh array antenna is designed. In Section 6.3, a 94-GHz microstrip grid array antenna and a 94-GHz microstrip mesh array antenna is presented using the same idea.

6.1 Design of 24-GHz linearly-polarized microstrip grid array antenna

In this section, we investigate how to design a high-gain microstrip grid array antenna with pencil beam at 24-GHz band for automotive radar application. The radiating elements are arranged specially into a square shape to achieve pencil beam and the parameters of the antenna are chosen theoretically to obtain high gain. The design procedure of the 24-GHz microstrip grid array antenna is described. To confirm the feasibility of the proposed design procedure, a 41-element microstrip grid array antenna with a broadside beam is fabricated. According to all the measured results, the developed microstrip grid array antenna can be considered as a good candidate for automotive radar systems.

6.1.1 Design procedure

The configuration of 24-GHz microstrip grid array antenna is exhibited in Fig. 6.1. It consists of one dielectric and two metallic layers. Rogers 5880 with dielectric constant of 2.2 and tangent loss of 0.0009 with a thickness of 0.787 mm is chosen as

the substrate. The grid array radiator, which is on the top surface of the substrate, is composed of rectangular meshes whose long sides are l long and w_l wide and whose short sides are s long and w_s wide. It is fed by a coaxial cable with its outer conductor connected to the ground plane and its inner conductor connected to the junction of the mesh near the geometrical centre of the radiator.

The application of 24-GHz automotive radars to monitor blind spots requires the maximal directivity of the antenna to be larger than 20 dBi, the impedance and radiation bandwidth to be larger than 0.25 GHz and the single-shaped beam to be smaller than 15° [98]. Considering the specific maximal directivity and the required pencil-beam patterns, one can find that the required number of radiating elements to be 41, which leads to an estimation of the length a and width b of the substrate as 60×60 mm². The resonance of the grid array radiator requires the length of long sides $l = \lambda_g \approx 9.8$ mm, where λ_g is guided wavelength at 24.15 GHz. The width of long sides w_l controls the loss of transmission line and the radiation of the cross-polar components. According to [53], in order to obtain low transmission line loss and cross polarization, the same width w_l for all short sides for $w_l/h < 1$ with the characteristic impedance smaller than 200Ω is chosen. Considering realizability from fabrication and reliability for use in automobiles, we prefer $w_l = 0.4$ mm. Similarly, the resonance of the grid array radiator also requires the length s to be half of l , that is $s = l/2 = 4.9$ mm. It is more involved to determine the widths w_s because w_s sets the array radiation coefficients with the first order approximation of the coefficient being proportional to width [53]. For simplicity in fabrication, the short sides keep the same width with an optimized value $w_s = \lambda_g/4 \approx 2.4$ mm. The measured return loss, antenna gain and radiation patterns will be included in the next section, which fully verified the design with good measured performance.

Fig. 6.1. Configuration of 24-GHz microstrip grid array antenna.

6.1.2 Simulated and measured results

The designed 24-GHz microstrip grid array antenna is fabricated for experiments. Fig. 6.2 shows the photo of the fabricated antenna with dimensions of $60 \times 60 \times 0.787$ mm³. Four holes with diameter of 4 mm at the four corners of the substrate are reserved for stabilizing the radiator in order to flatten the whole antenna structure. The S-parameter measurements were carried out with a Rohde-Schwarz ZVA 67 four-port vector network analyzer up to 67 GHz.

Fig. 6.2. Photograph of the fabricated 24-GHz microstrip grid array antenna.

Fig. 6.3 shows simulated and measured $|S_{11}|$ as a function of frequency. The simulated input impedance bandwidth is 1.99 GHz from 22.46 GHz to 24.45 GHz (or 8.24% at 24.15 GHz). The measured impedance bandwidth is 2.45 GHz from 22.23 GHz to 24.68 GHz (or 10.14% at 24.15 GHz). Both of the simulated and measured $|S_{11}|$ cover the 24-GHz band from 24 GHz to 24.15 GHz and they are in good agreement. Fig. 4 shows the simulated and measured peak realized gain as a function of frequency. The simulated 3-dB gain bandwidth is 1.54 GHz from 23.45 GHz to 24.99 GHz with a maximal gain of 20.55 dBi at 24.15 GHz. The measured 3dB bandwidth is 1.65 GHz from 23.54 GHz to 25.19 GHz with a maximal gain of 19.26 dBi at 24.32 GHz. The 3-dB gain bandwidths also cover the 24-GHz band. The radiation patterns at 24.15 GHz in E-plane and H-plane are shown in Fig. 5 and Fig. 6, respectively. The simulated and measured 3-dB beamwidths are 16° in both E-plane and H-plane. They clearly demonstrate that the developed microstrip grid array antenna radiates pencil-beam patterns.

Fig. 6.3. Simulated and measured $|S_{11}|$.

Fig. 6.4. Simulated and measured peak realized gain as a function of frequency.

Fig. 6.5. Radiation patterns at 24.15 GHz: (a) E-plane, (b) H-plane.

6.2 Design of 24-GHz dual-polarized microstrip mesh array antenna

In this section, a microstrip mesh array antenna on RT/duroid 5880 substrate is designed for 24-GHz application. The microstrip mesh array antenna enlarges the system polarization diversity with the capability of being utilized to support the simultaneous reception in two orthogonal polarizations, as well as transmission in either polarization.

6.2.1 Configuration

The configuration of the proposed microstrip mesh array antenna with dual feeds on RT/duroid 5880 substrate is shown in Fig. 6.6. The total size of the proposed antenna is $65 \times 65 \text{ mm}^2$ ($5.2\lambda \times 5.2\lambda$), where λ is the free space wavelength at 24 GHz. The mesh array on the top layer shown in Fig. 6.6(b) can be decomposed into two grid arrays of 32 square loops with the same dimensions. The orange grid array radiates linearly polarized waves in the x -direction and the blue grid array radiates linearly polarized waves in the y -direction. The long sides operate as transmission lines and the short sides operate as both radiation elements and transmission lines. The length of the long sides is l , the length of the short sides is s , the width of the long sides is w_l , and the width of the short sides is w_s . Eight stubs with length of l_p and width of w_p are added around the two feeding points to match the input impedance to 50Ω . The distance between the stub and the middle point of the short side is l_d . The substrate between the mesh array and the ground is 0.787 mm, which is chosen to avoid the occurrence of the TE_1 mode surface wave. The microstrip mesh array antenna is fed by two coaxial cables with their inner pins connected to the radiator and their outer conductor connected to the ground plane.

The mesh array radiator is composed of two grid arrays. From the theory of grid array antenna [53], obtaining a resonant frequency of f_0 requires the length of l to be one-guided wavelength and s to be half-length of the long sides. It should be noticed that to achieve high isolation between the two microstrip grid arrays, the two grid arrays must be on the same layer and the crossing points must be at the middle of each short side. The electric fields are minimal at the middle point of each short side, and thus the coupling is the smallest when the two short sides cross at this middle point.

This requires the length of the long sides of the microstrip mesh array antenna to be strictly two times larger than the one of the short sides.

Fig. 6.6. The 24-GHz microstrip mesh array antenna: (a) 3D exploded view, and (b) top view of the radiator: the orange grid array radiates linearly polarized waves in the x -direction and the blue grid array radiates linearly polarized waves in the y -direction. ($s = 5.3$ mm, $l = 10.6$ mm, $w_s = 2.5$ mm, $w_l = 1.5$ mm, $l_p = 0.8$ mm, $w_p = 0.5$ mm and $l_d = 0.6$ mm).

The long sides operate as transmission lines connected to the short sides with high radiation resistance, so they should have high characteristic impedance. Meanwhile, due to the fabrication rules and the fact that the losses of the line increase quickly above $200\ \Omega$ with the decrease of line width, the characteristic impedance of the long sides should be lower than $200\ \Omega$.

The use of microstrip line would induce a minority orthogonal current component which is perpendicular to the propagation direction of the microstrip line which might decrease the isolation between two polarizations, this problem can be controlled by adjusting the width of the short sides. Moreover, the intersection of short sides introduces direct coupling from the complementary array of another polarization, which alters the input impedances at the input ports. Therefore, impedance-matching technique with matching stub is adopted to obtain good impedance matching.

6.2.2 Effect of width of the short sides

The width w_s is an important parameter for controlling the isolation between the two polarizations. The simulated $|S_{21}|$ of the microstrip mesh array antenna in the frequency range 23 to 25 GHz with w_s varying from 0.5 mm to 2.5 mm using HFSS is shown in Fig. 6.7. It can be seen from Fig. 6.7 that the isolation at 24.15 GHz between ports one and two increase monotonically with an increase of w_s , which is due to the increase of the isolation at the two microstrip grid arrays' crossing point. The E-field distribution on the mesh array radiator when $w_s = 1$ mm and 2.5 mm are shown in Fig. 6.8 (a) and (b), respectively. With reference to the figures, E-field at the crossing point is strong for $w_s = 1$ mm while weak for $w_s = 2.5$ mm, which lead to poor isolation for $w_s = 1$ mm and high isolation for $w_s = 2.5$ mm. A larger w_s results in higher isolation between ports 1 and 2 and a higher antenna gain. Eventually, from the above analysis

with the consideration of the fabrication process rules, $w_s = 2.5$ mm is chosen for our design.

Fig. 6.7. Effect of w_s on $|S_{21}|$. All other parameters are kept as described in Fig. 6.6.

(a)

(b)

Fig. 6.8. Simulated E-field distribution on the mesh array radiator at 24.15 GHz: (a) $w_s = 0.1\text{mm}$, (b) $w_s = 0.35\text{ mm}$.

6.2.3 Impedance matching technique using stub

Matching stubs play important roles in the microstrip mesh array antenna design. The length of stub is a crucial parameter for impedance matching. The simulated $|S_{11}|$ for different stub lengths of $l_p = 0.2, 0.8$ and 1.4 mm is shown in Fig. 6.9, where the case without the stubs is also shown (w/o stub). The $|S_{11}|$ increases when l_p increases from 0.2 mm to 0.8 mm , and decreases when l_p increases from 0.8 mm to 1.4 mm . It indicates that the $|S_{11}|$ can be controlled by the length of l_p . The $|S_{11}|$ level is the lowest when $l_p = 0.8\text{ mm}$, which is about a guided wavelength. From the above analysis, $l_p = 0.8\text{ mm}$ is chosen as the optimum value.

Fig. 6.9. Effect of l_p on $|S_{11}|$. All other parameters are kept as described in Fig. 6.6.

6.2.4 Antenna performance

The microstrip mesh array antenna on RT/duroid 5880 substrate is simulated in HFSS with optimized parameters: $s = 5.3$ mm, $l = 10.6$ mm, $w_s = 2.5$ mm, $w_l = 1.5$ mm, $l_p = 0.8$ mm, $w_p = 0.5$ mm and $l_d = 0.6$ mm. The simulated $|S_{11}|$ for the antenna excited at one port and the other port opened, the simulated $|S_{11}|$ and $|S_{21}|$ for two-port excitation are shown in Fig. 6.10. The simulated one-port excitation $|S_{11}|$ and simulated two-port excitation $|S_{11}|$ are almost the same due to the high isolation between ports 1 and 2. A simulated isolation of 23.10 dB has been achieved at 24.15 GHz, which is sufficient for the microstrip mesh array antenna to operate as two independent antennas for 24-GHz application.

The simulated one-port excitation, two port excitation peak realized gains versus frequency for the microstrip mesh array antenna is shown in Fig. 6.11. For one-port excitation, the simulated maximum realized gain is 18.16 dBi at 24.17 GHz and the 3-dB gain bandwidth is 0.65 GHz from 23.85 GHz to 24.50 GHz (or 2.69% at 24.15 GHz). For two-port excitation, the simulated maximum realized gain is 19 dBi at

24.17 GHz and the 3-dB gain bandwidth is 0.70 GHz from 23.85 GHz to 24.55 GHz (or 2.90% at 24.15 GHz).

The radiation patterns of the microstrip mesh array antenna at 24.15 GHz under one port excitation while the other port is opened is shown in Fig. 6.12. With reference to the figure, the co-polarized fields are higher than their cross-polarized counterparts by more than 10 dB which indicates a pure linear polarization.

The simulated radiation patterns of the microstrip mesh array antenna under two-port excitation at 24.15GHz are shown in Fig. 6.13. They reveal that both the radiation patterns in $\phi = 0^\circ$ plane and $\phi = 90^\circ$ plane are the same, which indicates that the mesh array antenna has achieved dual linear polarization in the x and y directions.

Fig. 6.10. Simulated $|S_{11}|$ and $|S_{21}|$ under one-port and two-port excitation.

Fig. 6.11. Simulated peak realized gain under one-port and two-port excitation.

Fig. 6.12. Simulated radiation patterns at 24.15 GHz under one-port excitation. (a) E-plane and (b) H-plane.

Fig. 6.13. Simulated radiation patterns at 24.15 GHz under two-port excitation. (a) E-plane and (b) H-plane.

6.3 94-GHz microstrip grid array antenna and microstrip mesh array antenna.

The 24-GHz microstrip grid array antenna and microstrip mesh array antenna can be scaled to higher frequencies for more applications. In this section, a microstrip grid array antenna and microstrip mesh array antenna are simulated and measured for 94-GHz imaging application. Low temperature co-fired ceramic (LTCC) with dielectric constant of 5.9 and tangent loss of 0.002 is chosen as their substrate.

6.3.1 94-GHz microstrip grid array antenna.

The exploded view, radiator, and testing pads of the 94-GHz microstrip grid array antenna are shown in Fig. 6.14. It consists of two ceramic and three metallic layers. The radiator on the top layer (rad) has 32 optimized rectangular loops on an LTCC substrate backed by a metallic ground (gnd). The long sides of the loops are operated as transmission lines and the short sides of the loops are operated as both radiators and transmission lines. The 32 optimized rectangular loops have 41 short sides or 41

radiating elements. The length of the long sides of the loops (l), the length of the short sides (s), as well as the width of the long sides (w_l) and the width of the short sides (w_s) are respectively $l = 1.62$ mm, $s = 0.81$ mm, $w_l = 0.1$ mm and $w_s = 0.25$ mm. The dielectric substrate of Ferro A6M LTCC has a 5.9 dielectric constant and a 0.002 loss tangent. The ceramic substrate layer (sub₁) between the top layer and the ground plane is 0.19 mm thick. A circle opening of diameter 0.3 mm is set on the ground plane (gnd). According to the LTCC layout rules, the whole thickness of all the layers should be larger than 0.5 mm. Therefore, a ceramic substrate layer of 0.29 mm (sub₂) is added below the ground plane layer. A pin of diameter 0.1 mm and length 0.5 mm is used to feed the radiator at the point f near the center of the microstrip grid array antenna. Eight fencing vias are located around the feeding pin to connect the ground plane (gnd) and the metal ring on the bottom. The width of the metal ring is 0.1 mm. A 0.15 mm $\times$ 0.15 mm square pad which covers the feeding pin on the bottom layer, together with the metal ring, forms the ground-signal-ground (GSG) testing pads.

Fig. 6.14. The 94-GHz microstrip grid array antenna: (a) 3D exploded view, (b) radiator (top view), and (c) testing pad (bottom view).

Fig. 6.15. Photographs of the fabricated microstrip grid array antenna: (a) 3D top view, (b) 3D Bottom view, and (c) zoom-in of the testing pad.

The photographs of the fabricated microstrip grid array antenna are shown in Fig. 6.15. It has a size of $10 \times 10 \times 0.5 \text{ mm}^3$. The metal ring and the square pad on the bottom layer form the testing pads to be touched by a GSG probe of $200 \text{ }\mu\text{m}$ pitch. All the measurements have been conducted with the custom probe-fed measurement set-up which is able to characterize probe-fed antennas in terms of reflection coefficient, gain, efficiency and 3D radiation pattern until 140 GHz [99].

The simulated and measured $|S_{11}|$ versus frequency are shown in Fig. 6.16. They are in excellent agreement. The measured 10-dB impedance bandwidth is 5.12 GHz from 90 GHz to 95.12 GHz (or 5.45% at 94 GHz).

Fig. 6.16. Measured and simulated $|S_{11}|$ versus frequency.

The simulated and measured peak realized gains versus frequency are shown in Fig. 6.17. The simulated maximum realized gain is 17.2 dBi at 94 GHz and the simulated 3-dB gain bandwidth is 6.34 GHz from 90.4 GHz to 96.8 GHz (or 6.74% at 94 GHz). The measured maximum realized gain is 17.1 dBi at 95.8 GHz and the measured 3-dB gain bandwidth is 5.5 GHz from 92.0 GHz to 97.5 GHz (or 5.85% at 94 GHz). The simulated directivity is also shown in Fig. 4. The total efficiency calculated from the measured peak realized gain and simulated directivity is 93.19% at 94 GHz , and the

total efficiency calculated from the simulated peak realized gain and simulated directivity is 95.95% at 94 GHz.

Fig. 6.17. Measured and simulated peak realized gain versus frequency.

The simulated and measured radiation patterns are shown in Fig. 6.18. Both the E-plane and the H-plane patterns have the main beam of radiations in the broadside direction with low cross-polarization level. The simulated half-power beamwidths of the radiation beam are 26° and 22° in the E-plane and the H-plane, respectively. The measured E- and H- plane half-power beamwidths are 25° and 22° , respectively. The measured co-pol radiation patterns agree well with the simulated radiation patterns. It is interesting to note that the simulated radiation pattern in E-plane is un-symmetric, which is due to the fact that the feeding point is not located at the geometric center of the grid array radiator. The simulated cross-polarization is up to -3 dB in diagonal plane ($\phi=45^\circ$), which demonstrates that the proposed antenna is a linearly-polarized antenna and it achieves the minimal cross- polarization when it receives wave with orthogonal polarization.

The simulated current distribution on the grid array antenna at 94 GHz is shown in Fig. 6.19. The instantaneous currents on the long sides are out of phase thus a weak cross polarization lower than -17 dB are achieved. The currents on the short sides are almost in phase which result in a main beam in the broadside direction.

Fig. 6.18. Simulated and measured radiation patterns of the grid array antenna at 94 GHz in: (a) E-plane, (b) H-plane.

Fig. 6.19. Current distribution on the microstrip grid array antenna at 94 GHz.

Table 6.1 lists the key data of the 94-GHz microstrip grid array, the microstrip patch array on LTCC substrate [100], SIW antenna array [101], and slot array on Rogers\Duroid6002 substrate [102]. Note that the microstrip grid array in this work achieves the highest aperture efficiency of 41.09% due to its uniform current distribution. The high total efficiency of 93.21% of the microstrip grid array antenna is due to its simple feeding network of smaller feeding loss.

Table 6.1. Planar antenna arrays for 94-GHz applications

Array type	No. of elements	Size mm ²	Measured max gain dBi	Computed aperture efficiency	Radiation efficiency
Patch [100]	4×4	9.3×9.3	14.2	19.59%	46.77% (94 GHz)
SIW [101]	4×4	30×20	18.8	10.25%	85% (95.25 GHz)
Slot [102]	1×4	40×34	19	3.95%	76.5% (80 GHz)
Grid array in this work	41	10×10	17.1	41.09%	93.21% (94 GHz)

6.3.2 94-GHz microstrip mesh array antenna

The configuration of the proposed microstrip mesh array antenna with dual feeds on Ferro A6M LTCC substrate is shown in Fig. 6.20. The total size of the proposed antenna is $10 \times 10 \text{ mm}^2$ ($3.13\lambda \times 3.13\lambda$), where λ is the free space wavelength at 94 GHz.

(a)

(b)

(c)

Fig. 6.20. The 94-GHz microstrip mesh array antenna: (a) 3D exploded view, (b) radiator (top view), and (c) testing pads (bottom view). ($s = 0.865$ mm, $l = 1.73$ mm, $w_s = 0.35$ mm, $w_l = 0.2$ mm, $l_p = 0.2$ mm, $w_p = 0.12$ mm and $l_d = 0.156$ mm).

The mesh array on the top layer shown in Fig. 6.20(b) can be decomposed into two grid arrays of 32 square loops with the same dimensions. The blue grid array radiates linearly polarized waves in the x -direction and the orange grid array radiates linearly polarized waves in the y -direction. The long sides operate as transmission lines and the

short sides operate as both radiation elements and transmission lines. The length of the long sides is l , the length of the short sides is s , the width of the long sides is w_l , and the width of the short sides is w_s . Eight stubs with length of l_p and width of w_p are added around the two feeding points to match the input impedance to $50\ \Omega$. The distance between the stub and the middle point of the short side is l_d . The substrate between the mesh array and the ground is 0.19 mm, which is chosen to avoid the occurrence of the TE_1 mode surface wave. Another substrate with thickness of 0.29 mm is added under the ground plane to improve the mechanical hardness of the antenna. The microstrip mesh array antenna is fed by two pins of diameter 0.1 mm and length 0.5 mm at f_1 and f_2 , respectively. Eight vias are set around each pin to connect the ground plane and the metal ring on the bottom layer. Two circle pads with the diameter of 0.12 mm are covered on the feeding pins and form the ground-signal-ground (GSG) testing pad with the metal rings, as shown in Fig. 6.20 (c).

The photographs of the fabricated microstrip mesh array antenna are shown in Fig. 6.21, which have the following optimized parameters: $s = 0.865$ mm, $l = 1.73$ mm, $w_s = 0.35$ mm, $w_l = 0.2$ mm, $l_p = 0.2$ mm, $w_p = 0.12$ mm and $l_d = 0.13$ mm. The procedures to measure the radiation performances of the fabricated antenna are included in reference [99].

(a)

(b)

(c)

Fig. 6.21. Photographs of the fabricated microstrip mesh array antenna. (a) 3D Top view, (b) 3D Bottom view, and (c) zoom-in view of testing pads.

The antenna performance under two excitation cases are measured. One excitation case is one-port excitation, that is the antenna excited at one port and the other port opened, and the other one is two-port excitation, that is the two ports of antenna are connected to two signal sources, respectively, with equal amplitude and phase. The simulated and measured $|S_{11}|$ for the antenna excited at one port and the other port opened, and the simulated $|S_{11}|$ for two-port excitation are shown in Fig. 6.22. The measured one-port excitation $|S_{11}|$ at both port one and port two overlap due to the symmetrical structure of the antenna. The simulated one-port excitation $|S_{11}|$ and simulated two-port excitation $|S_{11}|$ are almost the same due to the high isolation between ports 1 and 2. The 10-dB impedance bandwidths of the microstrip mesh array antenna are 6.91% (89.2 GHz to 95.7 GHz) for both simulated one-port and two-port excitation, and 6.91% (90 GHz to 96.5 GHz) for measured one-port excitation. Discrepancies between the simulated and measured one-port excitation $|S_{11}|$ are caused by fabrication tolerances. To diagnose the discrepancies, a post simulation on the substrate of dielectric constant of 5.8 and loss tangent of 0.002 is carried out. The $|S_{11}|$

in the post simulation shifts to high frequencies and is in good agreement with the measured results.

Fig. 6.22. Simulated and measured $|S_{11}|$ under one-port and two-port excitation.

The simulated isolation between the two ports of the microstrip mesh array antenna is shown in Fig. 6.23. A simulated isolation of 21.2 dB has been achieved at 94 GHz, which is sufficient for the microstrip mesh array antenna to operate as two independent antennas for 94-GHz application. No measured isolation is available due to limitation of the measurement setup (which supports one-port only).

Fig. 6.23. Simulated $|S_{21}|$ of the mesh array antenna under two ports excitation.

The simulated one-port excitation, two port excitation and measured one-port excitation peak realized gains versus frequency for the microstrip mesh array antenna is shown in Fig. 6.24. For one-port excitation, the measured maximum realized gain is 13.3 dBi at 93.6 GHz and the measured 3-dB gain bandwidth is 1.6 GHz from 92.5 GHz to 94.1 GHz (or 1.70% at 94 GHz), while the pre-simulated maximum realized gain is 14.4 dBi at 93.0 GHz and the simulated 3 dB gain bandwidth is 2.0 GHz from 91.8 GHz to 93.8 GHz (or 2.13% at 94 GHz). The post simulated directivity and peak realized for one-port excitation are also shown in Fig. 8. As expected, the post simulated peak realized gain 3-dB gain bandwidth shifts up to a high frequency and agrees well with the measured result. The total efficiency calculated from the measured peak realized gain and post simulated directivity is 69.7% at 94 GHz, and the total efficiency calculated from the post simulated peak realized gain and post simulated directivity is 76.5% at 94 GHz. The measured gain decreases faster than the simulated ones when frequency is higher than 94 GHz the worse measured matching at high frequency band affects the measured gain and the shrinkage of the LTCC substrate makes the size of the fabricated antenna differ from the simulated one. For two-port excitation, the pre-simulated maximum realized gain is 13.5 dBi at 93.2 GHz and the 3dB gain width is 92 GHz to 95 GHz, which covers the 94 GHz band.

The radiation patterns of the microstrip mesh array antenna at 94 GHz under one port excitation while the other port is opened is shown in Fig. 6.25. With reference to the figure, the co-polarized fields are higher than their cross-polarized counterparts by more than 20 dB in simulation and 10 dB in measurement which indicates a pure linear polarization. It also can be observed that the simulated and the measured co-polar patterns at 94 GHz agree very well and both have very low back lobes.

Fig. 6.24. Simulated and measured peak realized gain under one-port excitation and two-port excitation.

The simulated radiation patterns of the microstrip mesh array antenna at 94 GHz are shown in Fig. 6.26. They reveal that both the radiation patterns in $\phi = 0^\circ$ plane and $\phi = 90^\circ$ plane are the same, which indicates that the mesh array antenna has achieved dual linear polarization in the x and y directions.

Fig. 6.25. Radiation patterns of mesh array antenna at 94 GHz under one-port excitation in: (a) E- plane, (b) H- plane.

Fig. 6.26. Radiation patterns of mesh array antenna at 94 GHz under two-port excitation in: (a) E- plane, (b) H- plane.

6.4 Summary

A new dual-polarized microstrip mesh array antenna on RT/duroid 5880 substrate has been developed for 24-GHz band application as an extension of microstrip grid array antenna. In the proposed design, high isolation between two polarizations is achieved by adjusting the width of short sides and good impedance matching can be obtained by adding stubs near the input ports. When the two antenna ports are excited simultaneously, the proposed antenna achieved a simulated maximum realized gain of 19 dBi at 24.17 GHz, a 10-dB impedance bandwidth of 5.63% and a 3-dB gain bandwidth of 3.19% at 24.15 GHz. As a one-port excited antenna, it achieved a measured maximum realized gain of 13.3 dBi at 93.6 GHz, a 10-dB impedance bandwidth of 6.91% and a 3-dB gain bandwidth of 1.70% at 94 GHz. Those performances indicate good return loss and high gain even in a frequency as high as 24 GHz which suffers from high metallic and substrate loss. An obvious advantage of the microstrip mesh array antenna is that it can support the simultaneous reception in two polarizations or transmission in either polarization which enlarges the polarization

diversity of the system and reduces the system volume. Then, successful designs of a 94-GHz microstrip grid array antenna and a 94-GHz microstrip mesh array antenna prove extensive potential applications of the proposed idea at high frequencies.

Chapter 7. COMPARISON OF MICROSTRIP GRID ARRAY

ANTENNA AND PATCH ARRAY ANTENNA

Recent years, microstrip grid array antenna has been used in various application such as 24 GHz, 60 GHz, 77 GHz, 120 GHz and 145 GHz application, because it has all the usual benefit of the patch array antenna such as compact size, easy fabrication, conformable structure, light weight, and low cost. In this report, the performance of two-loop microstrip grid array antenna, conventional patch antenna and wideband patch antenna on the same substrate are compared. Although two-loop microstrip grid array antenna has the largest radiator size, narrowest impedance bandwidth and 3-dB gain bandwidth, it has the highest peak realized gain and the lowest cross-polarization level. 145-element microstrip grid array antenna is designed and compared to the wideband patch array antenna in [17]. The 145-element microstrip grid array antenna has the advantage of higher peak realized gain and higher aperture efficiency as a result of simple feed structure. Moreover, cross-polarization level is smaller with an almost equal sidelobe level. Although the impedance bandwidth and gain bandwidth of 145-element microstrip grid array antenna are narrower than those of patch array antenna, they still cover 24-GHz band.

7.1 Two-loop microstrip grid array antenna compared with patch antennas.

Two-loop microstrip grid array antenna is the basic radiating structure of microstrip grid array antennas. Fig. 7.1 shows two-loop microstrip grid array antenna on RT/duroid 5880 with a dielectric constant of 2.2, tangent loss of 0.0009, length of 30 mm, width of 30 mm and thickness of 0.254 mm. Two-loop grid array radiator is

located on top surface of the substrate. The long sides of the square loop have a length of 8.96 mm and a width of 0.4 mm. The short sides of the square loop have a length of 4.48 mm and a width of 2.5 mm. The grid array radiator is fed by a coaxial cable with its inner pin connected to the radiator and the outer conductor connected to the ground plane. The grid array radiator can be regarded as an array of patch or dipole elements fed by microstrip lines, so the resonant input resistance of microstrip grid array antenna can be controlled by offsetting the position of the feed point from the edge of the mesh junction.

Fig. 7.1. Two-loop microstrip grid array antenna fed by coaxial cable.

Fig. 7.2 shows the conventional patch antenna fed by coaxial cable, which is on the same substrate as two-loop microstrip grid array antenna. The patch on the top surface of the substrate has a length of 3.86 mm and a width of 4 mm. The feed point position offset $b_p = 1.1$ mm.

Fig. 7.2. Conventional patch antenna.

The inset-fed rectangular patch loaded with a pair of straight slots shown in Fig. 7.3 is proposed in [17]. This patch antenna achieves wide impedance bandwidth since the loaded slots create an effective width for the patch. The optimum physical dimensions of the wide-band patch antenna in Fig. 7.3 are listed in Table 7.1.

Fig. 7.3. Inset-fed microstrip rectangular patch loaded with two straight slots.

Table 7.1. Optimum physical dimensions.

Symbol	Value (mm)	Symbol	Value (mm)
l_p	3.87	l_t	3.18
w_p	6.60	w_t	0.15
l_s	3.45	l_i	1.45
w_s	0.25	w_i	0.15
p_s	0.25	w_{ms}	0.76

Fig. 7.4. $|S_{11}|$ of the compared antennas.

The two-loop microstrip grid array antenna in Fig. 7.1 is compared with the patch antennas in Fig. 7.2 and Fig. 7.3. The return losses, radiation pattern at 24.15 GHz, peak realized gain, and efficiency of the compared antennas are shown in Fig. 7.4-7.7, respectively. The performance of the antenna is summarized in Table 7.2. Although the wide-band patch antenna has widest 10-dB impedance bandwidth and 3-dB gain bandwidth, it has medium peak realized gain. Two-loop grid array antenna has the highest peak realized gain. The radiation patterns in E-plane of three antennas are similar. The two-loop microstrip grid array antenna has a narrower 3-dB beamwidth in H-plane and a lower cross-polarization level than those of conventional patch antenna

and wide-band patch antenna. The radiation efficiency of two-loop microstrip grid array antenna is slightly lower than the other two patch antennas due to transmission line loss.

Fig. 7.5. Peak realized gain of compared antennas.

Fig. 7.6. Radiation pattern in E-plane of the compared antennas.

Fig. 7.7. Radiation patterns in H-plane of compared antennas.

Fig. 7.8. Radiation efficiency of compared antennas.

Table 7.2. Simulated performance comparison

	Two-loop microstrip grid array antenna (Fig. 7.1)	Conventional patch antenna (Fig. 7.2)	Enhanced-bandwidth patch antenna (Fig. 7.3)
Radiator size	11.46 mm×4.88 mm	3.86 mm×4 mm	3.87 mm×6.6 mm
-10-dB impedance bandwidth	0.23 GHz (24.05 – 24.28 GHz)	0.72 GHz (23.88 – 24.60 GHz)	1.1 GHz (23.65 – 24.75 GHz)
Maximum peak realized gain	11.28 dBi at 24.16 GHz	7.41 dBi at 24.16 GHz	7.67 dBi at 24.35 GHz
3-dB gain bandwidth	0.7 GHz (23.8 – 24.5 GHz)	2.5 GHz (23.0 – 25.5 GHz)	2.1 GHz (23.6 – 25.7 GHz)
-3 dB beamwidth in E-plane (24.15 GHz)	104° (-56° to 48°)	106° (-56° to 50°)	100° (-48° to 52°)
-3 dB beamwidth in H-plane (24.15 GHz)	24° (-12° to 12°)	64° (-32° to 32°)	62° (-32° to 30°)
Maximum cross-pol level in E-plane (24.15 GHz)	-45.54 dB	-43.10 dB	-40.91 dB
Maximum cross-pol level in H-plane (24.15 GHz)	-24.82 dB	-18.78 dB	-10.15 dB
Maximum radiation efficiency	86.58%	96.53%	92.67%

7.2 Microstrip grid array antenna compared with patch array antenna and waveguide slot array antenna.

In this section, a 145-element microstrip grid array antenna is designed and compared to the patch array antenna in Ref. [17]. Both of the antennas on the same substrate of RT\duroid 5880 with dielectric constant of 2.2, tangent loss of 0.0009, length of 100 mm, width of 100 mm and thickness of 0.25 mm. Fig. 7.9 shows the sketch of the 8×8 corporate-fed planar array whose detail parameters can be obtained in the Table I in Ref. [17]. Fig. 7.10 shows the configuration of the 145-element microstrip grid array antenna, whose long sides have a length of 9.56 mm and width of 0.2 mm, and short sides have a length of 4.78 mm and width of 1.5 mm. Two stubs with length of 2.5 mm and width of 1 mm are specially designed for impedance matching. The simulated $|S_{11}|$, peak realized gain, radiation patterns and radiation efficiency are shown in Fig. 7.11-7.14.

Fig. 7.9. Sketch of an 8×8 corporate-fed planar array.

The list of comparisons is given in Table 7.3. In this design, the microstrip grid array antenna has 145 radiating elements showing a 1.45% 10-dB impedance bandwidth and

3.23% 3-dB gain bandwidth. The 10-dB impedance bandwidth and 3-dB gain bandwidth of microstrip grid array antenna are narrower than those of patch array antenna. This is because, as number of radiating elements of microstrip grid array antenna increases, the bandwidths decrease. Moreover, the stepped-impedance lines used to achieve impedance matching limit the impedance bandwidth as well. Nevertheless, both the 10-dB impedance bandwidth and 3-dB gain bandwidth of microstrip grid array antenna cover the 24-GHz band.

The peak realized gain of microstrip grid array antenna is significantly higher than that of patch array antenna and consequently shows higher aperture efficiency. The microstrip grid array antenna has the advantage of simple feeding structure, leading to less transmission line loss and more compact structure for the radiators. On the identical substrate, the microstrip grid array antenna has 145 radiating elements, while the patch array antenna only has 64 radiating patch, namely 128 radiating slots.

Besides, the microstrip grid array antenna and the patch array antenna have almost equal sidelobe level in both E-plane and H-plane. But they achieve low sidelobe level through different methods. For the patch array antenna, Taylor one parameter method is used to synthesize the excitation distribution in both E-plane and H-plane. The microstrip grid array antenna achieve low sidelobe level by simply remove the long sides far from the feed point. The sidelobe level of microstrip arrays is affected by many factors, such as excitation amplitude and phase accuracies, mutual coupling, diffraction effects, positioning errors, and errors due to imperfect element matching and feed network isolation. Since the substrate is electrically thin, where the surface wave is weak, the mutual coupling and diffraction effects are not crucial for both antennas. For patch array antenna, complicated coplanar microstrip feed network incurs gain loss and degradation in sidelobe level and cross polarization due to the

feed radiation from the discontinuity and resistive loss. Besides, the sidelobe level suppression of patch array antenna adopting Taylor one parameter method is more sensitive to the excitation amplitude and phase accuracies of each radiation patch. The resulting sidelobe level can be affected for even small errors in amplitude and phase.

The sidelobe level of microstrip grid array antenna can be reduced by simply removing the long sides at the edge of the grid array radiator without reducing the radiating elements and decreasing the gain. The sidelobe of the microstrip grid array antenna is mainly degraded by the unsynchronized current far from the feed point. Typically, the long sides of microstrip grid array antenna serve as transmission lines with their currents out-of-phase and cancelling out each other, and the short sides are operated as both radiators and transmission lines with their currents in phase and adding up. However, since the number of radiators increases, the current on both long sides and short sides far from the feed point becomes unsynchronized, causing high sidelobe and low gain. Fig. 7.15 shows the current distribution of the 145-element microstrip grid array antenna with long sides far away from the feed point. It can be seen that the currents on the long sides and short sides far away from the feed points are unsynchronized. To reduce the unsynchronized currents which are undesired, the long sides at the edge of the grid array radiator are removed without changing the current distribution near the feed point, as shown in Fig. 7.16. Fig. 7.17 (a) and (b) compare the radiation pattern of the 145-element microstrip grid array antenna with or without long sides at the edge of the grid array radiator in the E-plane and H-plane, respectively. The sidelobe level in the E-plane is greatly suppressed by removing the long sides at the edge of the grid array radiator, while the sidelobe level keeps unchanged in the H-plane. Besides, since no amplitude tapering method is used for

microstrip grid array antenna, the requirement for tolerance of fabrication is looser than that of patch array antenna.

The cross-polarization of microstrip grid array antenna is generally -65 dB and -35 dB below the maximum value of the co-polarization in E-plane and H-plane, respectively, which are better than those of patch array antenna. The cross-polarization level of microstrip grid array antenna is controlled by the long sides of the loops. As mentioned above, the long sides of each loop are approximately one guided wavelength long at the resonant frequency and consequently form out-of-phase currents. These out-of-phase currents produce a cross-polarized pattern with null plane along E-plane and H-plane. As the number of loops increases, the undesired cross-polarization levels are rapidly decreased due to the additional null planes. For patch array antenna, higher cross-polarization levels can be caused by the radiation from complicated feed network with unsmooth bents. Additionally, small variations in slot dimension and feed position may affect the symmetry of the antenna and excite undesired higher order modes, consequently leading to degradation of cross-polarization level.

In addition, the performance of the waveguide slot array antenna in Ref. [30] are summarized in Table 7.3 for comparison. The configuration of the waveguide slot array antenna is shown in Fig. 7.18. The waveguide slot array antenna with 16×16 elements has the widest -10 dB impedance bandwidth, highest peak realized gain and highest aperture efficiency among the three compared antennas. However, a cavity is needed for the waveguide slot array antenna to suppress the cross-polarization level, leading to bulky volume and difficulty to integrate it with other circuit components.

Fig. 7.10. Configuration of 145-element microstrip grid array antenna with two stubs.

Fig. 7.11. $|S_{11}|$ of 145-element microstrip grid array antenna with two stubs and patch array antenna.

Fig. 7.12. Peak realized gain of 145-element microstrip grid array antenna with two stubs and patch array antenna.

Fig. 7.13. Radiation efficiency of 145-element microstrip grid array antenna with two stubs.

Fig. 7.14. Radiation patterns at 24.15 GHz. (a) E-plane and (b) H-plane.

Fig. 7.15. Current distribution on the grid array radiator with long sides at the edge: (a) magnitude and (b) vector.

Fig. 7. 16. Current distribution on the grid array radiator without long sides at the edge: (a) magnitude and (b) vector.

Fig. 7. 17. Radiation patterns for the 145-element microstrip grid array antenna with or without long sides at the edge of the grid array radiator at 24.15 GHz: (a) E-plane and (b) H-plane.

Fig. 7. 18. Configuration of the 16×16 -element waveguide slot array antenna.

7.3 Conclusion

This chapter compares a 145-element microstrip grid array antenna and a bandwidth-enhanced patch array antenna. Both antennas are designed on single-layer RT/duroid 5880 substrate with size of $100 \text{ mm} \times 100 \text{ mm} \times 0.25 \text{ mm}$. The 145-element microstrip grid array antenna has the advantage of higher peak realized gain and higher aperture efficiency as a result of simple feed structure. Moreover, cross-pol level of microstrip grid array antenna is smaller with an almost equal sidelobe level since the currents on the long sides are cancelled out. Although the impedance bandwidth and gain bandwidth of 145-element microstrip grid array antenna are narrower than patch array antenna, they still cover 24-GHz band.

Table 7.3. Comparison among 145-element microstrip grid array antenna , patch array antenna and waveguide slot array antenna.

	Microstrip grid array antenna (Fig. 7. 10)	Patch array antenna in Ref. [17] (Fig. 7.9)	Waveguide slot array antenna in Ref. [30]
Dimension	10 mm×10 mm×0.254mm	10 mm×10 mm×0.254mm	60 mm × 61 mm × 3.2 mm
10-dB impedance bandwidth	0.35 GHz (24 – 24.35 GHz)	1.78 GHz (23.22 – 25 GHz)	16.1 GHz (70.7–86.8 GHz)
Peak realized gain	25.41 dBi at 24.15 GHz	22.5 dBi at 24.15 GHz	33.1 dBi at 80 GHz
3-dB gain bandwidth	0.78 GHz (23.70 – 24.48 GHz)	Larger than 1 GHz (23.65 – 24.65 GHz)	Larger than 23 GHz (67 –90 GHz)
Sidelobe level in E-plane	-18.35 dB at 24.15 GHz	-20.79 dB at 24.15 GHz	-27.1 dB at 78.5 GHz
Sidelobe level in H-plane	-18.30 dB at 24.15 GHz	-19.15 dB at 24.15 GHz	--
Cross-pol level in E- plane	-65 dB at 24.15 GHz	-20 dB at 24.15 GHz	-30 dB at 78 GHz
Cross-pol level in H- plane	-35 dB at 24.15 GHz	-20 dB at 24.15 GHz	-30 dB at 78 GHz
Aperture efficiency (24.15 GHz)	43.21%	22.11%	62.46%

Chapter 8. CONCLUSION AND FUTURE WORK

This chapter will conclude the thesis and shed light on the possible improvement and update of the microstrip grid array antenna for future work.

8.1 Conclusion

In this thesis, the microstrip grid array antennas have been extensively investigated for 24-GHz applications. The mutual coupling mechanism between sub grid arrays and three types of microstrip grid array antennas are deeply studied, and the research highlights are re-captured in the following summary.

1. The mutual coupling between sub grid arrays is studied for the first time. It is demonstrated that the mutual coupling between sub grid arrays on electronically thin substrate is mainly caused by polarized currents in the dielectric substrate and can be reduced by proper array factors of each sub grid arrays. The large microstrip grid array antenna form by sub grid arrays with high isolation has large design flexibility to implement STAR application, monopulse antenna and generating OAM modes. A design example microstrip grid array antenna with 5 sub grid arrays on RT/duroid 5880 substrate for STAR application has been given. The proposed microstrip grid array antenna can achieve high measured isolation of 65.07 dB at 24.15 GHz between transmitting and receiving ports.

2. A novel high-gain multi-port microstrip grid array structure of 145 elements has been proposed with large design flexibility, which enables the proposed structure easy to implement one or two unbalanced antennas as well as one or two balanced antennas. Stepped-impedance lines are specially designed on the long sides of the meshes near the feeding ports to achieve good matching. Isolations between different ports have been studied and a simple technique of cutting slots in the ground plane to perturb the

electromagnetic coupling between ports so as to increase the isolation has been simulated. It has been shown that the slot in the ground plane can dramatically improve isolation without degrading the radiation performance of the antenna. A design example multi-port microstrip grid array structure with 4 ports on RT/duroid 5880 substrate has been given. The designed multi-port microstrip grid array structure has excellent performances when it is operated as one or two unbalanced antenna as well as one or two balanced antenna. More interestingly, the novel multi-port microstrip grid array structure has the unique capability to function as MIMO antenna with envelope correlations between radiation patterns of the two antenna elements lower than -40 dB over the 10-dB input impedance bandwidth of the proposed structure.

3. Microstrip grid array antennas excited by slot are investigated for the first time. A novel microstrip grid array antenna excited by capacitive coupling slot has been designed for 24 GHz applications. And then, a microstrip grid array antenna excited by coaxial-fed slot is proposed. A cavity is added below the slot to increase the front-to-back ratios of the antenna. The inductance introduced by the slot reduces the resonance frequency of the grid array mode, which can be alleviated by reducing the length of the short sides on the grid array radiator near the slot. The symmetry of the radiation pattern is determined by the field balance on the both sides of the slot, which can be achieved by displacing the cavity, slot and the coaxial line below the slot from the center of the ground plane.

4. A new dual-polarized microstrip mesh array antenna on Rogers 5880 substrate has been developed for 24-GHz band application. In the proposed design, high isolation between two polarizations is achieved by adjusting the width of short sides and good impedance matching can be obtained by adding stubs near the input ports. The antenna

achieves good return loss and high gain even in a frequency as high as 24 GHz which suffers from high metallic and substrate loss. Finally, an obvious advantage of the microstrip mesh array antenna is that it can support the simultaneous reception in two polarizations or transmission in either polarization which enlarges the polarization diversity of the system and reduces the system volume.

5. This thesis also compares a 145-element microstrip grid array antenna and a bandwidth-enhanced patch array antenna. Both antennas are designed on single-layer RT/duroid 5880 substrate with size of $100\text{ mm} \times 100\text{ mm} \times 0.25\text{ mm}$. The 145-element microstrip grid array antenna has the advantage of higher peak realized gain and higher aperture efficiency as a result of simple feed structure. Moreover, cross-pol level of microstrip grid array antenna is smaller with an almost equal sidelobe level since the currents on the long sides are cancelled out. Although the impedance bandwidth and gain bandwidth of 145-element microstrip grid array antenna are narrower than patch array antenna, they still cover 24-GHz band.

In summary, this thesis introduces some interesting concepts, which can potentially solve some important problems of microstrip grid array antenna and inspire in-depth thought to develop an advanced microstrip grid array antenna for high-gain, low-profile and easy-fabrication applications.

8.2 Future work

This thesis solves several important problems for microstrip grid array antenna in 24-GHz applications and verifies that the microstrip grid array antenna has many advantages for high-performance millimeter wave application over the conventional antenna arrays such as patch arrays, dipole arrays and slot arrays. Further analysis, design and implementation should be carried out to increase the performance and take

full advantage of the microstrip grid array antennas. The recommendations are listed as follows.

First, the problem of unsynchronized current of microstrip grid array antenna with large radiating array is not perfectly solved and the reasons that cause unsynchronized current are not thoroughly understood. In this thesis, the problem of unsynchronized current of microstrip grid array antenna is alleviated by multi-port microstrip grid array structure. However, the multiple ports increase the antenna complexity. Previous studies reveal that unsynchronized currents might arise from the mutual coupling between radiating elements. Since grid array radiator is an interconnected radiator where the radiating elements are connected by transmission lines, the dominant coupling between radiating elements is directly coupling through transmission lines instead of surface wave coupling, near field coupling or far field coupling. The junctions of short sides and long sides of the meshes introduce extra parasitic capacitors and inductors which lead to phase errors for each radiating element. The phase errors have negligible effect on the radiating elements near the feed point, but they accumulate at the radiating elements far from the feed point resulting in significant unsynchronized currents which cause distorted radiation pattern. The circuit model of the junction can be built up to figure out how the parasitic capacitors and inductors affect the phase error. By controlling the phase error, the single-endedly fed large microstrip grid array antenna with synchronized currents can be achieved.

Second, microstrip grid array antenna used for OAM application can be further explored. Chapter 2 shows that microstrip sub grid array can achieve high isolation in a compact size demonstrating that microstrip sub grid array is a potential candidate for OAM application. By choosing a proper size of sub grid array and arranging it in a circle shape, symmetrical OAM modes can be generated.

Third, the application of microstrip grid array antenna can be extended to higher millimetre frequencies even Terahertz frequencies. THz waves are attractive for wireless communications because they can offer huge bandwidth, which is essential for increasing data capacity. The link budget for Terahertz short range radar application requires high gain antennas. However, as frequency increases, the surface wave power in the substrate becomes strong which cause large dielectric loss. Besides, the conductive loss dramatically increases in Terahertz frequency. For microstrip grid array antenna designed at Terahertz frequency, new materials such as graphene and new packaging technology such as fan-out wafer level package can be attempted.

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