Novel Analog Devices & Techniques For 5G Communication &
Instrumentation Using LWAs, FSSs And Metasurfaces

by

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ABSTRACT

With the emergence of 5G networks, the demand for faster, more reliable, and ubiquitous radio systems in communication, instrumentation, and sensors poses unprecedented challenges in microwave and millimeter-wave engineering, ranging from spectral regulation to electromagnetic compatibility (EMC). The recent trend in 5G technology has been to place an increasing emphasis on digital signal processing (DSP). However, while offering device compactness and processing flexibility, DSP suffers from fundamental drawbacks, such as high cost analog-digital conversion, high power consumption and poor performance at high frequencies. To overcome these drawbacks, analog techniques are finding renewed interest as a possible alternative to DSP-based processing, with particular promise for millimeter-wave applications. This thesis presents novel devices and techniques in this context for 5G applications. First, a linearly-polarized leaky-wave antenna array front-end is demonstrated at 60 GHz for real-time beam-switching for 5G communications applications. Leaky-wave antennas are designed to radiate their traveling-wave excitation continuously along their structure, intrinsically allowing frequency-scanning of their radiated beam. Second, an integrated analog spectrum analyzer is demonstrated over the unlicensed IEEE 802.11ad 5G band (57-64 GHz) with a nominal physical resolution of 1 GHz. The device is based on a novel conformal periodic side-fire Leaky-Wave Antenna with near-field focusing to provide a compact footprint convenient for use as a tool on a laboratory bench. The antenna and receivers are fully integrated into the substrate of a standard Printed Circuit Board (PCB), allowing for shielded spectral decomposition which protects the signal under test from electromagnetic (EM) radiation in the surrounding environment. Thirdly, optically transparent and ink minimized frequency selective surfaces (FSS) screen-printed on Polyethylene Terephthalate (PET) at 28 GHz are proposed and experimentally demonstrated. Useful for 5G EMC applications, these FSSs are an excellent choice for frequency selective shielding. Finally, a technique for the phase-only synthesis of multi-beam patterns in reflective metasurfaces based on a direct Fourier transform mapping of the near-field to far field is shown. Compared to Woodward-Lawson beam sampling, this synthesis features sharper transitions in shaped beams such as sector patterns, providing utility in potentially stringent indoor short-range radiation pattern specifications. These variety of proposed technologies provide an attractive solution to difficulties faced in common 5G applications such as spectral filtering, fundamental signal processing operations, and short-range indoor communications.

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ABBREVIATIONS

5G	5 th Generation
DF	Direction Finding
EM	Electromagnetic
EMC	Electromagnetic Compatibility
EMI	Electromagnetic Interference
FEM	Finite Element Method
FSS	Frequency Selective Surface
HFSS	High Frequency Structure Simulator
LOS	Line of Sight
LWA	Leaky-Wave Antenna
mm-Wave	Millimeter Wave
PCB	Printed Circuit Board
PEC	Perfect Electrical Conductor
PET	Polyethylene Terephthalate
PPW	Parallel-Plate Waveguide
RF	Radio Frequency
SIW	Substrate Integrated Waveguide
TE	Transverse Electric
TEM	Transverse Electromagnetic
TM	Transverse Magnetic
VNA	Vector Network Analyzer
WL	Woodward-Lawson
WR	Waveguide Rectangular

Chapter 1

THESIS MOTIVATION, OBJECTIVES & CONTRIBUTIONS

1.1 mmWave Real-Time Analog Wave Control

The 5th (5G) and future wireless generations are expected to require a frequency spectrum to support a large number of electronic devices. This has sparked interest in the millimeter-wave (mm-Wave) band where a large spectrum is readily available for use. Beyond the most common high-speed point-to-point wireless communication applications, 5G is expected to exploit the mm-wave spectrum in dense urban environments with short-range non-line-of-sight (LOS) channels. This transition to the mm-wave spectrum comes with a challenge to devise high-performance novel components and systems. Such systems and components include efficient multi-functional antennas, low-cost instrumentation systems such as spectrum analyzers, as well as devices for electromagnetic interference (EMI) protection at these frequencies.

One other common application in **wireless communication systems** is that of beam-forming. Beam-forming involves the real-time engineering of radiation patterns to meet various specifications in wireless systems. In cases where the radiation patterns produced by off-the-shelf antennas do not meet specifications, combinations of antennas, custom designed antennas, and specific excitation schemes can meet radiation pattern specifications. A common problem that occurs in radiation pattern formation is the angular coverage of the beam. In millimeter-wave applications, the beams produced by antennas tend to be highly directive and naturally narrow-beam. For applications where the angular beam coverage needs to be wider than is typically demonstrated by millimeter-wave antennas, novel and efficient techniques must be employed to improve coverage. Compared to microwave devices, these mm-wave devices (28 GHz and 60 GHz) feature smaller dimensions and increased sensitivity to tolerance as a result. Because of this, novel antenna beam-forming solutions must be employed to reduce the complexity of the hardware.

Next, of increasing interest is the implementation of mm-wave short range communication sys-

tems in indoor environments. Given the close proximity of indoor environments such as office spaces, along with the increasing density of 5G devices in these environments, EMI protection is of great importance. The use of EM shielding can be applied on the surfaces of walls to minimize EM interference with adjacent wireless systems and enhancing wireless security by minimizing the leakage of radio waves outside designated areas. Given the large scale and ubiquitous implementation of 5G systems, techniques for cost effective EMI protection are greatly needed.

Another difficulty faced in millimeter-wave 5G networks is the reliable implementation of fundamental signal processing operations such as spectral analysis for **instrumentation applications**. Traditional techniques to perform spectral analysis include discrete Fourier transforms (DFTs) based on digital computing for RF signals, where an analog test signal is first sampled, discretized and then digitally processed to compute its Fourier transform [3, 4]. However, this approach suffers from performance degradation and challenging implementation as frequency increases and signal bandwidths become large [5]. With the upcoming next generation wireless systems, novel technologies for both instrumentation and communication are required particularly in the IEEE 801.11ad (60 GHz unlicensed band) and 28 GHz mm-Wave band which are capable of handling large signal bandwidths with reduced latencies [6] [7].

1.2 Thesis Objectives & Contributions

This thesis is concerned with devising novel devices and techniques using analog means to target three focused applications: passive beam-forming based on LWAs and metasurfaces, spectral filtering & EMI reduction using FSSs and spectrum determination using an integrated side-fire antenna. In particular, this thesis has four main objectives:

1. Perform millimeter-wave **beam-forming** using compact and easy to implement **multi-port slot pair antenna arrays** at 60 GHz for wireless applications. The beam-forming is based upon applying varying phase gradients to each antenna through simply varying the cell periodicity so as to improve angular coverage radiated by the array. Slot pair antennas have the advantage of self-matching by virtue of their pair-wise geometry though their $\approx \lambda/4$ spacing. While this electrical spacing holds only for a particular frequency (60 GHz) and is thus narrowband in nature, its purpose is to eliminate the narrowband stopband commonly present in general leaky-wave structures, including slot antennas. This improves the system feasibility by eliminating the need for additional matching networks.
2. Perform real-time mm-wave **spectral analysis** by using a compact bench-top, shielded, and PCB based **integrated side-fire LWA** around the IEEE 802.11ad unlicensed band (57-64 GHz). The periodic LWA features the property of beam-scanning, and can be applied more generally to spatial-spectral decomposition where the beam location in space can be mapped roughly linearly to its frequency. The LWA itself is based on a novel self-matched T-junction array which radiates broadside though a shielded PCB. The device is made compact through

beam focusing by applying curvature to the antenna for near-field operation.

3. Reduce EMI in 5G short-range indoor communication systems through spatial **spectral filtering** using band-stop FSSs. The FSSs are screen-printed on PET using silver ink nanoparticles enabling thin, flexible, and optically transparent operation. Due to the FSSs flexibility, they can readily be applied to conformal surfaces in buildings for more robust applications. Due to their high optical transparency, they are largely unnoticeable in their surrounding environments, ideal for placement on windows, for example. Finally, the surfaces were ink-minimized by applying a periodic meshing scheme to the metallized component of the cells, improving optical transparency further and greatly reducing the amount of silver ink in printing for feasible cost-effective mass production.
4. Develop a **general beam-forming synthesis technique** using phase only modulations of unit cells in **reflective metasurfaces**. Of particular interest is the formation of arbitrarily shaped multi-beam radiation patterns of non LOS 5G applications. In these applications, large-scale surfaces consisting of $> 10^5$ cells are needed; a quantity prohibiting computationally intensive synthesis methods and optimization. The synthesis method is simple, and the phase only operation of the surfaces allows removal of cross-polarized field components in reflection. Compared to antenna array software, this synthesis method involves no optimization, and is more suitable for the synthesis of large arrays ($> 10^5$ elements) rapidly. The surfaces can also be implemented on flexible and screen-printed substrates.

This thesis has consequently led to the following publications:

1. Demonstration of a linearly polarized leaky-wave antenna front end for beam-forming at 60 GHz for wireless applications.
 - [1] M. K. Emara, **D. J. King**, H. V. Nguyen, S. Abielmona, and S. Gupta, Millimeter-Wave Slot Array Antenna Front-End for Amplitude-Only Direction-Finding, *IEEE Trans. Antennas Propagat.*, vol. 68, no. 7, pp. 5365-5374, July 2020.
 - [2] M. K. Emara, **D. J. King**, H. V. Nguyen, S. Abielmona, and S. Gupta, A beam-scanning slot array antenna for millimeter-wave direction finding applications, *EuCap, Krakow, Poland*, April 2019.
 - [3] M. K. Emara, **D. J. King**, H. V. Nguyen, S. Abielmona and S. Gupta, Multi-port Slot Array Antenna for Millimeter-wave Direction Finding and Beam-forming Applications, United States, Patent filed, 2019.
2. Demonstration of 60 GHz integrated side-fire LWA with a suppressed stop band in both rectilinear and conformal structures; the latter showing promise for shielded mm-wave real-time spectral analysis.

- [4] **D. J. King**, M. K. Emara and S. Gupta, Millimeter-Wave Integrated Side-Fire Leaky-Wave Antenna and its Application as a Spectrum Analyzer, *IEEE Trans. Antennas Propagat.*, June 2020, *under review*.
- [5] **D. J. King**, M. K. Emara, and S. Gupta, Millimeter-wave near-field spectrum analyzer based on integrated side-fire antenna, *EuCap, Krakow, Poland*, April 2019.
- 3. Demonstration of flexible and ink-minimized screen-printed frequency selective surfaces (FSSs) with increased optical transparency for 5G EMI mitigation.
 - [6] **D. J. King**, K. Hettak, R. Chaharmir and S. Gupta, Flexible Ink-minimized Screen-Printed FSS with Increased Optical Transparency for 5G EMI Mitigation, *IEEE Transactions on Electromagnetic Compatibility*, Jul 2020, *under review*.
 - [7] **D. J. King**, K. Hettak, R. Chaharmir, J. Ethier and S. Gupta, Ink Minimized Meshed Frequency Selective Surfaces with Increased Optical Transparency for 5G, *IEEE APS-URSI, Montreal, Canada*, July 2020.
- 4. Phase-only synthesis methods for general beam-forming reflective metasurfaces.
 - [8] **D. J. King**, K. Hettak, R. Chaharmir, J. Ethier and S. Gupta, Phase-only Synthesis Methods for General Beam-forming Reflection Metasurfaces, *IEEE APS-URSI, Montreal, Canada*, July 2020.

1.3 Thesis Organization

This thesis is organized as follows: Chapter 2 demonstrates a linearly polarized LWA front end for increased radiation pattern coverage on a sectoral basis at 60 GHz for 5G applications. Chapter 3 proposes an analog, low-profile and shielded spectrum analyzer for operation at mm-wave frequencies around the 60 GHz band based on a novel side-fire LWA configuration for instrumentation applications. Chapter 4 demonstrates flexible and ink-minimized screen-printed FSSs with increased optical transparency for 5G EMI mitigation at 28 GHz, as well as reduced cost for increased feasibility in mass production. Chapter 5 finally describes a direct Fourier transform phase-only synthesis method for beam-forming reflective metasurfaces, featuring promise for accurate production of shaped beams over other synthesis methods. Finally, summary and future steps are provided in Chapter 6.

Chapter 2

MULTI-PORT LEAKY-WAVE ANTENNAS (LWAs) FOR BEAM-FORMING

2.1 Introduction and Specifications

The design of robust millimeter-wave communication systems requires understanding the channel dynamics with respect to space and time at these frequencies. mm-wave signals are highly susceptible to blocking, and they have communication limitations due to their poor signal attenuation compared to microwave signals. Therefore, by employing highly directional antennas, co-channel interference to or from other systems can be alleviated using non-LOS propagation. Because of the ability to shape, switch, or scan the propagating beam, phased arrays play an important role in advanced wireless communication systems [8].

Beamforming is essentially a spatial filtering operation typically using an array of radiators to capture or radiate energy in a specific direction over its aperture. Beamforming allows the formation of beams in space and time unachievable by arrays without designed excitation schemes. Modern communication systems deploy smart antenna systems, which can combine array gain with diversity gain along with interference mitigation to further increase the capacity of the communication link. This is achieved by electronic beam steering using a phased array, which is a multi-element radiation device with a specific geometric configuration. The output spatial power distribution is determined by the vector sum of the fields radiated by individual elements. It can be expressed in terms of the individual element radiation pattern and the array factor, which is a function of the array geometry and amplitude/phase shifts applied to individual elements [9].

The functionality of phased arrays, however, can also be achieved through proper design and choice of the antennas themselves, without the need for additional phase-shifting hardware, which can be expensive at mm-wave frequencies. The requirements of phase shifters include large phase-control range, small phase-shift step size, low insertion loss, low gain, and low variation in loss over all phase states. Furthermore, phase shifters need to achieve low power consumption, occupy

a small chip area and be simple to control. The loss and loss-variations of a phase shifter can be (partly) overcome using an extra variable gain amplifier (VGA) stage in front of the phase shifter, but such a VGA not only consumes large chip area and power consumption but also becomes difficult to design at mm-wave frequencies. [10]. Additionally, for MIMO applications, the number of phase shifters and attenuators increases as the number of antennas and beams increases, which can be very large. [11]. Travelling-wave antennas such as slot array antennas have their leakage (or radiation) phased by the travelling-wave nature of the excitation. By varying the location of the leakage sources, or analogously the radiators (slots), the beam can be steered depending on the excitation frequency. While this is narrowband owing to the lack of tunability, an array of such antennas can improve bandwidth, while also providing real-time operation otherwise not achievable in traditional phased arrays.

In this chapter, a novel multi-port LWA front-end based on SIW for angular coverage improvement of isolated slot-pair antennas is proposed and experimentally demonstrated for the IEEE 802.11ad band around 60 GHz. The proposed antenna front-end is low-profile and compatible with the standard PCB process.

The proposed antenna system is based on a parallel LWA array. A LWA is typically a 2-port travelling wave structure that gradually radiates EM energy to free-space as the wave propagates along it [12]. It is characterized by a complex propagation constant $\gamma(\omega) = \alpha(\omega) + j\beta(\omega)$, where α is the per unit length leakage along the structure and β is the propagation constant. LWAs can be considered as series-fed phased arrays, and are typically broadband and directive in nature. While α (and the physical aperture) controls the beamwidth of radiation, β provides the necessary frequency-dependent phase-shifts to scan the radiation beams. Specifically, a LWA features a frequency dependent beam scanning (antenna assumed to lie in the $x - y$ plane), which is governed by

$$\theta(\omega) = \sin^{-1} \left[\frac{\beta(\omega)}{k_0} \right]. \quad (2.1)$$

This beam-scanning property of LWAs can be exploited in DF devices when operating in the receive mode for the desired frequency ω_0 . Consider a general 1-D array of N LWAs, as shown in Fig. 2.1. This is a multi-port antenna with $2N$ ports. Each of the LWAs of this array is designed to be associated with different $\beta_n(\omega_0)$, which can be achieved by modifications of the corresponding geometrical features of the radiating aperture. Consequently, each LWA radiates a beam at ω_0 pointing along a different angle θ_n . The slot pair cell periodicity is designed such that the radiation patterns intersect at -3 dB, creating a coverage sector as shown in Fig. 2.1. The coverage sector is thus determined by the number of antennas and the beamwidth. Moreover, the leakage factor or the physical length of the antenna is designed to radiate all the power input into the antennas for maximum efficiency.

Fig. 2.2 shows a graphical representation of the radiation specifications. When excited from the

Figure 2.1 An illustration of the proposed LWA front end consisting of N slot array antennas and $2N$ ports. Each antenna is designed with a unique periodicity (and hence a unique beam scanning law). The coverage sector is also shown (based on the beam scanning law of each antenna and their radiation patterns) at the design frequency, f_0 [1].

left, a beam centred at $\theta > 0^\circ$ is radiated. When excited from the right, a beam centred at $\theta < 0^\circ$ is radiated. Depending on the cell periodicity, the beam tilts at a certain angle. By designing the periodicity, we can achieve beam patterns which intersect at half power points along θ .

The frequency range of operation is 58-61 GHz. Across this channel, it was specified that in simulation transmission (S_{21}) for each antenna be < -10 dB, and reflection (S_{11}) be < -15 dB. It is desired to radiate as much power as possible to maximize operation efficiency, as well have as wide of beam transversally as possible, forming a fan beam.

2.2 Slot Array Antenna Design

To demonstrate the operation of this antenna system, let us consider a practical LWA structure first. While there is a wide variety of LWA structures reported in literature, waveguide-based slot array antennas are particularly suitable for our intended mm-wave frequency band due to their good radiation performances (high gain and efficiency) and design flexibility [13, 14]. Slot waveguide antennas are typically based on rectangular or parallel-plate waveguides operating in their fundamental mode, TE_{10} or TEM respectively, where a physical opening is made in the form of slots to allow the guided EM energy to gradually radiate out to free-space [15]. These openings

Figure 2.2 Beam pattern specifications. This is achievable by exploiting frequency scanning in periodic antennas.

are either continuous or discrete (typically periodic). While a continuous aperture radiates from its fundamental mode, periodic slot arrays radiate from the first space harmonic, $n = -1$ [16]. The slot array antennas are compatible with SIW structures, which makes them simple for fabrication using PCB technology. These antennas can be formed through etching slots and apertures out of the copper surface, and through the drilling of vias to form conductive walls isolating the antennas.

Periodic slot array antennas consist of a unit cell which, when repeated, forms the overall radiating aperture of the antenna. The unit cell may either be using a single slot (e.g. longitudinal slots) or compound slots for enhanced radiation characteristics. By controlling the geometrical features of these slots, the radiation characteristics such as beamwidth (through the leakage-factor) and polarization can be easily controlled. Fig. 2.1 shows one such configuration known as the reflection-canceling slot pair [17], used as a unit cell in this work, forming a 1-D LWA. This unit cell is transversely asymmetric and can be optimized to close the open-stopband at broadside, to achieve a wide-band performance and continuous frequency based beam-scanning through broadside [16]. Specifically, the slot dimensions, spacing, and offset from the center of the waveguide are optimized for low-reflection and hence high broadside gain. The beam-scanning law of such antenna follows

$$\theta(\omega) \approx \sin^{-1} \left(\frac{\beta_0 + n\pi/p}{k_0} \right), \quad (2.2)$$

where θ is the angle measured from broadside, k_0 is the fundamental wavenumber, n is the space harmonic number (typically -1), β_0 is the phase constant of the fundamental mode and p is the slot period. Eq. 2.2 assumes the leakage constant of the antenna, α , is small and can be neglected for the calculation of the main beam direction [18].

A 1-D array of such linearly-polarized LWAs can now be formed as shown in Fig. 2.3(a), which consists of 3 LWAs, i.e. $2N = 6$ ports, as an example. The antennas were implemented on Rogers RO4350 ($\epsilon_r = 3.66$, $\tan \delta = 0.004$, $h = 30$ mil = 0.762 mm). Each antenna is well-matched at 60 GHz, which is obtained by optimizing the slot-pair configuration. Moreover, each LWA is designed so that its radiation patterns at 60 GHz, corresponding to each excitation port intersects at -3 dB, as shown in Fig. 2.2(b). This is achieved by varying the periodicity p across the array, which controls its propagation constant $\beta(\omega)$ and thus the beam-scanning law of (2.2). The -3 dB intersection point was chosen such that there is balanced power between adjacent beams. This LWA structure can now be used to demonstrate its beam-forming capability. A full antenna beam-forming system is excited through feeding networks. Various feeding networks include corporate feeds, series feeds, and space feeds. Corporate feeds have the desirable characteristic of equal line length to each element, making the network symmetric that helps with balancing mutual coupling effects. Most important, the signal phase and amplitude to each element is naturally identical, as often used in practice. The equal line lengths also make the network frequency independent, and thus wide bandwidth. Series feeds offer advantages of simple construction of a compact structure. A disadvantage of the series feed is that the common feed line attenuates the signal to successive elements, causing loss and complicating the design. Series feeds can introduce dispersion that distorts short pulses and limits bandwidth. The series and parallel feeds are sometimes referred to as “constrained” feeds because of the physical constraint of continuous connectivity from the feed to the elements. In contrast, the space feed uses a primary antenna illuminating a separate unit consisting of pickup elements connected to a secondary radiating array face. Typically only used for large phased arrays, the space feed array takes advantage of the free space environment between the primary feed and the pickup array, leading to size, weight, and cost reduction compared to a fully connected feed network. The amplitude distribution of the array is determined by the primary feed pattern and the pickup array element placements. The phase introduced by the different path lengths from the primary feed to each pickup element is compensated for by the phase shifters on the radiating elements, which also are used for beam scanning [19]. However, a feeding network for this antenna array was not designed.

Figure 2.3 (a) Antenna array simulation model in FEM-HFSS. Airbox surrounded by radiation boundary conditions. Waveguide feeds are WR-15. (b) Slot antenna geometry. Slot parameters for each of the three antennas shown in table 2.1. (c) Antenna transition geometry. The parameters are given by: $a = 3.759$ mm, $b = 1.879$ mm, $c = 3.81$ mm, $s = 3.41$ mm, $x = 2.438$ mm, $z = 1.43$ mm, $a_i = 2.5$ mm, and $a_c = 5.438$ mm.

Table 2.1 Antenna Array Dimensions

Antenna	1	2	3
Number of Cells	13	14	15
P (mm)	2.75	2.565	2.395
ℓ (mm)	1.25	1.3	1.3
w (mm)	0.25	0.25	0.25
s_1 (mm)	0.69	0.69	0.69
s_2 (mm)	0.285	0.285	0.285
a_i (mm)	2.5	2.5	2.5

2.3 Experimental Results

Fig. 2.4 shows a photograph of the overall fabricated antenna front end, as well as a photomicrograph showing slot dimensions for a specific cell of antenna 1.

Figure 2.4 (a) Fabricated slot antenna array. (b) Photomicrograph of a slot pair showing fabricated dimensions.

Each dimension of the slot geometry in fabrication has at most a $\pm 50 \mu\text{m}$ error compared to simulation. It was observed the largest deviation in fabrication were the slot dimensions themselves, rather than their spacing or periodicity. Errors in the slot dimensions simply perturb the expected leakage factor α , resulting in radiation characteristics with slightly varying amplitudes compared to simulation. Given the relatively small error in slot dimensions, this amplitude variation is negligible in practice. The key parameters in the beam-forming operation are the slot spacing and period, which determine broadside matching and the steering angle. The tolerance on both of these parameters are sufficiently small for practical device operation. In general, key parameters in beamforming include beamwidth, SLLs, nulls, phase accuracy, and amplitude tapering. Here, the goal is simply to steer adjacent beams without amplitude tapering. Amplitude tapering can be used to reduce SLLs such as binomial tapers, however SLLs in the unweighted excitation scheme were considered sufficiently low.

2.3.1 S-Parameter Characterization

In this section, the measured S parameter results are shown. The S parameters were measured using a vector network analyzer (VNA) from Tokyo Institute of Technology. Coupling between antennas in the array was determined by measuring transmission across adjacent antennas. For example, if port 1 is excited, and the output is measured at port 5 or 6, then the coupling between antenna 1 and antennas 2 or 3 are respectively measured. Coupling was consistently measured to be very small (on the order of -60 dB), and thus is not a consideration in practical design. This is because each antenna is fully substrate integrated and their immediate surrounding environment in the plane of the device is that of the PCB, with long vias forming their boundaries acting as conductor walls.

In measurement, either all unused ports (4 total) were left as open, or they were matched with loads, as in Fig. 2.5(a) and (b). In simulation, adapters and feeding discontinuities were not modelled or accounted for. S-parameter discrepancies between simulation and measurement can be

accounted for due to this difference.

Figure 2.5 Measurement conditions for S parameters. (a) Antenna 2 is excited from the left and measured directly across, with all other ports open. (b) Antenna 1 is excited from the left and measured directly across. All other ports are matched.

Fig. 2.6 - Fig. 2.7 shows the total set of S parameter measurements compared with simulation with all other ports open and matched. In Fig. 2.6 a) and Fig. 2.7 a), measurements with ports 4, 5 or 6 connected refer to S parameters taken while those ports are connected to the second port of the VNA. From the conservation of power, $|S_{11}|^2 + |S_{21}|^2 - P_{diss} = P_{rad}$, where P_{rad} is the total radiated power and P_{diss} is the total dissipated power due to material loss. This relation holds for each frequency.

(a)

(b)

Figure 2.6 (a) First set of S parameters with other ports open. (b) Second set of S parameters other ports open. Coupling and transmission are shown.

Figure 2.7 (a) First set of S parameters with other ports matched. (b) Second set of S parameters other ports matched. Coupling and transmission are shown

2.3.2 Radiation Patterns and Gain Characterization

Fig. 2.8 shows the radiation patterns and gains of the antennas in both simulation and measurement. Three frequency points were taken (58, 60, and 61 GHz). The measured resolution was 1° . For the measured results, the -3 dB points did not exactly intersect, largely due to the measurement resolution, as well as material variation causing the antenna to have a different electric length compared to simulation, resulting in a different beam angle. In addition, fabrication error in the slot period can also shift the patterns.

2.4 Conclusions & Future Work

A 60 GHz linear antenna array for sectoral beam-forming and coverage enhancement for wireless applications and was designed, fabricated and measured. The specified reflection and transmission was <-15 dB and <-10 dB, respectively. In the array, antennas 1 and 2 met the reflection specification, while antenna 3 did not. None of the antennas met the transmission specification

Figure 2.8 (a) - (c) Simulated vs. measured normalized antenna radiation patterns at 58, 60, and 61 GHz, respectively. (d) - (f) Simulated vs. measured antenna gains at 58, 60, and 61 GHz, respectively.

due to slot length restrictions and board length limitations. Additionally, the dielectric properties of the samples used may have discrepancies compared to their reported values. For the radiation patterns, the intersections were not quite at -3 dB as specified. This is due to coarse angular increments on the measurement, as well as material deviation. The coverage at 60 GHz was found to be 41.5° . In all cases, maximum SLLs are below 0 dB. These SLLs are deemed sufficient for main beam coverage expansion. In more general implementations of this system, the LWAs could be designed with amplitude tapers to reduce SLLs such as Chebyshev weightings.

This antenna system has also been demonstrated in a 5G direction finding (DF) application [1], as well as implementation for circularly polarized operation [20]. Future work includes the formation of an LWA array with more general individual beam patterns using both amplitude and phase weightings to improve coverage in a wider range of applications.

Chapter 3

SIDE-FIRE LWA BASED INTEGRATED SPECTRUM ANALYZER

3.1 Introduction

In chapter 2, the beam-forming of slot array antennas was performed to improve the radiation coverage for millimeter-wave systems in free space. The beam-forming here was real-time, and was performed using simple phase-only gradients to steer the individual antenna beams. However, millimeter-wave beam-forming can be applied more generally, and for a wider range of applications. For instance, nonlinear phase gradients can provide directly shaped beams, and variation in the antenna structure itself can provide additional beam operations such as near-field focusing. Interestingly, engineered beams also find application in spectral analysis.

Spectral analysis is among the most fundamental signal processing operations required in engineering and forms the basis of various systems in application areas such as signal distortion measurement from audio equipment, spectrum monitoring for governmental frequency allocation of various radio services, and electromagnetic emissions testing of equipment [21, 22]. Traditional techniques include discrete Fourier transforms (DFTs) based on digital computing for Radio Frequency (RF) signals, where an analog test signal is first sampled, discretized and then digitally processed to compute its Fourier transform [3, 4]. However, this approach suffers from performance degradation and challenging implementation as frequency increases and signal bandwidths become large [5]. With the upcoming next generation wireless systems, novel technologies for instrumentation and communication are required particularly in the IEEE 801.11ad (60 GHz unlicensed band) mm-Wave band which are capable of handling large signal bandwidths with reduced latencies [6] [7].

Consequently, alternative techniques have been explored in the literature to perform spectrum analysis using analog methods. In the field of optics, material dispersion is utilized to real-time separate the various spectral components of a broadband test signal, commonly utilized in Real

Time Fourier Transformers (RTFTs) [23], Arrayed Waveguide Gratings (AWG) [24], prisms or to use the wavelength dependent diffraction orders in periodic diffraction gratings [25] and Virtual Image Phased Arrays (VIPA) [26], for instance. While such systems are not as flexible as a digital system, they offer high acquisition bandwidth and ultrafast system operation. Inspired from these optical systems, various solutions have been proposed at RF using both guided and radiative systems based on dispersion (frequency dependent group delay, $\tau(\omega)$) engineered Bragg gratings [27], spatial interferometers [28] and leaky-wave Antennas (LWAs), for instance, exploiting their frequency scanning properties [29, 30, 31, 32, 33]. However, all of these solutions suffer from either lack of device integration or are based on unshielded microstrip implementations which are not particularly suitable for higher mm-wave bands (e.g. 60 GHz, for instance) leading to significant free-space radiation, not ideal for minimal electromagnetic interference.

Previously, an idea of a novel mm-wave near-field real-time spectrum analyzer (RTSA) based on an integrated side-fire LWA has been proposed in [34]. Here this idea is further explored in detail and experimentally demonstrated. The subsequent contributions of this work are two-fold: 1) It is based on a novel side-fire LWA using integrated waveguide periodic apertures on its side walls, providing a typical frequency scanned power leakage *along the plane of the antenna* as opposed to radiation normal to the antenna plane in conventional designs. The LWA unit cell is designed with a self-matching mechanism that suppresses the stop-band characteristics of periodic structures, thereby enabling seamless frequency scanning from the backward to forward region through broadside. Self-matching refers to the passive matching of the device itself through its own geometrical parameters, without the need for external matching networks. 2) Its application to broadband spectrum analysis at mm-wave frequency bands. In this work, a detailed design analysis of a side-fire antenna and its application to spectrum analysis is presented along with full experimental demonstrations around the 60 GHz band. The system is the first of its kind at mm-wave frequencies, and is fully integrated inside a low profile parallel-plate waveguide (PPW) with no radiation to free-space, eliminating any electromagnetic interference (EMI) with neighbouring instruments. The system is further designed to operate in the near-field region by introducing a curved shape of the structure resulting in compact size and low-profile compatible with a standard PCB process, where latencies can be expected to be on the order of milliseconds [35].

This chapter is organized as follows. Sec. 3.2 describes the principle of the proposed side-fire leaky-wave antenna inspired from a cascaded matched T-junction cell with an asymmetric notch (or inductive post in fabrication). Through cascading the T-junctions, the output ports can be opened for radiation forming a periodic LWA. Detailed eigenmode analysis is further provided to highlight important features of the structure. Sec. 3.3 shows how the side-fire LWA can be used as a real-time spectrum analyzer through the property of spatial-spectral decomposition. By introducing curvature, a compact size for the spectrum analyzer can be achieved. Furthermore, a simple analytical method is used to model the beam-scanning property of this structure providing further insight. Sec. 3.4 provides experimental results showing the fabricated structures and the corresponding measurement results demonstrating the successful operation of the side-fire LWA

for spectral decomposition. Finally, conclusions are provided in Sec. 3.5.

3.2 Proposed Side-Fire Leaky-Wave Antenna (LWA)

LWAs can be considered as series-fed phased arrays, and are typically broadband and directive in nature. While α (and the physical aperture) controls the beamwidth of radiation, β provides the necessary frequency dependent phase shifts to scan the radiation beams. The Majority of LWAs operate as interference-based devices, and are periodic in nature with sub-wavelength periodicities, where they radiate from the $n = -1$ space harmonics which fall within the fast-wave region with both backward and forward radiation [36, 17, 37]. On the other hand, uniform LWA structures based on metamaterial principles enable full-space frequency scanning using fundamental mode directly [38, 39].

The majority of LWAs provide frequency-scanned peak radiation normal (in the $y - z$ plane) to the plane of the antenna (say $x - y$ plane, with antenna along y -axis for instance). An exception is a rectangular waveguide based LWA, where a continuous slot is opened enabling peak radiation from the sides, in the $x - y$ plane [40]. Such a structure however is restricted to operate in the forward region only with no broadside radiation capability. Now, our objective here is to design a LWA, which is integrated into the substrate, and radiates in the $x - y$ plane with radiation in *both forward and backward regions including broadside*. In this section, we will present a step by step procedure to show how such a structure may be devised starting from a conventional waveguide T-junction.

3.2.1 Conventional Waveguide T-Junctions

The microwave and mm-wave T-junction is an important and commonly used component in microstrip circuits such as filters, amplifiers, ring hybrids, and power dividers [41]. Depending on the symmetry of the junction, power can be split in various ways. The S matrix can be written as

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} \\ S_{21} & S_{22} & S_{23} \\ S_{31} & S_{32} & S_{33} \end{bmatrix}. \quad (3.1)$$

An ideal conventional matched power splitter has $|S_{11}| = |S_{22}| = |S_{33}| = 0$, and $|S_{12}| = |S_{13}| = |S_{23}| = 1/2$, with reciprocity [41]. A simple waveguide T-junction operating in its fundamental TE_{10} mode with equal power split is shown in Fig. 3.1(a), with its typical response shown in Fig. 3.1(b) exhibiting strong reflection characteristics. This impedance mismatch can be improved through the introduction of a notch located in an optimized region of the junction, acting as a matching element [42, 43]. Fig. 3.1(b) along with its typical response is shown in Fig. 3.1(c) with a significantly improved broadband matching. In actual fabrication, this notch is manifested as an inductive post.

Let us switch the input and output ports of a conventional T-junction, so that the input port now becomes an output port, and one of the output ports becomes the new input, as shown in Fig. 3.1(a), indicated with *Port-Switched Simple T-junction*, with its corresponding response shown in Fig. 3.1(d). Naturally, the power balance is disturbed between the two output ports and poor matching is still present. Next, let us introduce the notch to match this section, where by it is seen that by an optimum design and placement of the notch, an optimal matching is achieved as shown in Fig. 3.1(e), i.e. *Port-Switched Matched T-junction*. This configuration must be compared with a matched T-junction, where the notch was placed symmetrically with respect to the two output ports. For a Port-Switched Matched T-junction, a longitudinal offset s is needed which introduces a *transverse* asymmetry in the structure, where the transverse plane is $y = 0$. Now by controlling the width of Port 3, power flow in this section can be controlled. Figure 3.2 shows the sensitivity of the matching effect of this notch for the cascaded T-junction at 60 GHz, which is the desired design frequency for the side-fire antenna.

3.2.2 Side-Fire LWA

The port-switched matched T-junction now acts as a power splitter by splitting the input from port 1 to output ports 2 and 3, respectively, with identical port impedances at port 1 and 2. This represents a fundamental building unit which is self-matched due to optimal design of the notch element. This allows a series cascade of this unit to form an N element periodic structure with $N+2$ ports and a spatial period p , which can now be seen simply as a $1 : N$ power divider, as illustrated in Fig. 3.1(a). Since each unit cell element is self-matched using a transversally asymmetric placed notch, the resulting periodic structure also exhibits a broadband matching, whereby its stopband characteristics can be easily controlled. The typical reflection and transmission characteristics for $1 : 5$ power divider are shown in Fig. 3.1(f), where a broadband matching is clearly evident. Due to the periodicity and the traveling-wave nature of this structure, the phase of the output ports are frequency dependent. For instance, when $p \approx \lambda_g$, all the wavefronts that exit these ports are in phase, as shown in the Fig. 3.1(f). We will refer to this frequency f_0 as the *broadside frequency of the structure*. For frequencies, where $p > \lambda_g$ ($f > f_0$), there is a progressive phase delay leading to a phase tilt, which will later be shown to correspond to *forward radiation*. Finally, for $p < \lambda_g$ ($f < f_0$), there is a progressive phase advance corresponding to *backward radiation*. These wavefront advances and delays leading to phase tilts are further shown in Fig. 3.1(f) for two specific frequencies.

The cascaded $1 : N$ power divider can now be allowed to have its outputs open to free space readily forming a periodic LWA with beam-scanning capability. Thanks to the self-matched transversally asymmetric unit cell based on a notch, a seamless transition from the backward to the forward region is possible including broadside radiation. This resulting structure represents a periodic LWA consisting of N unit cells with radiating apertures, allowing free-space radiation from the side walls of a rectangular waveguide operating in its fundamental TE_{10} mode. Since this structure

Figure 3.1 Principle of constructing an N -port power divider starting from a conventional 3-port waveguide T-junction. Various design parameters are: $s = 12$ mm, $w_1 = w_2 = 42$ mm, $p = 63.32$ mm, notch length $\ell = 20$ mm, and notch width $g = 2$ mm. The fields are plotted at $f = 4$ GHz for (b-e).

is compatible with SIW technology, it can now be readily integrated inside a dielectric slab, where the radiation from the aperture array can either be coupled into free-space or confined between top and bottom conductors of the dielectric forming a PPW, as shown in Fig. 3.3(a). If the leaky-wave region outside the main waveguide is confined inside a PPW, the entire structure remains shielded and completely enclosed.

A cell for such a periodic antenna has a stop band at broadside that requires suppression [44]. This behavior can be examined through considering the eigenmodes of the unit cell [45, 46]. The cell schematic is shown in Fig. 3.3(a). Depending on the dimensions and location of the notch, various dispersion relations of the unit cell may be computed, depending on the symmetries of the unit

Figure 3.2 Sensitivity of matching notch of cascaded T-junction at 60 GHz with $w_1 = w_2 = 2.45$ mm and $p = 3.7$ mm. (a) Sensitivity of s for $g = 0.12$ mm and $\ell = 1.2$ mm. (b) Sensitivity of ℓ for $g = 0.12$ mm and $s = 0.7$ mm. (c) Sensitivity of g for $\ell = 1.2$ mm and $s = 0.7$ mm.

cell [46, 47]. The eigenmode analysis assumes Floquet (periodic) boundaries on the two faces of the unit cell along the direction of wave propagation. Let us consider the case, when the leaky-wave region is confined inside a PPW, which is terminated with the medium impedance at the far end of the waveguide. While the width of the main waveguide, a , the period p , the aperture size t and width w_1 are kept constant, the length ℓ_n and offset s of the notch is varied to study its traveling wave characteristics. The FEM-HFSS computed eigenmodes for various cases of notch configurations is shown in Fig. 3.3(b), which shows regions of $\beta < 0$ corresponding to backward radiation, and $\beta > 0$ for forward radiation (for positively sloped branches of the dispersion curves). As expected from the works of Otto et. al. [46, 47], a perfectly closed stopband requires an optimal transversal asymmetry in the unit cell. For sub-optimal asymmetries a large stop-band is present, which prohibits any broadside radiation along $\theta = 0^\circ$. A fully symmetric unit cell on the other exhibits a frequency balanced condition, but is still a non-optimal condition which is manifested in the flat phase region around the broadside frequency. Ultimately, at the optimal set of notch parameters, the eigenmodes crossover at a single point as shown in shown in Fig. 3.3(b), and the reflection at this frequency corresponding to broadside is expected to be optimal [48].

To further confirm these eigenmode results, a driven model is analyzed in FEM-HFSS consisting of $N = 12$ cells, whose computed frequency dependent transmission/reflection response and far-field peak gain is shown in Fig. 3.3(c). As expected, the cell with the optimal asymmetry provides an excellent match throughout the frequency band compared to a suboptimally asymmetric case, while the symmetric unit cell still exhibits a noticeable increase in reflection around broadside in spite of an overall good matching response [47]. Fig. 3.3(d) shows the radiating wavefronts at three different frequencies, clearly showing the wave propagation along backward, broadside and forward directions. Therefore, the proposed side-fire structure represents a fully integrated periodic LWA with complete control over its dispersion relation and the subsequent closure of the stop-band to enable broadside radiation. Exploiting its integrated nature and fully shielded configuration, we will next describe an integrated analog spectrum analyzer for a 60 GHz mm-

Figure 3.3 The proposed side-fire Leaky-Wave Antenna (LWA) constructed using the N -port power divider of Fig. 3.1. a) A driven-mode LWA structure consisting of 12 cells and the unit cell used for eigenmode analysis in FEM-HFSS, b) the corresponding eigenfrequencies showing three unit cell configurations. c) The driven mode S-parameters and the peak Gain vs frequency. d) Near-fields E_z showing the radiated wavefronts in the backward and forward regions, along with that at broadside. Here $p = 64$ mm, $t = 17.78$ mm, $a = 42$ mm, $w_1 = 5.08$ mm, and $w_2 = 152.4$ mm. For optimal asymmetry: $\ell_n = 8.64$ mm, and $s = 0.32$ mm. For full symmetry: $\ell_n = 8.76$ mm, and $s = 0$ mm. For suboptimal asymmetry: $\ell_n = 2.54$ mm, and $s = 0.34$ mm.

Wave frequency band.

3.3 Proposed mm-Wave Spectrum Analyzer

3.3.1 Principle

Leaky-wave antennas have the property of frequency scanning. For a periodic leaky-wave antenna, the aperture spacing provides a phase shift between locations of radiation. The electrical spacing between the apertures is frequency dependent, which provides a phase gradient that enables beam steering. The beam scanning law from eq. (2.1) is applicable in this case. A real-time

spectrum analyzer can be formed using the process of spatial-spectral frequency decomposition, similar to that used in optical prisms [29, 30]. Accordingly, a broadband time domain test signal sent at the input port of a LWA would have its frequency components radiated at particular angles described by (??) with appropriate magnitudes. This enables one to deduce the presence of the frequency components of the signal and their respective magnitudes by inspecting the corresponding far-field radiation patterns. Such a determination of the spectral components of an unknown test signal based on spatial-spectral frequency decomposition principle, is purely analog and thus real-time in nature.

The radiated signals from the LWA are typically measured in the far-field regime of the radiation, which is approximately located at distances greater than $2d^2/\lambda_0$, where d is the largest dimension of the antenna aperture, and λ_0 is the free space wavelength. For typical mm-wave leaky-wave antennas which are several wavelengths long, this distance readily exceeds 5 m when operated around 60 GHz. Therefore, to perform spectral analysis using a typical mm-wave LWA, one would need a large space to make the measurement. However, this measurement would not be protected from the environment, and thus undesired signal interference would occur between neighbouring devices. Thus shielding is necessary, which makes the proposed side-fire LWA configuration of Fig. 3.3 particularly attractive. However, the far-field requirement of large system size is still a challenge, since for far-field distances of several meters, the PCB would be impractically large. A potential solution is then to add a physical curvature to the LWA aperture to focus the far-field radiation in the near-field region of the antenna, which is proposed next.

3.3.2 Curved Side-Fire LWA

A solution to the issue of large far-field distances where the beam is found is by introducing a curvature to the antenna [49]. In the case of parabolic reflectors, a parabolic surface allows for a beam focal point where the radiation is concentrated. Depending on the exact surface curvature, the focal location can change, subject to sufficiently small aberrations. [50, 19]. In this sense a convex lens with a close focal point can solve the far distance problem. This principle can be easily applied to the side-fire LWA configuration of Fig. 3.3, where instead of a linear array, we devise a curved array with desired frequency dependent beam focussing characteristics in its near-fields. Such a structure may behave like a lens, allowing the beam location to be within the range of a typical PCB lengths, while preserving the spatial-spectral decomposition information needed for spectral analysis.

With a determined antenna structure and periodic aperture producing a specified per unit length propagation constant $\beta(\omega)$, a numerical model to calculate the radiation and focussing properties is needed to understand and model this structure. Consider a uniform circular LWA as shown in Fig. 3.5(a), of an angular span 2α and a radius r , excited from the left. The field variation along the z -axis is assumed to be zero, i.e. a 2D problem is assumed. Each point on this LWA can be assumed to be a line source $J_z(x', y'; \omega)$ which is radiating with a certain phase accumulated along

Figure 3.4 Integrated Side-Fire LWA design. (a) FEM-HFSS cell model of the curved LWA based on substrate integration, showing the unit cell design parameters. Here $d_1 = 0.34$ mm, $d_2 = 0.34$ mm, $v = 0.7$ mm, $s_3 = 1.08$ mm, $\ell_1 = 3.46$ mm, $\ell_2 = 0.8$ mm, $\epsilon_r = 2.2$ mm, $s_1 = 0.7$ mm, and $s_2 = 1.08$ mm. The number of cells is 72, with a total antenna length of 293 mm.

(b) S_{11} and (c) S_{21} of the curved antenna with optimal via $h = 0.1$ mm compared to suboptimal fully symmetric via $h = 0$ mm.

the LWA. Each source coordinate follows the following parametric curve

$$x'^2 + y'^2 = r^2. \quad (3.2)$$

The total radiated electric field in the $x - y$ plane can be analytically expressed as [31]

$$\mathbf{E}(x, y) = \int_{-r \sin \alpha}^{r \sin \alpha} J_z(x', y') G(x', y'; x, y) dx' \hat{z} \quad (3.3)$$

where $G(x', y'; x, y)$ is the Green's function. This integral represents the contribution of each source at the observation point (x, y) along the entire extent of the antenna. For a 2D problem we have a

Hankel's function of the second kind, given by

$$G(x', y'; x, y) = -\frac{\omega\mu_0}{4} H_0^{(2)}(k\rho), \quad (3.4)$$

with

$$\rho(x', y'; x, y) = \sqrt{(x - x')^2 + (y - y')^2} \hat{\rho}, \quad (3.5)$$

where ρ represents the vector from the source to the observation point, as illustrated in Fig. 3.5(a). Finally, the position dependent current source is assumed to take the following form:

$$J_z(x', y') \approx e^{-j\beta s} \quad (3.6)$$

where the length $s = r\theta$ represents the arc length from $\{(x_1, y_1) - (x', y')\}$ and can be easily shown to be

$$s = r \cos^{-1} \left\{ 1 - \frac{(x' - x_1)^2 + (y' - y_1)^2}{2r^2} \right\} \quad (3.7)$$

Therefore, substituting (3.4), (3.6) inside (3.3) leads to the analytical form of the field anywhere in the observation area.

Next, an integrated spectrum analyzer based on a side-fire LWA configuration of Fig. 3.3 with a 1 GHz spatial resolution operating from 57-64 GHz was designed and simulated in FEM-HFSS, as shown in Fig. 3.4(a). To be compatible with standard PCB fabrication, the vertical solid walls of Fig. 3.3 are realized using a metallized via array as typically done in SIW implementations, and a curvature is introduced. An isolated via is used to emulate a notch and its precise placement is obtained to close the stop-band following the analysis of Fig. 3.3 to obtain the required asymmetry. Fig. 3.4(b,c), further shows the simulated S-parameters for the case of an optimal and suboptimal asymmetry, clearly showing broadband matching within the bandwidth of interest, i.e. 57-64 GHz.

Next, an analytical model is built for this side-fire configuration, where its propagation constant $\beta(\omega)$ corresponding to the $n = -1$ space harmonic is approximated as

$$\beta(\omega) \approx \sqrt{\left(\frac{\omega}{c}\right)^2 \epsilon_r - \left(\frac{\pi}{a}\right)^2} - \frac{2\pi}{p}. \quad (3.8)$$

This analytical result closely approximates the full-wave computed dispersion relation extracted from an eigenmode simulation of a single cell, and is thus used due to its simplicity to estimate beam-scanning of the structure in the near-field region of the antenna. Using (3.4), (3.6) and (3.8) in (3.3), the radiated near-fields were calculated along a line at the focal distance r , and compared with those obtained using the full-wave HFSS model, as shown in Fig. 3.5(b-d) for several frequencies. An excellent match is seen for all frequencies which successfully validates the model. While the HFSS model is rigorously correct, it is electrically large and computationally expensive. The analytical model on the other hand, captures the fundamental behavior of the structure using a

much simpler model based on array of line sources, while providing a reasonably good accuracy. It thus represents an ideal method for an efficient design and fast analysis of a side-fire LWA based spectrum analyzer.

Figure 3.5 Analytical method to compute the near-fields of a side-fire antenna. a) Illustration of a curved side-fire antenna showing the source and observation points. (b)-(e) Comparison of electric fields calculated along a line at the focal point using the wavenumber computed from FEM-HFSS eigenmode analysis with full-wave electric fields along a line at the focal distance ($r = 56$ cm).

3.4 Experimental Demonstration

In this section, two experimental demonstrations will be presented: 1) A straight integrated side-fire LWA radiating into free space and, b) a curved side-fire antenna radiating inside an integrated PPW acting as a mm-wave spectrum analyzer.

3.4.1 Straight Side-Fire LWA

Fig. 3.7(a) shows a photograph of a 15 cell straight side-fire LWA operating around the 60 GHz band, following the configuration of Fig. 3.4(a), except with a straight cell. The matching via is optimized in simulations to close the stop-band using the nominal permittivity of the dielectric

Figure 3.6 Procedure to determine the radius of curvature of the curved side-fire LWA and the arc length 2α to achieve a frequency resolution of 1 GHz within the operating bandwidth of 57 GHz - 64 GHz. (a) Beam scan variation with r at $f = 60.7$ GHz. b) Beamwidth as a function of angular span of the LWA for $r = 56$ cm.

specified by the manufacturer. A transition is further designed to excite the LWA antenna using a standard WR-15 waveguide, as shown in the input and output ends of the LWA structure. One side of the antenna is left open to free-space with a width of the PPW controlling the amount of radiation along the $-y$ direction. Fig. 3.7(b-c) shows the measured S-parameters compared with FEM-HFSS. While good matching is observed in measurements as a result of optimal via placement, a slight increase in the reflection is also observed along with appreciable mismatch with the simulated response. This mismatch between simulation and measurement can be attributed to the unknown material properties at 60 GHz, and the tolerances in fabrication, especially considering that the response is particularly sensitive to the matching via location.

To estimate the material properties, a straight waveguide resonator was also fabricated, which was then used to estimate the effective dielectric constants of the Arlon DiClad 880 material used in fabrication. The resonator results are presented in the Appendix, and the extracted dielectric permittivity and loss tangent are found to be ≈ 2.08 and 0.003 , respectively, compared to the design values of $\epsilon_r = 2.2$ and $\tan \delta = 0.0009$. The fabricated straight side-fire LWA was then re-simulated with the new material parameters, and the new S-parameters are shown in Fig. 3.7(b-c), improving the agreement with the measurements. Importantly, a frequency shift of almost 1.6 GHz is observed, with a new broadside frequency of $f_0 = 60.7$ GHz (observed in HFSS with peak radiation at $\theta = 0^\circ$). The resulting frequency dependent gain and the simulated radiation patterns are shown in Fig. 3.7(d-e) showing the seamless radiation from backward to forward regions including broadside at 60.7 GHz, as expected. Due to an actual higher loss tangent of the dielectric, the gain is seen to be significantly dropped with a slight drop at broadside, while maintaining a monotonically increasing trend, as expected.

Figure 3.7 Experimental demonstration of a straight side-fire LWA radiating into free-space. (a) Fabricated antenna prototype ($N = 15$ cells). (b) S_{11} of the antenna in FEM-HFSS with Arlon and the resonator characterized material compared to the VNA measurement. Here the stopband is closed at 61 GHz (broadside). (c) S_{21} of the antenna in FEM-HFSS with Arlon and the resonator characterized material compared to the VNA measurement. (d) Simulated gain as a function of frequency for the antenna with the characterized material and Arlon DiClad 880. (e) Radiation patterns of the simulated antenna with characterized material in 1 GHz steps over the band of interest.

3.4.2 Integrated Spectrum Analyzer

Next, an integrated spectrum analyzer based on a side-fire LWA will be demonstrated. Fig. 3.8(a) shows the general schematic of the proposed system. A curved side-fire LWA of Fig. 3.4 with larger

Figure 3.8 FEM-HFSS modeling of the integrated side-fire LWA based spectrum analyzer. (a) Schematic model showing an array of integrated horn receivers with $\Delta x = 570$ mm and $\Delta y = 323$ mm, and the curved side-fire LWA consisting of $N = 72$ unit cells. (b) Interpolated transmission into the receiving horn ports. (c) Normalized $|E_z|$ plots from HFSS of the spectrum analyzer showing the beam focusing. All horns and aperture dimensions shown in Tab. 1.

number of cells and input/output transitions to standard WR-15 waveguides is placed on the far right of the dielectric layer, where Port 1 is intended as the input to the system. The leakage from the side apertures is maintained inside the PPW structure, due to the beam-scanning property, the spatial-spectral decomposition is achieved in the $x - y$ plane, with broadside along $-x$ and $x < 0$ and $x > 0$ as the backward and forward radiation regions, respectively. The system is designed to operate between 57-64 GHz, with a frequency resolution of 1 GHz.

To capture the power radiated along different angles, an integrated horn array is chosen for the sake of a proof of concept experiment (detailed dimensions are provided in Tab. I). To achieve a 1 GHz resolution over the 57-64 GHz band, 8 horns for each 1 GHz frequency step are needed where the transmission in the form of S-parameters can be measured. The receivers are waveguide fed, and thus must be spaced to allow room for WR-15 flanges, which have a diameter of 19.05 mm. The unknown physical dimensions to achieve this resolution are: the horn spacing, radius of curvature and the angular span of the side-fire antenna. To determine these parameters, a simple

Figure 3.9 Dispersion diagram of the conformal leaky-wave antenna illustrating the issue of multi-beam radiation within the fast-wave region.

parametric analysis of Fig. 3.5(a) is performed. Fig. 3.6(a) shows the maximum scan range (using peak amplitudes of the focussed beam) over 57-64 GHz for each r at a fixed frequency f , for instance. Knowing the required separation between the horn receivers of ≈ 12 mm, r is set at 56 cm. Next, Fig. 3.6(b) shows the proper antenna arc length for a given radius of 56 cm to produce a sufficiently narrow beam, less than the 3-dB beamwidth of the horn. It was then set at $2\alpha = 30^\circ$. This would then allow 1 GHz resolution while preventing electromagnetic coupling to neighbouring receivers which could compromise the spectral decomposition.

Fig. 3.8(b) shows the FEM-HFSS simulated power transmission between the input port 1 and the various horn receiver ports. A linear frequency scanning can be seen across the ports, with clear separation between the power received by adjacent ports. Direct incidence of the beam on a port results in a transmission of more than -15 dB. Neighbouring ports in such a case would have less than -30 dB in transmission. To this end, a nominal frequency resolution of 1 GHz is readily shown. Fig. 3.8(c) further shows the near-field distribution inside the PPW for various frequencies, which are being received by different horn receivers demonstrating the frequency scanning behavior of the side-fire antenna. In addition, extra radiation beams are also observed in the forward region, which are analogous to grating lobes in regular antenna arrays.

To understand the nature of these grating lobes, the dispersion diagram of the side-fire LWA must be inspected as shown in Fig. 3.9. For the designed substrate permittivity and antenna dimensions (curvature, width, aperture period), more than one fast-wave harmonic appears in the radiative region. Since radiation is in the substrate, the slope of the light line is increased by a factor of $1/\sqrt{\epsilon_r}$ compared to that in free-space. As a result, the fundamental mode does not leave the radiative region since it is parallel to the light line. Therefore, the fundamental mode *always* remain inside the fast-wave region and radiates in the forward direction. At the same time, other spatial harmonics, particularly $n = \{-1, -2\}$ harmonics, also radiate within the band of interest [39] [36]. They

Table 3.1 Port and horn dimensions.

Dimen.	Port 1	Port 2	Port 3	Port 4	Port 5	Port 6	Port 7	Port 8	Port 9	Port 10
a (mm)	3.759	3.759	3.759	3.759	3.759	3.759	3.759	3.759	3.759	3.759
b (mm)	1.88	1.88	1.88	1.88	1.88	1.88	1.88	1.88	1.88	1.88
a_{ap} (mm)	3.302	3.302	3.297	3.297	3.297	3.297	3.297	3.27	3.27	3.27
L_1 (mm)	7.807	7.807	7.807	7.807	7.807	7.807	7.807	7.807	7.807	7.807
L_2 (mm)	5	5	5	25	5	25	5	25	5	25
L_3 (mm)	0	0	9.6	9	9.3	9	9.6	9.3	9.3	9
L_4 (mm)	4.284	4.284	4.284	4.284	4.284	4.284	4.284	4.284	4.284	4.284
g (mm)	2.81	2.81	2.81	2.81	2.81	2.81	2.81	2.81	2.81	2.81
θ_h ($^\circ$)	0	0	56	52	40	50	42	42	37	46

appear as extra beams in the full-wave simulated fields of Fig. 3.8(c), where $n = -1$ are the ones captured by the horn receivers. While these spurious grating lobes represent unwanted power loss through radiation, the dominant power is still found to be contained in the $n = -1$ harmonic sufficient for the current purpose of spectrum measurement.

The side-fire based integrated spectrum analyzer is next fabricated and some pictures of the prototype and the experimental setup is shown in Fig. 3.10(a-b). Fig. 3.10(a) shows the array of integrated horns on the receiver side connected to a standard WR-15 waveguide to a V-band coax cable. The integrated spectrum analyzer is next measured using a Vector Network Analyzer as shown in Fig. 3.10(b), where different measurements are taken across various horn ports. The S-parameters of the curved side-fire antenna is shown in Fig. 3.10(c,d) where in spite of an intended well-matched design in simulations, relatively larger reflections are observed, particularly around 62 GHz. This indicates a stop-band region and a high frequency shift in the response, as the broadside frequency was around 60 GHz. Transmission on the other hand stays around -15 dB. The spectrum analyzer was then re-simulated in HFSS using the newly characterized parameters of the dielectrics (provided in Appendix), and the resulting S-parameters are shown also in Fig. 3.10(c,d). The frequency shift is clearly captured as seen in S_{11} and an increase in reflection is observed in spite of being practically acceptable. However an appreciable difference between measurements and simulations still remains. This suggests that small variations in tolerances and material properties may significantly accumulate to give rise to large errors, since this integrated spectrum analyzer is still electrically large to the order of about 120λ . Fig. 3.11 shows the S param-

Figure 3.10 Experimental setup and results of the fabricated spectrum analyzer. (a) Zoomed in view of the receiver horn array during S-parameter measurement. (b) S-Parameter measurement of the VNA. (c) Measured S_{11} and (d) S_{21} of the spectrum analyzer compared to simulation with Arlon DiClad 880, and in simulation fitted material parameters (Appendix A).

eter sensitivity of the conformal antenna as a function of the matching via's diameter d_1 , horizontal offset from the centre h , and vertical offset from the centre of the back slot v . In all cases, it is apparent that the stopband around 62 GHz is quite sensitive to the via location and size. Based on the photomicrographs of Fig. 3.11 (a) and (b), variations on the order of 50 - 100 μm have been fabricated, resulting in a measured stopband. Next, the transmission across various horn ports receivers from the input LWA port is measured and are shown in Fig. 3.12. A clear shift in the transmission peak is observed across frequency as different horn receivers are measured, confirming the frequency scanning behaviour. There is a systematic shift in the measured peak compared to simulation with Arlon as the material over each horn receiver. There is also a drop in transmission. This is attributed to the lower permittivity of the actual material, as well as a higher than simulated loss tangent. The measurements results are next compared with the characterized material (appendix) response in FEM-HFSS and a very good agreement is observed where not only the frequency shift but the transmission levels are correctly reproduced. The spectrum analyzer thus operate between 59 GHz - 66 GHz (instead of 57 GHz - 64 GHz), with a resolution of approx-

Figure 3.11 S Parameter Sensitivity of the conformal antenna with characterized material. (a), (b) Photomicrographs of two cells showing notch dimensions and locations. (c), (f) Sensitivity of d_1 for $h = 0.1$ mm and $v = 0.7$ mm. (d), (g) Sensitivity of v for $h = 0.1$ mm and $d_1 = 0.34$ mm. (e), (h) Sensitivity of h for $v = 0.7$ mm and $d_1 = 0.34$ mm.

imately 1 GHz and an average transmission level of about -20 dB. The proposed side-fire LWA based integrated spectrum analyzer is thus successfully demonstrated.

3.5 Conclusions & Future Work

An analog, low-profile and shielded spectrum analyzer has been proposed for operation at mm-wave frequencies around the 60 GHz band based on a novel side-fire LWA configuration. The proposed side-fire LWA has been systematically developed from a conventional 3-port waveguide T-junction which has been modified to a LWA unit cell with an internal matching mechanism to

Figure 3.12 Measured S-parameters of the integrated spectrum analyzer of Fig. 3.10 demonstrating the spatial-spectral decomposition. Adjusted $\epsilon_r = 2.08$ from nominal $\epsilon_r = 2.2$ for Arlon DiClad 880. Adjusted $\tan \delta = 0.003$ from nominal $\tan \delta = 0.009$ for Arlon DiClad 880 following the material characterization given in the Appendix. (a)-(h) Correspond to ports 3-10, respectively.

suppress stop-band to enable broadside radiation. The internal mechanism is achieved by introducing a transversal asymmetry of the unit cell, by placing a notch inside the waveguide, which results in a seamless leaky-wave radiation from backward to forward region including broadside. The proposed antenna is fully compatible with SIW technology resulting in a low-profile and is compatible with standard PCB fabrication. The resulting periodic side-fire antenna radiates in the plane of the antenna, whereby the leakage power be either be allowed to radiate in free-space or kept confined inside a PPW structure. The proposed side-fire structure thus can be completely shielded, which is particularly attractive for devising an analog broadband spectrum analyzer. Such a system has been demonstrated around the 60 GHz band, where a convex side-fire antenna has been used to focus the radiated beams in the near-field of the structure to make the entire system more compact, operating between 59 - 66 GHz in experiments with a 1 GHz resolution. Furthermore, a simple mathematical model consisting of an array of point sources has also been proposed which is found to faithfully reproduce the beam-scanning characteristics of the curved side-fire LWA in the near-field, providing a fast design tool to engineer such a system.

The proposed side-fire antenna also provides a different interpretation of a LWA, where a periodic leaky-wave structure may be seen as a broadband multi-port power divider, whose outputs are connected to radiating apertures. The proposed spectrum analyzer furthermore offers several system features and benefits. The system is broadband in nature due to its wide-band radiating property of the stop-band suppressed side-fire LWA, directly operating at mm-wave frequencies based on SIW technology. It is a fast analog system minimizing digital computations compared to

a fully digital system, except at the data acquisition stage. The frequency resolution of the system depends on the length of the side-fire antenna. This resolution can further be improved by optimizing the length of the antenna and controlling the per-unit-length leakage from the unit cell, while keeping the overall size of the analyzer the same. The low-profile and compact size of the spectrum analyzer represents a bench-top type system that can be placed on standard equipment with no electromagnetic interference due to its fully-shielded configuration. Thus the proposed integrated side-fire antenna and the associated spectrum analyzer system represent an attractive solution for next generation 5G mm-wave systems in the area of wireless communication and instrumentation, for instance.

Chapter 4

OPTICALLY TRANSPARENT FREQUENCY SELECTIVE SURFACES (FSSs)

4.1 Introduction

In chapter 3, the benefits of designing 5G devices with EM shielding were highlighted. The proper operation of the spectrum analyzer was dependent on its ability to isolate its signal-under-test from the EM environment to prevent interference and spectral corruption. However, the need for EMI shielding is also present in other applications, particularly in free-space wireless communications, as opposed to integrated EM wave propagation in materials such as PCB substrates.

In the special case of an indoor environment, the use of EM shielding can be applied on the surface of an existing wall to minimize EM interference with adjacent wireless systems and enhancing wireless security by minimizing the leakage of radio waves outside designated areas. This can be readily achieved using spatial spectral filters known as *Frequency Selective Surfaces* (FSSs), that exhibit reflection or transmission across a given frequency band. They typically consist of a periodic array of unit cells whose geometrical properties control the filtering characteristics of the structure, with typical unit cell periods of $\approx \lambda/2$ [51]. It is an effective way to solve EMI problems and not be a source of pollution to that electromagnetic environment.

It is desired to design a *band-stop* FSS shielding design without degrading the functionality at their operating frequencies with less weight and cost while maintaining stability across wide angular range and polarizations (TE and TM) [52, 53, 54, 55]. The angular stability of the FSS is known to improve with smaller unit cell sizes, and with recent advances in EM metasurfaces, wide variety of novel sub-wavelength unit cell topologies have been proposed in the literature. Moreover, for deployment on transparent windows and walls, they are further required to exhibit significant optical transparencies. Potential solutions are FSS realizations on flexible and ultrathin transparent substrates such as Polyethylene terephthalate (PET) on which the metallic FSS patterns are screen printed using high conductivity inks with tolerances of printed electronics (PE)

technologies. While this significantly increases the optical transparency, the manufacturing cost is proportional to the amount of ink used and for large-scale metasurfaces, this may be economically prohibitive.

In light of these challenges, here we experimentally demonstrate a simple ink-minimization technique applied to a compact and a miniaturized cross-dipole/loop FSS (of unit cell periodicities $\approx \lambda/3$) to offer greater optical transparency by simultaneously reducing its manufacturing cost, while offering greater angular stability. Metallic ink minimization is a process in which a solid metallic surface is meshed using various perforations to realize microwave components and antennas, without compromising their electrical performance [56]. It has been shown, for instance, that the meshed pattern of planar antennas such as bow-tie, patch-dipole and monopole antennas had little effect on the performance compared to their solid metallic counterparts [57, 58]. We thus apply this technique of ink minimization to FSS unit cells and demonstrate the use of single-step screen printing as a simple fabrication tool for low-cost manufacturing of band-stop FSS shielding on low-cost flexible and optically transparent substrates for millimeter-wave applications at 28 GHz. Also given that such FSSs may be installed on conformal surfaces, we further exploit the flexible nature of these transparent substrates and present their spectral filtering characteristics by introducing finite spatial curvature.

This chapter is organized as follows: Sec. 4.2 presents the evolution of the conventional FSS unit cell to a more compact cross-dipole loaded meandered loop cell used later in this work, with increased angular stability and compact size. The screen printing process is further described along with the application of ink-minimization on the cell to reduce the amount of ink required to achieve the same response as that of a solid filled cells. The simultaneous benefit of increased transparency is further quantified and presented using a simple transmission model. Sec. 4.3 presents various FSS prototypes of solid and ink-minimized structures operating at the 5G 28 GHz frequency band, further showing the effect of minimum line-width employed in surface meshing process. Sec. 4.4 further explores the utilization of the FSS prototypes on conformal surfaces and its various transmission response is characterized in details. Finally, Sec. 4.5 presents the summary and possible future work.

4.2 Proposed Miniaturized FSS with Ink Minimization

We seek an FSS unit cell that offers stable angular response across large range of incoming angle of incidences. To achieve low sensitivity with respect to the incidence angle of the exciting wave, the screen size needs to be small. The various choices of the FSS unit cells will be discussed here next and it will be shown that a *cross-dipole loop structure* represents an excellent choice for applying the ink-minimization technique due to its greater angular stability and compact size, compared to typical conventional FSS unit cells.

4.2.1 Proposed Miniaturized Unit Cell

Typical FSS cells are conventional square loop unit cell elements, as shown in Fig. 4.1(a), which are composed of two half-wavelength dipoles that are connected to one another at each end, therefore, occupying large element sizes and large periodicities leading to problems, including difficulties in conforming to curved surfaces and dependence on the incidence angle of the incoming wave. In traditional designs, the frequency-selective properties result from mutual interactions of the elements [59]. Therefore, to observe a desired frequency selective behavior, a large number of unit cells must be present. Consequently, the overall size of the surface is ought to be electrically large in order to mimic an infinite frequency selective surface characteristics [60].

To address this problem, meandering of square loop frequency-selective surfaces are considered here as in [61][62], where the meandering of a single circular loop was shown to enable miniaturized cells for angular stability. In general, there are three methods to reduce the size of transmission-line circuits, which are: 1) Line folding; 2) lumped elements; and 3) distributed loading. Distributed loading is quite appropriate for millimeter-wave frequencies when used with distributed elements. This approach uses either distributed capacitive or inductive loading to reduce the size of footprint FSS element. A hybrid cross loop/dipole element is one such example which is comprised of standard cross loop and cross dipole elements, as shown in Fig. 4.1(b) along with its typical transmission response. It shows a smaller unit cell size compared to a square loop cell (18% reduction), with an improved angular stability ($\approx 27\%$ of the resonant frequency deviation), as also shown in Fig. 4.1(b). By introducing the cross dipole within a *meandered loop*, the overall area of the unit cell is further reduced by 20%, with an even greater angular stability (resonant frequency shift of 0.255 GHz vs 1.64 GHz for the square loop), as shown in Fig. 4.1(c).

All cells of Fig. 4.1 are equally capable of responding to TM as well as TE polarizations of the incident plane wave. All of the loop structures have nulls for both polarizations when the lengths of loads are approximately $\lambda/4$. When this occurs, the short circuit load appears at the cross loop input as an open circuit, effectively leading to no current flow. This happens for parallel as well as perpendicular polarizations as opposed to the modal interaction nulls in the center connected elements that occurs for only parallel polarization. This structure also provides opportunity for bandwidth control by changing the load impedance. This can be achieved by varying the separation between the arms of its constituent elements. A shorter (longer) distance decreases (increases) the bandwidth [63] [64].

We now wish to use this FSS unit cell to design a band-stop response at the 5G band of 28 GHz, and realize it using the low-cost screen printing process described next. In practical applications such as mm-wave imaging [65] and interference reduction in wireless indoor environments [66], FSS attenuations of < -20 dB are acceptable. Note that for 60 GHz operation this FSS cell could be scaled down appropriately, while having increased sensitivity to tolerances for its resonant frequency, with a similar Q factor at 28 GHz due to material conductivity. We further wish to apply the ink-minimization technique on this cell to further lower the cost, making it suitable for mass

Figure 4.1 Proposed Ink-minimized (meshed) FSS unit cell with miniaturization for 28 GHz operation. (a) Structure of a simple square loop cell and its typical transmission response as a function of the incidence angle (Resonant shift over the span is 1.64 GHz) (b) Structure of the conventional square loop cross dipole FSS cell, and its typical transmission response as a function of the incidence angle (Resonant shift over the span is 0.455 GHz) computed using Ansys FEM-HFSS using the Floquet unit cell boundaries. (c) Structure of the miniaturized meandered square loop cross dipole FSS cell showing improved angular stability (Resonant shift over the span is 0.245 GHz). (d) Structure of the meshed FSS cell with minimum $80 \mu\text{m}$ line-width and its typical transmission response under normal incidence. All cells are printed on PET of thickness 0.125 mm , $\epsilon_r = 3$ and metal thickness of $5 \mu\text{m}$.

manufacturing of large-scale surfaces.

4.2.2 Screen Printed Technology

The creation of conductive paths in electronic systems has traditionally involved the addition and subtraction of conductive material by means of deposition, mask, and etch processing. Printing techniques, such as screen printing has been proven as viable alternative methods for creating conductive paths and have the additional benefit of creating these paths in a simple two-step process involving only printing and curing. The deposition of Silver (Ag) nanoparticles during metallization has been deemed favorable over other metals on the account of its unique properties, such as chemical inertness, excellent electrical conductivity, and ease of ink formulation [67] [68].

For greater optical transparency, PET film was used along with Ag-nanoparticle-based conducting ink, where the silver ink was further formulated to meet the viscosity and surface tension using ethylene glycol. The process is relatively simple, cost-effective, and versatile, allowing for inexpensive printing of FSS patterns on large surfaces. The screen plate (mask), which is a stretchy fine lattice, was used to imprint the pattern on the substrate. Ag ink was placed on top of the screen with a squeegee bar to fill it with ink. The movement of the squeegee with pressure enabled the Ag

ink to pass through the pattern to create the Ag pattern designed on the PET substrate. The resulting Ag printed films were oven dried to remove excess solvent and material impurities [69, 70, 71]. A difficulty associated with screen printed technology, particularly at higher 5G frequencies (28 GHz, 60 GHz) is ink spreading. When the Ag ink is pressed onto the substrate, uneven distribution of pressure causes the ink to spread nonuniformly, perturbing the structure dimensions to the order of about 10 μm . This perturbation can contribute to resonance shifts in FSSs. A further difficulty is in the conductivity of the printed ink, which is lower than bulk silver (6.3 MS/m) resulting in reduced Q factor. A substantial cost of the process is the ink, whose amount and cost scale up significantly with electrically large FSS print sizes. We can reduce this cost by applying meshing solid metallic surfaces as a part of the ink-minimization process.

4.2.3 Ink Minimization for Increased Optical Transparency

A simple technique to reduce the amount of ink required to realize a given metallic pattern without compromising its electrical performance is using the process of *ink minimization*. Ink Minimization involves the *meshing* of the metallic component of a solid electromagnetic structures to produce a structure with similar performance with a large reduction in the used material. The principle is similar to perforated metallic sheets which are used in wide variety of EMI/EMC problems, with classical usages in Faraday Cages, for instance. As long as the perforations are electrically small, electromagnetic waves with large wavelengths cannot penetrate the surface. The perforated surfaces simply act as equivalent solid surfaces with high shielding effectiveness while improving optical visibility through them.

This motivates the application of ink minimization to our FSS unit cell of Fig. 4.1(b), as the goal is to minimize the metallic area (and thus the ink) to reduce cost and increase optical transparency. Fig. 4.1(c) shows an example of an FSS unit cell whose various metallic traces are meshed using square shaped perforations. The amount of meshing possible simply depends on the allowed tolerance on a printable line-width using the screen printing process, where 80 μm is shown here representing a typical line-width tolerance. While there could be variations in the exact shape and distribution of the mesh, a simple square shaped mesh with straight edges is chosen here for simplicity to avoid possible complications with fabricating curved shapes and holes. Fig. 4.1(f) shows the typical band-stop filtering response compared to that of a solid cell of Fig. 4.1(b). With slight adjustments in various dimensions of the cell and the unit cell period, nearly identical frequency response is observed with no change in the resonant frequency. Moreover, the ink-minimized cell maintained very good angular stability up to $\pm 60^\circ$ range.

The optical transparency of the surface is determined dominantly by the surface area of the ink in relation to unit cell period, and is defined here as the ratio of the transmitted power to the incident power (assumed unity). This can be simply expressed for an approximate estimation, as:

$$\mathcal{T}(\theta, \omega) = \left(1 - \frac{a}{\Lambda^2}\right) \tau^2(\theta, \omega), \quad (4.1)$$

where θ is the viewing angle and τ is the frequency dependent field transmission coefficient of a dielectric sheet using the Fresnel coefficients, Λ is the FSS unit cell period and a is the total surface area of the Ag ink. With a PET dielectric substrate of permittivity $\epsilon_r = 3.0$ and thickness $d = 0.125$ mm at $\lambda = 560$ nm (in the visible EM spectrum), the standard solid FSS cell of Fig. 4.1(b) has a $\mathcal{T}$ of 0.228. On the other hand, the proposed ink minimized FSS cell of Fig. 4.1(c) with an $80 \mu\text{m}$ tolerance has a transparency of 0.377, which represents an increase of 0.149, i.e. increase in optical transparency by 65%. Fig. 4.2 further shows the optical transmission for various viewing angles and metallization area ratios obtained using (4.1). For viewing angles approaching 90° , transparency approaches 0 since the FSS is not being directly viewed. For metallization areas approaching unity, the transparency also approaches 0 since the cells become dominantly metallic and cover the transparent PET underneath. Smaller metallization area ratios clearly provide increased transparency for all viewing angles.

Figure 4.2 Approximate optical transparency of the proposed ink-minimized FSS for different metallization area ratios a/Λ^2 and viewing angles θ , computed using (4.1).

4.3 Screen Printed FSS Prototypes & Measurement Results

4.3.1 Ink-minimized vs Solid FSS

The proposed ink minimized FSS unit cell is next fabricated using the screen printing process for a chosen line-width of $80 \mu\text{m}$, and the resulting prototypes are shown in Fig. 4.3(a-b), corresponding to a solid conventional FSS of Fig. 4.1(b) and (c), respectively. The prototypes are next characterized using the quasi-optical measurement setup of Fig. 4.3(c). The FSSs were next tested by measuring frequency-dependent transmission S_{21} with a VNA over the 22 - 36 GHz band. The FSSs were excited with pyramidal horns antennas and focused with dielectric lenses for achieving approximate plane wave incidence conditions [25].

Fig. 4.3(a-b) shows the transmission responses of both solid and ink-minimized cells subject to different angle of incidences in experiments and compared to full-wave simulations. In each case there are fabrication errors due to ink spreading as shown by the measured dimensions of the

Figure 4.3 Fabricated FSS prototypes and their transmission characterization. (a) Photomicrograph of solid cell and the corresponding measured transmission response as a function of incidence angle. (b) Photomicrograph of $80\ \mu\text{m}$ line-width ink-minimized FSS cell and its measured response. Conductivity characterized as $\sigma = 600000\ \text{S/m}$. (c) Photograph of the quasi-optical measurement system used to characterize the FSS prototypes [2].

cells when compared to their simulation models. This ink spreading, combined with variable ink conductivity results in resonant frequency shifts and reduced Q-factors, compared to simulation as shown in Fig. 4.3(a-b). However, the ink-minimized FSS maintained a good angular stability with good band-stop characteristics compared to the solid one. A large deviation in simulation compared to experiment is the reduced Q-factor for the various structures. This is largely due to the lower and unknown conductivity of the ink compared to that of silver, as it was the idealized Ag used in simulation. The resonance shift can be attributed to ink spreading, which perturbs the geometry of the resonator. The meshed FSS of Fig. 4.3(b) featured a metallic area reduction to 53.3% of the solid FSS of Fig. 4.3(a), thereby increasing its optical transparency according to (4.1).

4.3.2 Effect of line-width

The extent of the ink minimization is determined by how thin the lines of the mesh can be fabricated. To study this behavior, several ink-minimized FSSs were fabricated and measured under normal incidence conditions. Fig. 4.4 shows the measured transmission results for line-widths varying from 60 – 100 μm , compared with that of the solid cell. It is observed that both the Q-factor and the resonant frequency are effected by varying the line widths in experiment. The smaller the line widths in fabrication, lower the Q-factor and lower the shift in resonance frequency. Depending on the specifications, line widths as low as 80 μm can be acceptable for FSS applications, where FSS with 60 μm tolerance is found to exhibit large resonance shift with poor rejection, i.e. large transmission. Overall, with reduced metallization areas, as low as 39.4% of unmeshed structures, the optical transparencies of the FSSs can be increased significantly, by factors of up to 1.86 for smaller line widths (e.g. 60 μm). Further, with these reduced metallization areas, the cost of screen-printing these surfaces is reduced according to reduction in expensive Ag ink. With this, ink minimization provides an attractive solution for improving the feasibility of 5G printed surfaces in large scale infrastructure applications (e.g. throughout buildings in cities).

4.4 Conformal FSS

Given the flexibility of the PET substrate, the fabricated FSSs can be used in conformal applications such as in radomes for out of band backscattering reduction [72], spectral filtering through conformal surfaces such as windows, radiation diversity of antennas [73], and wearable EMI shielding [74]. Given the periodicity of the meshing, the conformal FSS has its mesh slightly perturbed due to the stretching and spatial curvature. Conformal FSSs can also demonstrate improved bandwidth over equivalent planar FSSs [75].

To model such curvature effects, a numerical model is made in FEM-HFSS as shown in Fig. 4.5(a) along with measuring the FSS in experiments. In the simulation model, the FSS is wrapped around a cylinder with a certain radius of curvature ρ . The FSS spans an angular sector subtended by $\Delta\theta$. To emulate a plane wave throughout the simulation volume, perfect electrical conductors

Figure 4.4 Effect of the minimum line-width used in the proposed ink-minimized FSS cell on the measured transmission response compared to the solid FSS cell.

(PEC) were placed along the broad outside faces of the simulation volume and perfect magnetic conductors (PMC) were placed along the narrow outside faces. The structure was excited by a wave-port with wave polarized along x -axis and transmission was measured through a receiving wave-port across the FSS. In measurement, the conformal FSSs were measured by attaching the planar FSSs over a Styrofoam cylinder of measured diameter 228 mm (or a radius of curvature of 114 mm, available at the time of measurement). Given the low attenuation constant of the Styrofoam cylinder at mm-wave frequencies [76] and its overall thickness, loss due to the cylinder was considered negligible and not accounted for in the full-wave simulation.

Fig. 4.5(b) shows the measured and simulated transmission of both the solid and $80 \mu\text{m}$ line width ink-minimized FSSs, showing a good agreement between measurements and simulations. For both structures there is slight shift in resonance, however the Q factor is noticeably reduced. This reduction in Q factor can largely be attributed to the lower conductivity of the ink, compared to the nominal conductivity of silver used in simulation (6.1 MS/m). The small shift in resonance can be attributed to ink spreading in fabrication which perturbs the geometry of the FSSs, and therefore the resonant frequency response. Thus the proposed ink-minimized FSS shows acceptable performance under conformal conditions.

Fig. 4.5(c) further shows the measured results of each FSSs for varying radii of curvatures ρ . To determine the corresponding Q-factors, each measured curve was smoothed and the frequency span, Δf between a 10 dB increase from the resonance was measured. This data is shown in Tab. I for both solid and ink-minimized FSSs.

For large ρ , the resonance and Q-factor has little variation compared to the planar case ($\rho = \infty$). For extreme large curvature of $\rho = 10 \text{ mm}$, while the solid cell lost its resonant behavior, the ink-minimized cell maintained a well-defined resonance and the Q-factor in spite of gradual appearances of additional resonance dips. Based on this characterization, it can be concluded that the

Figure 4.5 Performance evaluation of the proposed ink-minimized FSS when mounted on a curved surface. (a) General schematic and the FEM-HFSS model used to simulate the transmission response of the conformal FSS consisting of 20 cells. (b) Measured and simulated results of the conformal ink-minimized and solid FSS with $\rho = 114$ mm. (c) Measured results of the fabricated FSSs with varying ρ .

Table 4.1 Measured effect of curvature

Cell	ρ (mm)	Δf (GHz)	Resonance (GHz)
Solid	10	—	—
	60	2.28	28.04
	114	3.35	28.09
	∞	2.45	28.32
Meshed	10	0.6	27.13
	60	4.38	26.82
	114	4.42	26.78
	∞	3.24	26.69

proposed ink-minimized FSS offers good performance up to $\rho = 60$ mm on a conformal surface, as well as showing a bandwidth improvement over the planar meshed structure. The variation in resonance location for the conformal FSSs compared to the planar FSSs can be attributed to the variation in element spacing due to the curvature, forming a perturbed resonant structure.

4.5 Conclusion

A thin, flexible and transparent PET based screen-printed band-stop FSSs with ink minimization has been experimentally demonstrated at 28 GHz using both planar and conformal conditions. The technique of meshing the proposed conventional FSS unit cell has been utilized to obtain an ink-minimized unit cell configuration while maintaining stable angular response of the surface within $\pm 60^\circ$. This not only led to reduced amount of expensive ink required in experiments, thereby lowering the manufacturing cost, but increased the optical transparency of the structure by a factor of 65% using an $80 \mu\text{m}$ line-width tolerance. While the meshed line widths of $80 \mu\text{m}$ and $100 \mu\text{m}$ in the fabricated FSSs show reasonable Q-factors, $60 \mu\text{m}$ line width exhibited considerable degradation in performance. Finally, the operation of the ink-minimized FSSs have been shown using a conformal geometry with a large radius of curvature (up to 60 mm) and a good performance has been confirmed. Overall, ink minimization of FSSs features increased feasibility through improved transparency and reduced cost, making it an attractive solution for large scale implementation. Future work includes the improvement in fabrication tolerances to produce better results with thinner line widths, the application of various other meshing schemes including non-uniform meshing, and the application of meshing to more general printed structures such as electromagnetic metasurfaces.

Chapter 5

PHASE-ONLY SYNTHESIS FOR METASURFACE BASED BEAM-FORMING

5.1 Introduction

In the previous chapter, FSSs were demonstrated as important tools for indoor 5G wireless systems, particularly for their application in EMI reduction through their spectral filtering operation. However, these surfaces are simple in terms of their cell arrangements as they are fully periodic. Each cell is exactly identical, providing a single resonant response for stop-band functionality. These surfaces can be generalized, however, such that the cells are not identical. With the additional degrees of freedom provided by the non uniform cell distribution, a much wider range of functionalities are provided. Unlike the FSSs which operate by filtering transmission, these more general surfaces can also operate via reflection. These generalized surfaces are referred to as electromagnetic metasurfaces.

Electromagnetic metasurfaces are 2D arrays of sub-wavelength resonating cells, whereby controlling the spatial distribution and properties of the individual particles, the amplitude, phase and polarization state of the scattered fields can be controlled in transmission or reflection. One emerging application of such metasurfaces is as smart reflectors to control millimeter-wave wave-propagation in wireless environments in the context of 5G communication [77], diffuse the beams or provide multi-sector coverage in a given wireless environment, for instance.

These reflectors are electrically large (typically hundreds of wavelengths), consisting of large number of sub-wavelength elements, which can be used to either diffuse the beams or provide multi-sector coverage in a given wireless environment, for instance.

The general design of metasurfaces for achieving arbitrary reflection beam profiles, require complete control over the amplitude and phase response of the cells. The amplitude variation can typically be achieved either using controlled resistive elements (thus adding ohmic losses) on

the unit cell or exploiting polarization rotation (generating cross-polarization) [78]. In certain applications or scenarios [79], such a strong cross-polarization is undesired and a natural question arises: Is it possible to generate an arbitrary shaped beam from the metasurface by synthesizing phase-only spatial modulation without polarization rotation? This question is addressed in this paper by taking inspiration from antenna array theory, where two specific methods are presented: Woodward-Lawson (WL) and Amplitude-to-Phase mapping technique, and applied to metasurface based smart reflectors.

5.2 Phase-Only Woodward-Lawson (WL) Synthesis

Consider the objective of synthesizing a metasurface to generate a desired far-field pattern of $G_c(\theta)$ as a response to a normally incident plane-wave. The continuous aperture excitation can be approximated as its Fourier transform, which when spatially discretized takes the form of complex source excitations $w(n)e^{j\psi(n)}$ of say, N elements. The corresponding array factor for this uniform array of N elements is given by

$$G_c(\theta) = \sum_{n=0}^{N-1} w(n)e^{j\psi(n)} e^{jkd n \cos(\theta)} \quad (5.1)$$

where k is the free space wavenumber, and d is the element spacing. Now the objective is to synthesize a reflective metasurface that can approximate $G_c(\theta)$ using phase only-modulation, i.e.

$$G_p(\theta) = \sum_{n=0}^{N-1} e^{j\phi(n)} e^{jkd n \cos(\theta)} \quad (5.2)$$

WL synthesis is a standard radiation pattern synthesis technique commonly used in antenna arrays based on sampling the desired pattern at discrete locations. The synthesized pattern is a sum of functions of the form $b_m \sin(N\phi_m)/N \sin(\phi_m)$ [19]. In the case of phase-only WL synthesis, excitation amplitudes are kept constant at unity, and only the phase weights are used to steer the sub-array factor composing functions to sample the desired pattern. The resulting phase profile of the entire array using WL method typically consists of a quasi-periodic series of linear profiles. Fig. 5.1(a) shows an example for a beam-diffuser. Using the phase-only WL method, the phase only weights are obtained and the resulting pattern is compared with the ideal complex array factor in Fig. 5.1(a). As can be seen, WL method achieves the main beam-width as desired.

To realize these phase-weights, a simple metasurface unit cell consisting of a single metallic square loop, backed by an infinite ground plane, is considered, as shown in the inset of Fig. 5.1(b). Since the goal is not to generate a cross polarized component, rotation of the cell is not permitted hence the synthesis is constrained to phase-only. Keeping the period $d \ll \lambda_0$ as a constant parameter, the size of the loop, a , can be used to vary the reflection phase, as shown in Fig. 5.1(b), with a phase range of about 318° . Next, a 1D metasurface model is created to realize the phase profile obtained

from the WL method earlier, and exited with a normally incident plane-wave. Due to finite dimensional tolerances of about $2.5\ \mu\text{m}$ in the inkjet printing on Polyethylene terephthalate (PET), the desired continuous phase profile is spatially discretized in HFSS. Fig. 5.1(a) shows the HFSS simulated response, which closely matches with the synthesized response with slight discrepancies in the side-lobe-levels (SLLs) due to mutual coupling.

Note that the phase slope can be reduced by increasing substrate thickness, allowing for better sampling due to tolerance. Different cell geometries such as the Jerusalem cross [80] can achieve full 360° phase range, which improves SLL. However, full 360° phase range is not completely necessary.

Because of the cell symmetry, a reflection based metasurface implementing this cell can operate with dual polarization, and for a given linearly polarized E-field, doesn't generate any cross-polarization.

Figure 5.1 Phase synthesis using Woodward-Lawson method to create a wide off-broadside beam. a) The synthesized field pattern. b) Phase response of a physical metasurface unit cell at 60 GHz with varying loop-lengths (computed in FEM-HFSS using Floquet boundary conditions).

5.3 Complex Excitation to Phase-Only Mapping

For a sector beam similar to that of Fig. 5.1, for instance, WL synthesis results in a wide Transition Factor (TF) which is the angular span between 90% level and 10% level of the local discontinuity [19]. A possible solution to provide more accurately shaped beams, is a direct mapping of complex-weights in the array factor to phase-only weights [81].

For large arrays ($> 10^5$ elements) synthesis methods are still required to combat the computational demand of optimization techniques [82] [83].

To demonstrate such a mapping, let us equate $G_c(\theta)$ and $G_p(\theta)$,

$$\sum_{n=0}^{N-1} w(n) e^{j\psi(n)} e^{jkdn \cos(\theta)} = \sum_{n=0}^{N-1} e^{j\phi(n)} e^{jkdn \cos(\theta)}. \quad (5.3)$$

Using Euler's formula, this simplifies to

$$\cos\{\phi(n)\} + j \sin\{\phi(n)\} = w(n) \cos\{\psi(n)\} + j w(n) \sin\{\psi(n)\}. \quad (5.4)$$

Equating the real and imaginary parts leads to

$$\cos\{\phi(n)\} = w(n) \cos\{\psi(n)\}, \quad (5.5)$$

and

$$\sin\{\phi(n)\} = w(n) \sin\{\psi(n)\}. \quad (5.6)$$

The solution to the phase mapping $\phi(n)$ must satisfy both equations above. A ratio of these equations to solve $\phi(n)$ leads to the trivial case where $\phi(n) = \psi(n)$ and $w(n) = 1, \forall n$. In this case, the mapping is poor and not a feasible near field to synthesis for $G_c(\theta)$.

An alternative approach to satisfying both equations approximately is to introduce a fast variation term into the phase in question, $\phi(n)$ through multiplication. This fast variation term is given by

$$s(n) = (-1, 1, -1, 1, -1, \dots, 1), \quad (5.7)$$

where the central two terms have the same sign to maintain pattern symmetry. This results in

$$\cos\{s(n)\phi(n)\} = w(n) \cos\{\psi(n)\} \quad (5.8)$$

and

$$s(n) \sin\{\phi(n)\} = w(n) \sin\{\psi(n)\}. \quad (5.9)$$

The odd symmetry of $\sin$ allows the left hand side of the equation above to vary rapidly by shifts of $\pm\pi$. In this case, both equations are still not closely satisfied. However, the amplitude $w(n)$ and the phase $\psi(n)$ can be decoupled and considered separately by preserving $\psi(n)$ exactly and solving the amplitude mapping problem directly as

$$\cos\{s(n)\phi(n)\} = w(n) \quad (5.10)$$

and

$$s(n) \sin\{\phi(n)\} = 0, \quad (5.11)$$

where $\psi(n) = 0$. Then, (5.3) holds approximately in the far-field because rapidly varying patterns in the near-field represent slowly varying patterns in the far-field due to the properties of Fourier

transform pairs. This means that the contribution due to (5.3) has little effect on the far-field pattern, and hence $w(n)$ and $\psi(n)$ have been decoupled. Then, to retain the phase information in the mapping, $\psi(n)$ can be preserved and used directly in the phase only pattern as

$$G_c(\theta) \approx \sum_{n=0}^{N-1} e^{js(n) \cos^{-1}\{w(n)\}} e^{j\psi(n)} e^{jkdn \cos \theta}. \quad (5.12)$$

Here, the phase-only near-field distribution is

$$\phi(n) = s(n) \cos^{-1}\{w(n)\} + \psi(n). \quad (5.13)$$

which is a function of both $\psi(n)$ and $w(n)$, retaining the general complex pattern information. To illustrate this mapping, Fig. 5.2(a) shows the resulting far-field pattern (s-Vec {1}), compared to the WL method. The desired main-beam is almost perfectly reconstructed with a very sharp transition factor with low side-lobes (< -20 dB). The drawback of this method, is the requirement of rapidly varying phase profile due to the s-Vector properties (Fig. 5.2(c)), which may be challenging to implement practically using a sub-wavelength unit cell of the metasurface. A possible remedy is to not alternate the sign of s_n every element, but every m elements for example. Fig. 5.2(a) shows one such case for $m = 2$ (s-Vec {2}). While the main beam remains unaffected, the SLLs increase significantly.

The increased SLLs observed for higher grouping of s-Vector elements with $m > 1$, can be attributed to the phase discretization effect [84]. To demonstrate that phase discretization is indeed the cause of high SLLs, a smaller cell of $d = 0.1\lambda$ (compared to $d = 0.2\lambda$) is used in Eq. 5.13. Fig. 5.2(b) shows the newly synthesized patterns, where the SLLs are significantly lowered in both cases of m , to the point where they can be lower than the ones obtained using WL. Fig. 5.2 further shows the pattern obtained using the unit cell phases of Fig. 5.1 considering $2.5 \mu\text{m}$ fabrication tolerances, to show the feasibility of the design. Practical considerations for the implementation of the mapping to reflective metasurfaces include the fabrication tolerance. For increasingly coarse variation in the cell dimensions, as well as random variations in the dimensions, the SLLs increase. Depending on the specifications, cell resolutions exceeding $20 \mu\text{m}$ can produce exceedingly high SLLs and ripple.

5.4 Conclusion

Two methods for synthesizing the phase profile of a reflective metasurface have been proposed based on WL method and amplitude-to-phase mapping techniques. While WL method provides low SLLs, it poorly approximates beams with sharp transition factors. In such cases, a direct complex amplitude-to-phase mapping, through the s-Vector, has been presented that closely approximates any arbitrary pattern characteristics using phase-only modulated metasurfaces. It has been shown that this mapping performs better than the WL method using smaller metasurface unit

Figure 5.2 Phase synthesis using amplitude-to-phase mapping technique. a) Comparison of realized patterns using s-Vectors with standard Woodward-Lawson method, for two cases of element groupings ($d = 0.2\lambda$, $N = 200$). b) Impact of the smaller unit cell size on the pattern characteristics ($d = 0.1\lambda$, $N = 400$.) c) Phase profile for WL synthesis. d) Phase profile for complex mapping with s-vector grouping of 1.

cells.

Chapter 6

SUMMARY AND FUTURE WORK

6.1 Key Conclusions

This thesis presents novel devices and techniques in this context for 5G applications. First, A 60 GHz linear antenna array for sectoral beam-forming and coverage enhancement for wireless applications and was designed, fabricated and measured. Second, an analog, low-profile and shielded spectrum analyzer was proposed for operation at mm-wave frequencies around the 60 GHz band based on a novel side-fire LWA configuration. The low-profile and compact size of the spectrum analyzer allowed a bench-top type system that can be placed on standard equipment with no electromagnetic interference due to its fully-shielded configuration. Thirdly, optically transparent and ink minimized FSSs screen-printed on PET at 28 GHz were proposed and experimentally demonstrated. The technique of meshing the proposed conventional FSS unit cell has been utilized to obtain an ink-minimized unit cell configuration while maintaining stable angular response of the surface within $\pm 60^\circ$. This not only led to reduced amount of expensive ink required in experiments, thereby lowering the manufacturing cost, but increased the optical transparency of the structure by a factor of 65% using an 80 μm line-width tolerance. Finally, a technique for the phase-only synthesis of multi-beam patterns in reflective metasurfaces based on a direct Fourier transform mapping of the near-field to far field was shown. Compared to Woodward-Lawson beam sampling, this synthesis features sharper transitions in shaped beams such as sector patterns, providing utility in potentially stringent indoor short-range radiation pattern specifications. These variety of proposed technologies provide an attractive solution to difficulties faced in common 5G applications such as spectral filtering, fundamental signal processing operations, and short-range indoor communications.

6.2 Future Work

Future research on the topics discussed in this thesis could potentially involve the following topics:

1. Chapter 2: The linear periodic LWA array has been demonstrated in 5G DF applications, as well as designed for circular polarization. The slot pair antennas can be further designed to feature arbitrary radiation patterns (amplitude and phase) for more complex sectoral operations. In case of dynamic coverage requirements, as suggested, various ports can be selectively excited using switches, which could be a part of future experimental demonstration. Furthermore, all ports can also be simultaneously excited to create a passive solution for a wide customizable coverage using conventional power divider, essentially converting the multi-port antenna configuration into a standard single port antenna structure.
2. Chapter 3: The integrated side-fire based LWA has been demonstrated using a proof-of-concept horn receivers. Improved mechanism for measuring the spectral S-parameters could be a part of the future work for simultaneous measurements of various port receivers and in real-time. Further, fully rigorous mathematical modeling of the radiation to facilitate characterization for future designs could be useful for fast efficient designs of the entire system including the placements of the receivers. In regards to the underlying side-fire antenna configuration, the electromagnetic description of stop-band suppression through cell asymmetry need to be understood in more details in periodic non-TEM structures, and some efforts need to be made to fully develop it.
3. Chapter 4: The ink-minimized FSS requires more improvement in accounting for fabrication tolerances to produce better results with thinner line widths. While basic square meshing has been used in the presented FSS structure, the application of various other meshing schemes including non-uniform meshing can be explored for comparable or superior electromagnetic performance while offering higher optical transparencies. In addition, the application of ink-minimization procedure using PET materials to more general printed structures such as electromagnetic metasurfaces can be interesting avenue to explore for general wave transformations, in addition to applying it at higher 5G bands such as 60 GHz.
4. Chapter 5: The chapter presented two analytical synthesis techniques based on Woodward Lawson and s-Vector based methods, for general beam-forming using phase-only metasurfaces, along with supporting array-factor and full-wave demonstrations. The synthesis methods require more development in the future, along with experimental verification of shaped multi-beam radiation patterns, while maintaining little to no cross-polarization response with symmetric dual-polarization cells. Possible synthesis using bianisotropic unit cells to provide additional degrees of freedom while preserving polarization could also be performed. Finally, generalization to 2D surfaces needs to be done, which is deemed straight forward but essential in practical applications.

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Appendix A Material Characterization

Figure A.1 Fabricated waveguide resonator for effective material characterization showing S_{21} . Arlon DiClad 880 ($\epsilon_r = 2.2$ and $\tan \delta = 0.0009$), and FEM-HFSS with fitted material ($\epsilon_r = 2.08$ and $\tan \delta = 0.003$). Resonator length and width are 50 mm and 2.5 mm, respectively.

Given the sensitivity of the devices to electrical material parameters such as ϵ_r and $\tan \delta$ (particularly the beam location of the spectrum analyzer considering the large focal length) a material characterization of the panel used to fabricate the devices was performed. This was done by fabricating a resonator and measuring its frequency dependent transmission S_{21} , as shown in Fig. A.1. The location and magnitudes of the peaks can then be fitted by running additional simulations of the resonator with new values of ϵ_r and $\tan \delta$. These values which match measurement most closely provide a more accurate characterization of the panel material beyond the standard values of Arlon DiClad 880 provided in HFSS. However, given that the frequency dependence of the material parameters varies from what is predicted in software, over a large bandwidth the curves cannot be perfectly fit over the band, hence some mismatch will occur. The curves were fit best to match over the band of interest (57-64 GHz) and the extracted effective material parameters were found to be $\epsilon_r = 2.08$ and $\tan \delta = 0.003$.