ANALYSIS AND DESIGN OF A MULTIFUNCTIONAL SPIRAL ANTENNA

A Dissertation

by

TENG-KAI CHEN

Submitted to the Office of Graduate Studies of Texas A&M University in partial fulfillment of the requirements for the degree of

DOCTOR OF PHILOSOPHY

August 2012

Major Subject: Electrical Engineering

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ABSTRACT

Analysis and Design of a Multifunctional Spiral Antenna. (August 2012)

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Chair of Advisory Committee: Dr. Gregory H. Huff

The Archimedean spiral antenna is well-known for its broadband characteristics with circular polarization and has been investigated for several decades. Since their development in the late 1950's, establishing an analytical expression for the characteristics of spiral antenna has remained somewhat elusive. This has been studied qualitatively and evaluated using numerical and experimental techniques with some success, but many of these methods are not convenient in the design process since they do not impart any physical insight into the effect each design parameter has on the overall operation of the spiral antenna. This work examines the operation of spiral antennas and obtains a closed-form analytical solution by conformal mapping and transmission line model with high precision in a wide frequency band.

Based on the analysis of spiral antenna, we propose two novel design processes for the stripline-fed Archimedean spiral antenna. This includes a stripline feed network integrated into one of the spiral arms and a broadband tapered impedance transformer that is conformal to the spiral topology for impedance matching the nominally-high input impedance of the spiral. A Dyson-style balun located at the center facilitates the transition between guided stripline and radiating spiral modes. Measured and simulated results for a probe-fed design operating from 2 GHz to over 20 GHz are in excellent agreements to illustrate the synthesis and performance of a demonstration antenna. The research in this work also provides the possibility to achieve conformal integration and planar structural multi-functionality for an Unmanned Air Vehicle (UAV) with band coverage across HF, UHF, and VHF. The proposed conformal mapping analysis can also be applied on periodic coplanar waveguides for integrated circuit applications.

DEDICATION

To my family and wife

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I would like to thank my advisor, Dr. Gregory Huff, for his guidance, patience, and overwhelming drive to explore new ideas. Under his supervision, I was able to dive into the topics that unfamiliar to me prior to doing research in the EML laboratory. Not only has he taught me a vast number of technical skills and abilities, he has also showed me how to present more clearly. I will always appreciate his support and guidance in all areas of my education.

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NOMENCLATURE

CPW	Coplanar Waveguide
PCPW	Periodic Coplanar Waveguide
CPS	Coplanar Strip
СМ	Conformal Mapping
PPC	Parallel Partial Capacitance
SPC	Series Partial Capacitance
FDTD	Finite Difference Time-Domain
FEM	Finite Element Methods
AR	Axial Ratio
VSWR	Voltage Standing Wave Ratio
UAV	Unmanned Aerial Vehicle

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CHAPTER I

INTRODUCTION

1.1 Preface

The highly demand for broadband wireless communications has been increasing because of a number of reasons including higher communication data rate by transmitting data on several separate spectrum and integration of several wireless services [1], [2]. As a result, a lot of researches and development have been focused on the study of modeling, design [3], optimization [4], and measurement methodology of broadband antennas [5]. TABLE I shows several examples of various wireless communication systems with their operating frequencies and bandwidths, illustrating that a very broadband antenna is necessary to integrate these system onto one platform. However, the wideband antennas (except spiral antennas) barely provide circular polarization [6], [7]. A circular polarization antenna is widely employed in radar, navigation, satellite and some mobile communications systems because of its insensitivity to orientations and polarization diversity between transmitters and receivers, and is generally required for various environments where omnidirectional receiving is desired [8]. Although several broadband circular polarized antennas have been proposed in recent years [9]-[11], their axial ratio (AR) bandwidth are hardly achieved up to 40%.

This dissertation follows the style of the *IEEE Transactions on Antennas and Propagation*.

System	Freq(MHz)	Polarization
Keyless entry	350	VP
GSM 900/1800	890-960/1710-1880	VP
PCS	901-941/1850-1990	VP
GPS	1563-1587/1164-1188	RHCP
Satellite Radio	2400-2460	RHCP
WiFi 11b/g/a	2.4GHz/5GHz	
UWB	3.1-10.6GHz	
3G wireless	1755-1850/2500-2690	
LTE (4G)	4 bands in 1900-2620	
WiMax	2.3 GHz\2.5 GHz \3.5 GHz	

TABLE IOVERVIEW OF RADIO SERVICES.

The spiral antennas, which easily cover the bandwidth over 20:1 with excellent circular polarization [12], have long been a popular choice for broadband radiating systems and continue to be relevant in modern communication systems and in other high-performance devices based on multifunctional or hybrid concepts. This includes the integration of various services and sensors into a single system as a means of adding new degrees of functionality, but this can also create challenges for the radiating structure (and potentially throughout the entire front-end of the wireless system). The spiral configurations can support these applications, but from a mathematical standpoint (design equations, etc.) only the equiangular spiral, the Archimedean spiral, and the rectangular counterpart of the Archimedean spiral have received wide attention. These spirals will be discussed and reviewed in the next section.

1.2 Literature Review

Apparently, the researches on examining the properties of Archimedean spiral antennas are more than equiangular spiral antennas. From a design point of view and their properties, both spiral antennas can be classified in the class of frequency independent antennas. The term "frequency independent antenna" is first coined by V. H. Rumsey [13] for a class of antennas whose properties is independent of frequency for all frequencies between the low and high frequency limits. Rumsey also stated that if the shape of the antenna could be specified entirely by angles, its performance would be independent of frequency, whose formula is recognized as an equiangular spiral. Therefore, the Archimedean spiral antennas are criticized by Mayes and Dyson that they are not belong to frequency independent antennas [14].

However, there is evidence that the Archimedean spiral is a better candidate than the equiangular spiral. A comparison between the Archimedean spiral and log spiral by fabrication and measurement undertaken in [15] concludes that the Archimedean spiral is a better choice based on its well-defined dispersion relation. The work in [16] also suggests an Archimedean spiral over an equiangular spiral because of constant arm width spiral easy to mount termination resistors for better performance. Despite these, the equiangular spiral and the Archimedean spiral share some similar properties and the research works on equiangular spiral are applicable to the Archimedean spiral as well. The following sub-section introduce the past research works on spiral antennas as well as the conformal mapping methods on solving coplanar waveguide problems.

1.2.1 Equiangular Spiral Antennas

After Rumsey proposed an idea about frequency independent antenna in 1957 and suggested a solution to the equiangular spiral antenna [17], Dyson created an infinite balun and conducted an experiment on slot-type spiral to demonstrate the idea of frequency-independency [18], [19]. This work also showed that the input impedance of the antenna is lower by increasing the width of the metal arm. In 1961, Cheo, Rumsey, and Welch published an analytical solution to frequency independent antenna, which showed the current distribution attenuated significantly as increase of spiral length and various radiation modes [20]. Sivan-Sussman demonstrated those various modes of the equiangular spiral antenna experimentally in 1963 [21].

Since the prior research works on equiangular spiral antennas in 1960s, no publication on equiangular spiral appeared until 1991 [22], which demonstrated experimentally that the spiral microstrip antennas can achieve very wide bandwidth of 2-18 GHz. The first numerical work on equiangular spiral antennas is proposed in 1996 by Wentworth and Rao using the method of moments [23]. The calculated impedance in this paper contradicted the self-complementary impedance of 188 Ω is not quite accurate since the mesh of spiral arm is coarse. In 2005, Lou and Jin presented the time-domain finite element method (TDFEM) for an equiangular spiral antenna, which did show an impedance of 188.5 Ω for a self-complementary spiral [24]. However, the subject of this paper is not on the spiral antenna itself but on the numerical verification. McGadden and Scott then presented the analysis of equiangular spiral antenna using finite-difference time-domain (FDTD) method, which is the first comprehensive numerical parametric

study on equiangular spiral antenna [25]. They also proposed a very easy formula for evaluating the self-complementary impedance on a half-space dielectric substrate by Booker's relations, which is the well-known Babinet's principle. Nevertheless, this formula does not take the geometric parameters of spiral and substrate height into consideration and is a highly approximated one.

In addition to these analysis works, other active researches are to obtain a unidirectional beam and to improve the antenna design and performance. A unidirectional beam can be achieved by placing a shallow cavity with absorbing strips [26] or placing an electromagnetic band gap (EBG) reflector [27] backed to the spiral. A simpler feeding structure by tapered microstrip balun [28] or by parallel-plane perpendicular-current feed [29] are also proposed to improve and to realize a completely planar spiral antenna compared to the conventional center-fed or vertically-connected coaxial line. Although these techniques are presented on equiangular spiral antennas, they can be used on Archimedean spiral antennas as well.

1.2.2 Archimedean Spiral Antennas

The Archimedean spiral antenna is known for its broadband characteristics with circular polarization, though it was not claimed as a type of frequency independent antenna. The fundamental design of Archimedean spiral was first reported in the 1950s by Turner [30]. The first attempt to solve the Archimedean spiral analytically was by Curtis in 1960 with approximation as a semicircle spiral [31]. The semicircle spiral model showed good agreement with experimental results.

In the same year of 1960, Kaiser proposed the well-known band theory of spiral antennas [32], which states that for the two-wire spiral transmission lines with negligible wire width, the radiation occurs in the regions where currents in the neighboring arms are in-phase. Although there is no rigorous mathematical description for the band theory, its easier-understood concept can explain several notable properties of spiral antennas. As is well accepted, the feeding structure near the spiral center determines the upper frequency limit of spiral antennas and the maximal circumference of the spiral determines the lower frequency limit. At radiation region where two currents are inphase, the radiation behavior of a spiral antenna is similar to one-wavelength loop antenna. The first mode radiation (also called normal mode in [12]) arises from a balanced excitation (also called odd excitation in [12]) and the second mode radiation (split-beam mode in [12]) arises when currents are in-phase in the feed point, which is called even excitation in [12]. The higher-order mode may occur if the spiral structure is large enough. However, only odd-mode radiation can be excited by a balanced excitation since no radiation occurs when the two currents are out-of-phase at even-wavelength circumference, unless there is any amplitude or phase error at the input terminal. As discussed in [33], the current distribution decay rapidly along the arm of the Archimedean spiral and thus the radiation from the higher-order mode is smaller than the first mode radiation.

After those research works in 1960s, no publication on Archimedean spiral antennas until 1986. Nakano *et al.* analyzed an Archimedean spiral antenna with the presence of an infinite PEC reflector by using the moment method with thin-wire

assumption [34]. Champagne *et al.* also applied the moment method with thin-wire assumption to analyze the impedance loaded multi-arm Archimedean spiral antenna, and first time revealed the split-beam mode on four-arm spiral by numerical calculation [35]. In 1997, Nakano *et al.* presented an analysis of the monofilar spiral antenna backed by a ground plane, which has a single arm configuration with a simpler feed system than two-arm Archimedean spiral [36]. It is worth to note that the excitation on this single arm Archimedean spiral antenna is called the unbalanced excitation, which is not contrary to the balanced excitation for two-arm Archimedean spiral antenna and should be clarified.

The analyses mentioned in the above paragraph are performed in free space. In 1998, Nakano *et al.* proposed a moment method analysis for a two-arm thin-wire antenna on a semi-infinite dielectric material [37] and for a single-arm printed wire on an infinite conductor-backed substrate [38]. In 2002, they analyzed a two-arm square Archimedean spiral antenna printed on two types of substrates, finite-size dielectric substrate with infinite and finite conductor-backed plane, respectively [39]. The finitedifference time-domain (FDTD) analysis in this paper showed the difference on radiation pattern, axial ratio (AR), and gain while less difference on input impedance between these two configurations. A notable result in [40] by examining a two-arm square Archimedean spiral antenna sandwiched by dielectric layers with infinite ground plane showed that the oscillation in the input impedance and AR is due to the reflection from the air/dielectric interface, which can be lessened by adding anti-reflection layer for impedance matching. In 2010, Nakano *et al.* compared the performance of single-arm and two-arm Archimedean spiral antennas in [41]. It should be addressed that the terminology in the work [42] confused a lot of researchers. The unbalanced mode excitation mentioned in this paper meant that one arm is fed by a coaxial cable and the other arm is a floating circuit.

The research conducted by Nakano *et al.* are mainly on investigation of the spiral antenna properties in different configurations using the moment method and FDTD. Other numerical works to evaluate the characteristics of spiral antenna can be found by finite-volume time-domain (FVTD) method [33], integral equation with thin-wire assumption [43], FDTD [44], finite element method (FEM) [45], and commercial full-wave solvers [46]. Over the past decade, the spiral was the subject of active researches and several modified designs on spiral configuration were proposed and investigated, such as stripline-fed spiral design [47], new combo-antenna design combining equiangular spiral, Archimedean spiral, and annular slot ring antenna [48], meander or zigzag spiral designs [49], coplanar strip spiral [50], three-arm (coplanar waveguide) spiral [51], and etc. [52]-[56].

In the above mentioned papers, the design and analysis of the spiral antenna were accomplished using the numerical techniques, which can provide accurate results, but are not convenient in the design stage and do not give any physical insight into the effect of each design parameter on the overall operation of the spiral antenna. On the other hand, the input impedance of spiral antenna is fully unpredictable until the optimized solutions are achieved through a trial-and-error iterative numerical process. The analytical input impedance is important for system design because of the need of impedance matching to system impedance, i.e. normally 50 Ohms. Recently, the lossy transmission-line model was proposed to characterize the input impedance of a spiral antenna [57]. However, they did not provide the model with the geometric parameters of spiral.

The balanced excitation required by spiral antennas often requires that wideband baluns and impedance transformers are utilized when excitation from a coaxial cable is desired. The unbalanced transition from the cable and the impedance mismatch between it and spiral antenna [58] are just two of the major design problems that can limit the desired wideband operation. As a result, numerous feed topologies and baluns have been investigated to provide the balanced feed for the spiral antenna (e.g. [51], [58]-[61]). In many cases the resulting feed structure is orientated perpendicular to the plane of the spiral. Structurally functionalized antennas and those for on-chip integration with other circuit components are just a few examples where a planar design is desirable, but a design that integrates the spiral, balun, and feed network into a multi-layer structure to enable other applications.

Overall, several issues introduced in this section and some notable results by past research works will be discussed more in this dissertation.

1.2.3 Conformal Mapping Method on Coplanar Waveguides

The coplanar waveguide (CPW) structure has been popular for Monolithic Microwave Integrated Circuit (MMIC) [62-65] and printed circuit board (PCB) [66] design because of its wide designable versatility. Since its introduction in 1969 [67], it

has been utilized in a lot of applications (not only in microwave fields [68]-[71]) and consequently has been investigated extensively on its characteristics.

The electrical properties of various CPW configurations have been well characterized by various techniques including conformal mapping [72]-[80], finite difference time domain (FDTD) method [81]-[84], spectral domain method [85], [86], integral method [87], and relaxation method [88]. Simple analysis such as conformal mapping assumes quasi-TEM propagation where the guiding wavelength is much larger than the structure dimensions. Other methods that are more rigorous use full-wave techniques that allow frequency dependence of phase velocity and characteristic impedance.

At the lower frequency that antennas and RF circuits usually operate, the conformal mapping analysis results in accurate calculations of the electrical parameters with several features including less computation time, ease of programming, and facilitating to get the effects of various physical dimensions on the electrical parameters. These analytical expressions by conformal mapping are accurate up to frequencies of 20-40 GHz compared to full-wave techniques [75], and ideal for computer aided design (CAD), especially for sensitivity analysis and variation prediction.

The mapping function for ideal CPW and coplanar strips (CPS) on semi-infinite substrate and infinite lateral ground plate was first presented by Wen [67], where the method was later extended for the design of a CPW directional coupler [89]. Davis *et al.* provided finite boundary correction for CPW with finite substrate thickness in 1973 [90]. Until 1980, Fouad Hanna provided the same correction for CPS [91] and in the same

year, he and Veyres proposed a well-known paper applying partial capacitance (PC) techniques to solve CPW with both finite metal dimensions and finite substrate [80]. Since then, a lot of research employed conformal mapping method and partial capacitance techniques to derive analytical expressions for the impedance and the permittivity of CPW with backed conductor plane [73], [92], CPW with upper and lower shielding [64], broadside-coupled CPW [93], parallel-coupled CPW [79], [94]-[96], parallel CPW [97]-[99], and asymmetrical CPW and CPS [100]-[103].

The partial capacitance techniques can be easily extended to multilayer cases as demonstrated in [65], [75], [78], [104] since it allows the total line capacitance to be approximated as the parallel combination of the capacitances of homogeneous substrate, which is cataloged as parallel partial capacitance (PPC) approximation. Since its first introduction in1980, the first theoretical justification was presented in 2003 [105] and showed that the PPC approach is valid when the dielectric constant is decreasing layer by layer away from the CPW structure. A modified PC or series partial capacitance (SPC) approach was thus suggested by several authors [106]-[108], but SPC approach is still valid when the dielectric constant is increasing layer by layer away from the CPW structure. A generalization of mixed PPC and SPC approach is not proposed yet. The quasi static solutions by conformal mapping can be easily modified to obtain dispersion effects [109] and conductor loss [110], [111].

1.3 Dissertation Organization

The organization of this dissertation has four main parts. The first part contains introduction including a review of the relevant works in Chapter I. Chapter II presents the modified band theory for an Archimedean spiral antenna with non-negligible metal width and derives a novel analytical solution on the characteristic impedance operating in the radiation region. The derivation is based on the assumption of periodic coplanar waveguide (PCPW) model and conformal mapping methods. Chapter III describes the immature but successful design process to design a stripline-fed two-arm Archimedean spiral antenna by proposing an inward-fed topology based on a center-fed spiral design with an impedance transformer that is conformal to the spiral. This work demonstrates the function of a stripline-fed Archimedean spiral antenna and supports the goal of developing planar broadband radiating structures. Chapter IV analyzes the radiation mechanism and the impedance properties of the stripline-fed Archimedean spiral antenna presented in Chapter III. Based on this analysis, a new design process containing intuitive physical concepts is proposed in Chapter V to design this multifunctional planar antenna. In this chapter, the modified design process employs the simple concept of impedance matching and analytical formulas to design the antenna as well as examination by full-wave simulation, while the full-wave simulation is totally applied in the design process I. This is also followed by measured and simulated results for a probe-fed fabricated structure designed to operate from 2 GHz to over 20 GHz on the same substrate to demonstrate the design process. The application of this multifunctional antenna on circular UAV design is introduced in Chapter VI. The last part of the

dissertation presents the analytic solutions of periodic coplanar waveguides for integrated circuits applications in Chapter VII. Finally, conclusions and future work are given in Chapter VIII.

CHAPTER II

ON THE CHARACTERISTIC IMPEDANCE OF TWO-ARM GAP-FED ARCHIMEDEAN SPIRAL ANTENNAS

The Archimedean spiral continues to be a widely studied antenna topology thanks to its broadband impedance and radiation characteristics. These have been investigated experimentally and computationally since their initial development in the late 1950's [31], [32], and a number of numerical methods have been developed and utilized in the decades following their introduction to model these broadband attributes. Examples of this include the method of moments based on a thin-wire assumption [37], [43], finite-volume time-domain (FVTD) [33], finite-difference time-domain (FDTD) [39], [44], finite element method (FEM) [45], [52], and similarly constructed commercial full-wave solvers [47], [49], [51], [112].

In many of these and other examples, the process of design and analysis relies on experimental observations and/or empirical methods to impart greater physical insight into this behavior. These methods have been collectively successful for synthesis, but a physically descriptive and rigorous analytical analysis of the Archimedean spiral antenna has received less attention. This is especially true concerning the input impedance. A semi-circular model [31] was first proposed in for this purpose, and the solution for an infinite number of equiangular spirals was obtained in [20]. The development of closedform analytical solutions for the input impedance, which both impart physical insight and remain amendable to the synthesis of two-arm spirals, has been somewhat limited. This may in part be attributed to the many variations the spiral can embody (cavity backed, etc.) as well as the ability to rapidly evaluate the structures using full-wave solvers, but it nonetheless remains of interest for both synthesis and analysis.

The radiation behavior is a more studied topic when compared to the impedance and it is commonly accepted that the basic operation of the balanced two-arm Archimedean spiral can be accurately explained using Band Theory [32]. This theory states that for the two-wire spiral transmission line with negligible wire-width, the radiation occurs in annular regions where currents in the neighboring arms are in-phase. The lossy transmission-line model in [57] applies this concept using the radiation resistance of loop antennas as a means to capture the impedance behavior. By extending this explanation to include microstrip [113], stripline [47], and other printed antenna topologies, where wire width can no longer be considered negligible, the current distribution will reside on the edge of the conductor and the power will radiated when the two neighboring current distributions are in-phase.

This deviation from the original constructs of Band Theory leads to the concept of common slot-line mode radiation for the spiral antenna. Hence, when radiation occurs from the common slot-line mode both the field distribution and its physical structure are similar to the propagating TEM coplanar waveguide mode. With multiple turns this leads to the analysis of periodic coplanar waveguide (PCPW) structures and the primary focus of this analysis, from which a closed-form analytical solution can be obtained using conformal mapping. This results in a quasi-static analysis which is rigorously valid only at zero frequency, but it can provide an accurate prediction across a wide frequency band [64], [65], where RF and microwave spiral antennas operate. This can also provide physical insight into the radiation mechanism of the two-arm Archimedean spiral antenna and provide an efficient analytical solution for both synthesis and analysis.

This chapter proposed a conformal mapping approach to derive quasi-static closed-form solutions for the characteristic impedances of PCPW; this is used to characterize the input impedance of the balanced, gap-fed two-arm Archimedean spiral antennas operating in their radiation region (where Band Theory predicts radiation will occur). For completeness, the radiating mechanism of spiral antenna is reviewed first along with the basic design parameters of the spiral geometry. This is followed by the development of a model for the PCPW (assuming conductors of negligible thickness). The mapping between physical and finite image domains is discussed next as a more straightforward approach towards deriving the input impedance of the spiral. A comparison is then made with full-wave electromagnetic solutions to validate the accuracy of this approach over a wide range of design parameters. A brief summary on the conformal mapping process and the resulting characterization of the spiral concludes the discussion.

2.1 Archimedean Spiral Antenna

The absence of a comprehensive design process for the synthesis of Archimedean spirals with desired input impedance can present many challenges when finding the optimum winding, arm width, or arm spacing. The analysis in [12] (albeit some time ago) states that the conductor width and conductor spacing are the least critical parameters,
and suggests that their ratio should be unity. Many subsequent designs continue to follow this suggestion for a self-complementary topology, and the resulting input impedance $Z_{in} = 60\pi \ \Omega$ (~ 188 Ω) for the infinitely-wound self-complementary spiral antenna of this topology in free-space is described best through Babinet's principle [114]-[118]. This can, of course, be applied to find impedances for a wide range of complementary structures (including spirals) using the widely recognized relation in (1), where Z_{metal} and Z_{slot} are input impedances of the metal and slot radiating structures, respectively, and η is the intrinsic impedance of the media where the structure is immersed.

$$Z_{\text{metal}} Z_{\text{slot}} = \frac{\eta^2}{4} \tag{1}$$

This result is fundamentally important but it cannot provide a priori information on the input impedance of non-self- complementary spiral antenna since Z_{metal} and Z_{slot} are typically unknown. The value of η for the spiral antenna when it resides on dielectric media with a finite-thickness is also an unknown parameter in the design process, although this can be solved by numerical methods. The remainder of this section represents a precursor towards directly solving for these parameters (explicitly Z_{metal} for the gap-fed two-arm structure) by examining the similarity of the spiral's cross section with propagating modes on a TEM transmission line.

2.1.1 Antenna Parameters

Fig. 1 shows a two-arm planar gap-fed Archimedean spiral antenna. The antenna is excited in the center of the spiral with a tapered Dyson-style balun [18]. This tapered topology reduces the capacitive loading at the central input end of the spiral antenna. The boundaries of the two metal arms are defined by four spiral curves, and the wellknown expressions for these are shown in (2), where *r* is the radius of the spiral curve, θ is the winding angle in radians, $a = RC/2\pi$ is the radius change rate of the spiral, *RC* is the radius change for one turn of a spiral arm, and r_{in} is the inner radius of the spiral. The same values of *a* and r_{in} are applied on these four curves, and the outer taper of the arm into a point is given by a circular curve defined using an offset angle θ_{off} with the outer radius in (3), where *N* is the number of turns of the spiral antenna.

$$r = a\theta + r_{in} \tag{2}$$

$$r_{out} = 2\pi N a + r_{in} \tag{3}$$

Using c_1 and c_2 to describe the two edge curves of one arm of the spiral requires $c_1 = a\theta + r_{in}$ and $c_2 = a (\theta + \theta_{off}) + r_{in}$. The width of the metal strip can then be defined as (4), the resulting slot width can be obtained by noting that $S = a\pi - W$, and the metallization ratio can then be defined as (5).

$$W = |c_1 - c_2| = \theta_{off} a \tag{4}$$

$$\chi = \frac{W}{W+S} = \frac{\theta_{off}}{\pi}$$
(5)

For convenience, the second metal strip can simply be generated by rotating the first metal strip 180° about the axis of the spiral. The feed region expanded in the bottom

of Fig. 1 has a gap length g with gap width W_g . The ratio of gap width to gap length has to be designed to occupy a very limited footprint in an effort to reduce the gap capacitance at input and to have a negligible effect on the antenna's impedance or pattern. This is generally limited by the desired frequency response of the design. This results in the structure shown in Fig. 1. In a general context, these parameters are well defined in other works related to the spiral, but they have been included here for continuity of nomenclature in later sections and to provide a quick reference on their physical meaning.



Fig. 1 Layout of the two-arm gap-fed Archimedean spiral antenna in free space with non-negligible metal width and tapered-down arm terminations.

2.1.2 Band Theory and Periodic Coplanar Waveguide (PCPW)

There is no rigorous mathematical description of the Band Theory for Archimedean spiral antennas. However, it is a powerful concept that provides an intuitive, versatile, and easily digested description of the radiating mechanism. According to this theory [32], the radiation from a two-wire spiral transmission line with negligible wire-width occurs when the two currents in the neighboring arms are in-phase. The active region where this radiation occurs is similar to a loop antenna whose circumference on the spiral plane corresponds to one wavelength at the operating frequency f in (6), where r_{rad} is the radius of the radiation region, λ_g is the guided wavelength propagating along the spiral arm, v_p is the guided phase velocity, c is the speed of light, and ε_{reff} is the effective dielectric constant that the propagating wave experiences (if a dielectric substrate is present).

$$2\pi r_{rad} = \lambda_g = \frac{v_p}{f} = \frac{c}{f\sqrt{\varepsilon_{reff}}}$$
(6)

A related but topologically different situation arises when the strip width of a metallic spiral arm is non-negligible (e.g., no longer a thin wire). It is well-known that the current distribution resides on the edges of the metal strips in this case and a slotline mode propagates between the arms, so a path difference between the outer curve and inner curve of one spiral arm will be created. Following the principles of Band Theory, it can be deduced that power will also be radiated when the two neighboring current distributions on the same metal strip are in-phase. Fig. 2 shows the cross-section of a nominal spiral (top) and field distribution on these arms when the two neighboring currents on one spiral arm are in-phase (bottom).

This is similar to a pair of in-phase slot-line modes, and the similarly-named common slot-line mode radiation. Using the symmetry of the electric field distribution, a set of perfect electric conductor (PEC) walls and perfect magnetic conductor (PMC) walls can be placed at the middle plane of the slot and the metal strip, respectively. The notional electric field distribution shown in Fig. 2 is explicitly valid *only* for an infinite number of arms with equal current and field distributions. A similar distribution can be assumed for a finite number of arms over the frequencies where common mode radiation occurs. This field distribution leads to the periodic coplanar waveguide (PCPW) mode, for which a closed-form analytical solution is examined in the following section using conformal mapping.



Fig. 2 Cross-sectional view of a two-arm Archimedean spiral antenna along *yz*-plane (top) and the electric field distribution (bottom) when radiation occurs, which is similar to the coplanar waveguide mode but with periodicity along the spiral cross section (named periodic coplanar waveguide (PCPW) mode).

2.2 Conformal Mapping

Conformal mapping is introduced in this section to obtain the characteristics of a PCPW transmission line. For quasi-TEM propagation, the characteristic impedance Z_c in (7) and effective dielectric constant ε_{reff} in (8) can be completely determined from the effective per unit length (*P.U.L.*) capacitances [119], where *L* and *C* are the *P.U.L.* inductance and capacitance, respectively, and C_0 is the capacitance with no dielectric material present. From (7) and (8), *C* and C_0 are the only unknown parameters required to obtain the quasi-static characteristics.

$$Z_c = \sqrt{\frac{L}{C}} = \frac{\sqrt{\varepsilon_{reff}}}{cC}$$
(7)

$$\mathcal{E}_{reff} = \frac{C}{C_0} \tag{8}$$

2.2.1 Schwartz-Christoffel Transformations

The Schwartz-Christoffel transformation provides a suitable conformal mapping for the calculation of these unknown capacitances. It maps the upper-half of the complex *t*-plane onto the interior of a polygon in the *w*-plane and the real axis onto the boundary of the polygon. The resulting expression in (9) for a half-plane [120] assumes a constant *B*, the mapping function w = f(t), and the interior angle at vertex ρ of the polygon as $\pi \alpha_{\rho}$. For a closed convex polygon, the restriction on α_{ρ} is given by (10). For more than four vertices (n > 4), this transformation can only be evaluated by numerical integration.

$$\frac{df(t)}{dt} = B \prod_{\rho=1}^{n} \left(t - t_{\rho} \right)^{-(1-\alpha_{\rho})}$$
(9)

$$\sum_{\rho=1}^{n} \left(1 - \alpha_{\rho} \right) = 2 \tag{10}$$

2.2.2 Elliptic Integrals

Fig. 3 shows the general transformation when mapping the upper half-plane onto the interior of the rectangle with $\alpha_k = 1/2$ and n = 4 or n = 3 to form a parallel-plate capacitor. The integrals in this transformation involve rational functions which can be reduced to the square root of third- and fourth-degree polynomials and their products. These can be expressed by a linear combination of elliptic integrals of the first, second, and third kinds. For n = 4, the conformal transformation shown in Fig. 3(a) uses the mapping function in (11), where the constant *A* describes a translation and constant *B* describes the rotation and magnification (neither are required here). For a generic value of *t*, the integral is said to be incompletely expressed by the elliptic integral of the first kind as (12), where φ and *k* are called amplitude and modulus of elliptic integrals, respectively.

$$w = A + B \int_{\rho=1}^{t} \prod_{\rho=1}^{4} \left(t' - t_{\rho} \right)^{-\frac{1}{2}} dt'$$
(11)

$$F(\varphi,k) = \int_0^{\sin\varphi} \frac{dx}{\sqrt{(1-x^2)(1-k^2x^2)}}$$
(12)

It is notable that when $\varphi = \pi/2$, the integration in (12) is from 0 to 1 and $F(\pi/2,k)$ is expressed as K(k); this expression is called a complete elliptic integral of first kind. The integral in (11) can be found in a table of integrals (e.g., [121]), and the boundary of the rectangle in the *w*-plane is given by (13) and (14) with the geometric parameters given by (15)-(17).

$$\int_{t_4}^{t_3} \frac{dt'}{\sqrt{(t'-t_1)(t'-t_2)(t'-t_3)(t'-t_4)}} = -jg_4 K(r)$$
(13)

$$\int_{t_3}^{t_2} \frac{dt'}{\sqrt{(t'-t_1)(t'-t_2)(t'-t_3)(t'-t_4)}} = -g_4 K(p)$$
(14)

$$g_4 = \frac{2}{\sqrt{(t_1 - t_3)(t_2 - t_4)}}$$
(15)

$$r = \sqrt{\frac{(t_1 - t_2)(t_3 - t_4)}{(t_1 - t_3)(t_2 - t_4)}}$$
(16)

$$p = \sqrt{\frac{(t_2 - t_3)(t_1 - t_4)}{(t_1 - t_3)(t_2 - t_4)}}$$
(17)



Fig. 3 Schwartz-Christoffel transformation of (a) n = 4 and (b) n = 3 for the general case.

For n = 3, the conformal transformation can also be mapped onto the rectangle in *w*-plane using (18). This transformation leads to (19) and (20) with geometric parameters (21)-(23), where *k*' is the complementary modulus of elliptic integral. It is noteworthy here to indicate that the parameters g_3 and g_4 are not in Fig. 3 since they represent the magnification of mapping. These will be cancelled in the calculation of capacitance.

$$w = \int^{t} \frac{dt'}{\sqrt{(t'-t_1)(t'-t_2)(t'-t_3)}}$$
(18)

$$\int_{t_3}^{t_2} \frac{dt'}{\sqrt{(t'-t_1)(t'-t_2)(t'-t_3)}} = -g_3 K(k)$$
(19)

$$\int_{t_2}^{t_1} \frac{dt'}{\sqrt{(t'-t_1)(t'-t_2)(t'-t_3)}} = jg_3 K(k')$$
(20)

$$g_3 = \frac{2}{\sqrt{(t_1 - t_3)}}$$
(21)

$$k = \sqrt{\frac{(t_2 - t_3)}{(t_1 - t_3)}}$$
(22)

$$k' = \sqrt{1 - k^2} \tag{23}$$

2.2.3 Elliptic Functions

Fig. 4 shows the reverse mapping from a rectangle onto the upper-half plane using the mapping function in (24), where sn(z,k) is the Jacobian elliptic function with a modulus *k* determined by (25)-(26) (from [122]), K = K(k), and K' = K(k').

$$t = \operatorname{sn}^{2}(z, k) \tag{24}$$

$$k^{2} = 16q \prod_{n=1}^{\infty} \left(\frac{1+q^{2n}}{1+q^{2n-1}} \right)^{8}$$
(25)

$$q = \exp\left(-\pi \frac{K'}{K}\right) \tag{26}$$



Fig. 4 Reverse mapping from a rectangle in *z*-plane onto the upper-half *t*-plane.

2.3 Analysis

2.3.1 PCPW in Free Space

The symmetric field distribution of the PCPW facilitates the placement of the PEC and PMC boundaries shown in Fig. 2. This can be represented by the equivalent circuit in Fig. 5(a). The total capacitance C_0 for one periodic section of PCPW can then be evaluated using this model. The *z*-plane for this topology shown in Fig. 5(b) has the coordinates $z_b = W/2 + S/2$, $z_c = W/2$ and $z_d = 0$, and (27) maps these points onto the *t*-plane.

$$t = -\cos\left(\frac{2\pi z}{W+S}\right) \tag{27}$$



Fig. 5 Calculation of total capacitance C_0 for PCPW in free space; (a) equivalent circuit and (b) conformal mapping steps from *z*-plane onto *t*-plane and then onto *w*-plane.

The parallel-plate structure in the *w*-plane can be obtained using the transformation in (18), and from this the *P.U.L.* capacitance C_a can be obtained by (28) and (29), where χ is the metallization ratio defined in (5) and ε_0 is the permittivity in free space.

$$C_a = \varepsilon_0 \frac{K(k_0)}{K(k_0')}$$
(28)

$$k_0 = \sin\left(\frac{\chi}{2}\pi\right) \tag{29}$$

From the equivalent circuit in Fig. 5(a), the total *P.U.L.* capacitance C_0 and the characteristic impedance Z_0 of the propagating wave when the two slot line modes are in-phase are obtained using (30) and (31), respectively.

$$C_0 = 2C_a \tag{30}$$

$$Z_0 = \frac{1}{C_0 c} \tag{31}$$

2.3.2 PCPW in Substrate

Conformal mapping techniques can also be used to evaluate the PCPW embedded in dielectric material. Fig. 6(a) shows the cross sectional view of spiral antenna embedded in substrate with thickness of 2h and dielectric constant of ε_r . Fig. 6(b) shows the field distribution of the common slot-line mode where radiation occurs. The total *P.U.L.* capacitance *C* of substrate-embedded PCPW can be evaluated by the partial capacitance approximation [104] and modeling the air-dielectric interface as a PMC. The accuracy of this approximation [105] is limited to ε_r being greater than that of background medium.

Fig. 6(c) illustrates the conformal mapping steps for the calculation of C_1 ; the coordinates of the original PCPW structure on the complex *z*-plane are $z_a = z_b + jh$, $z_b = (W + S)/2$, $z_c = W/2$, $z_d = 0$, and $z_e = jh$. The rectangle on the *z*-plane is first mapped onto the *t*-plane with the mapping function in (32), where *k* is evaluated by (25) using (33) and K = K(k).

$$t = \operatorname{sn}^2\left(K\frac{2z}{W+S}, k\right) \tag{32}$$

$$q = \exp\left(-\pi \frac{2h}{W+S}\right) \tag{33}$$



Fig. 6 (a) Cross sectional view of a two-arm Archimedean spiral antenna embedded in substrate, (b) electric field distribution in a section of PCPW when radiation occurs and partial capacitance approximation, and (c) conformal mapping steps for evaluation of C1, from z-plane onto t-plane onto w-plane.

The resulting *t*-plane mapping points in (34) are mapped to the *w*-plane using (35)

$$t_{a} = \frac{1}{k^{2}}$$

$$t_{b} = 1$$

$$t_{c} = \operatorname{sn}^{2} (\chi K, k)$$

$$t_{d} = 0$$
(34)

$$w = \int \frac{dt}{\sqrt{t(t-1)(t-t_c)(t-t_a)}}$$
(35)

The *P.U.L.* capacitance C_1 can then be obtained using (36) and the modulus in (37).

$$C_1 = 2\varepsilon_0 \left(\varepsilon_r - 1\right) \frac{K(k_1')}{K(k_1)}$$
(36)

$$k_{1} = \sqrt{\frac{1 - \mathrm{sn}^{2}(\chi K, k)}{1 - k^{2} \mathrm{sn}^{2}(\chi K, k)}}$$
(37)

The effective dielectric constant ε_{reff} in (38) and the characteristic impedance Z_0 in (39) for the common slot line modes in PCPW can then be derived using (7) and (8).

$$\varepsilon_{reff} = 1 + \left(\varepsilon_r - 1\right) \frac{K(k_1')}{K(k_1)} \frac{K(k_0')}{K(k_0)}$$
(38)

$$Z_0 = \frac{60\pi}{\sqrt{\varepsilon_{reff}}} \frac{K(k_0')}{K(k_0)}$$
(39)

2.4 Results and Discussions

Analytically obtained results from the conformal mapping of the common slotline mode in PCPW are compared in this section to results from full-wave simulations [123]. The basic PCPW topology in free-space is examined first. Results for the substrate-embedded PCPW follow this using a range of commercially available substrate heights and dielectric constants. The technique is then applied to analyze the input impedance of the gap-fed two-arm Archimedean spiral antenna in both free-space and when embedded into a range of substrates with different heights and dielectric constants.

2.4.1 PCPW in Free-Space

Fig. 7 shows the characteristic impedance of PCPW operating in the common slot-line mode in free space as a function of the metallization ratio χ . A very wide range of *W* and *S* are accounted for in this plot ($0.05 \le \chi \le 0.95$), and the results shown are analyzed at 10 GHz in full-wave simulations. The results obtained by conformal mapping remains in very good agreement with simulations when no higher order slot-line modes exist.

2.4.2 Substrate-Embedded PCPW

Fig. 8 and Fig. 9 show the characteristic impedance of PCPW embedded in substrate of $\varepsilon_r = 2.2$ and $\varepsilon_r = 10.2$, respectively, with h = 0.254 mm, 0.508 mm, 1.27 mm, and 2.54 mm. The periodic cell used to analyze the PCPW has been fixed as W + S = 5mm in the analysis of these examples to make the range of metallization ratio χ meaningful and the analysis was again for 10 GHz. These results are intuitive since this field distribution now has dependence on both χ and substrate height h. Specifically, larger values of (W+S)/h correspond to larger conductor spacing S compared to h with fixed metallization ratio χ and an increased field distribution outside the substrate; this reduces the total *P.U.L.* capacitance and increases the characteristic impedance of the PCPW.



Fig. 7 Real-valued characteristic impedance as a function of metallization ratio obtained using conformal mapping and numerical simulation as PCPW operating in the common slot-line mode in free space.

The results for the higher dielectric constant of $\varepsilon_r = 10.2$ are generally in better agreement than for the lower dielectric constant of $\varepsilon_r = 2.2$. This can be explained by the higher field concentration in the higher dielectric constant, and a more accurate partial capacitance approximation to model the air-dielectric interface as PMC.



Fig. 8 Analytical and simulated real-valued characteristic impedance of the substrate-embedded PCPW in the common slot-line mode with dielectric constant of $\varepsilon_r = 2.2$, where CM is abbreviated conformal mapping.



Fig. 9 Analytical and simulated real-valued characteristic impedance of the substrate- embedded PCPW in the common slot-line mode with dielectric constant of $\varepsilon_r = 10.2$.

While both have excellent agreement for higher substrate heights, discrepancies are observed for thin substrates with smaller metallization ratio. When this is the case, the field distribution is less concentrated in the substrate as shown in Fig. 10(b) compared to Fig. 10 (a) and the partial capacitance approximation losses accuracy due to the modeling of the dielectric-air interface as a PMC.



Fig. 10 The main field distribution of PCPW in (a) thicker substrate with higher metallization ratio and (b) thinner substrate with lower metallization ratio.



Fig. 11 Visualization of the simulated Archimedean spiral antenna of metallization ratio from 0.0833 (left) and 0.9166 (right).

2.4.3 Two-Arm Gap-Fed Archimedean Spiral Antenna in Free-Space

The PCPW model is now applied to evaluate the input impedance of the gap-fed two-arm Archimedean spiral antenna in Fig. 1 (in free-space). Fig. 11 shows the spiral design to be simulated from the metallization ratio $\chi = 0.0833$ to $\chi = 0.9166$ with RC =10 mm, N = 5, an inner radius given by (40), $g = 0.1r_{in}$, $W_g = 0.01W$ (to reduce gap capacitance), and the outer radius obtained by (3). Notice that the conformal mapping analysis is only valid when operating at a frequency within its balanced-mode radiation.

$$r_{in} = \frac{S}{2\sin\left(\frac{S}{2a}\right)} \tag{40}$$

Fig. 12 shows the simulated real part of input impedance as a function of frequency. Frequency-independent behavior of the input impedance can be observed in Fig. 12 from 2 GHz to 8 GHz for $\chi = 0.5$. This verifies the well-known property of self- complementary structures [114]. For $\chi \neq 0.5$ (non-self- complementary), the input impedance in the radiation region is no longer a frequency-independent value. This is explained by the difference between frequency-independent antennas and Archimedean spiral antennas [14].

As expected, only the self-complementary structure exhibits frequency independent behavior. A less-pronounced dependence can also be observed in Fig. 12 at higher metallization ratios due to the more concentrated field distribution in the narrower slot. This makes the *P.U.L.* capacitance of transmission line less dependent on frequency.



Fig. 12 Real-valued input impedance of two-arm Archimedean spiral antenna in free space.

The low frequency operating point of the spiral antenna has to be determined in order to compare the input impedance of spiral antenna with the quasi-static PCPW model. According to the band theory, the lowest radiating frequency f_L in (41) is determined by the outer radius r_{out} and ε_{reff} .

$$f_L = \frac{c}{2\pi r_{out}\sqrt{\varepsilon_{reff}}}$$
(41)

In practice, the lower frequency limit can be two to three times this theoretical limit [50] given by (41). Using this observation, the outer radius $r_{out} \sim r_{in} + RC \times N$ of the aforementioned spiral example corresponds to approximately $3f_L = 3$ GHz (in free space). Fig. 13 shows that the input impedance at 3 GHz for the spiral antenna. Excellent agreement is achieved between simulated results of the gap-fed spiral and those obtained by conformal mapping and the PCPW model. Above the lower frequency limit, the input

impedance is frequency-dependent. Fig. 13 also shows the comparison between the input impedance at 8 GHz of spiral antenna in free space and those obtained by conformal mapping. This indicates that the PCPW model can predict the input impedance of the two-arm gap-fed Archimedean spiral antenna over a very wide frequency band, especially for a larger metallization ($\chi > 0.5$).

It is important to note that the real-valued input impedance found using the analytical PCPW model based on conformal mapping and the full-wave simulation of the physical spiral both accurately predict the self-complimentary value of $\eta_0/2$ at $\chi = 0.5$ (shown in Fig. 13(b)). It is also noteworthy that this analysis yields a different result from related work [124] on the analytical approximation of input impedance for non-self-complementary spiral antennas based on quasi-TEM coplanar strip theory. The most likely difference between the results from this work and those from resides in the interpretation of the band theory for wire spirals and the operation of spirals with non-negligible arm widths.

Fig. 14 shows the impact of convergence on the accuracy of the numerical simulations used to validate the PCPW model. The several values of the "delta S" parameter, which captures the difference in $|S_{11}|$ in results between successive simulations after adaptive meshing refinement, were examined along with different meshing and simulation parameters in an attempt to provide the most accurate results. The values of 0.01, 0.001, and 0.0001 are shown in Fig. 14. Values above 0.0001 cannot achieve self-complementary impedance ($\eta_0/2$) at $\chi = 0.5$, illustrating that for a highly curvilinear spiral a significant computational effort is required to obtain the results

derived from conformal mapping. Adjusting other simulation parameters yielded similar results.



Fig. 13 (a) Analytical and simulated input impedance of the two-arm Archimedean spiral antenna in free space at 3 GHz and 8 GHz, and (b) the zoomed region of red dash square in (a), where the black dash lines show that the self-complementary impedance of 188.5Ω at metallization of 0.5 for both spiral antenna and PCPW.



Fig. 14 Simulated input impedance by HFSS with different convergence setting of "delta S" = 0.0001, 0.001, and 0.01.

2.4.4 Substrate-Embedded Two-Arm Gap-Fed Archimedean Spiral Antenna

Fig. 15 shows the conformal mapping analysis of the PCPW model and simulations of a substrate-embedded gap-fed two-arm Archimedean spiral antenna. The antenna shown in Fig. 11 is now embedded in a substrate of $\varepsilon_r = 2.2$ with heights h = 0.254 mm, 0.508 mm, 1.27 mm, and 2.54 mm. The input impedance from simulations has been recorded at 3 GHz since it shows less frequency-dependence for the spiral antenna embedded in this substrate.

Note that the HFSS simulation in this case is performed with the convergence setting of delta S = 0.001 instead of 0.0001 due to heavily computational effort, resulting in somewhat inaccuracy. The discrepancies are observed for thin substrates. The behavior of the input impedance curves of spiral antenna shown in Fig. 15 (dash line with triangular markers) are similar to those characteristic impedances of PCPW

obtained by HFSS (dash line with triangular markers shown in Fig. 8). As mentioned before, the partial capacitance approximation cannot model the air-dielectric interface very well for these cases.



Fig. 15 Real-valued input impedance of a two-arm Archimedean spiral antenna embedded in a substrate of $\varepsilon_r = 2.2$ and varied height *h* simulated by HFSS (dash line with triangular markers) at 3 GHz, compared with the characteristic impedances of PCPW model obtained by conformal mapping (solid line).

Fig. 16 shows the simulated real-valued component of the input impedance for the same antenna once it has been embedded within a substrate of $\varepsilon_r = 10.2$ and h = 1.27mm. The HFSS simulation in this case is performed with the convergence setting of delta S = 0.01. As expected, it demonstrates a rapidly resonating input impedance, especially for the higher metallization ratio. This follows observations by Nakano et al. [40] that the oscillation is due to the reflection from the air-dielectric interface. According to [40], this can be reduced by adding another layer to achieve impedance matching for radiated wave, but at the expense of reduced bandwidth. In lieu of this, the comparison cannot be made at a single frequency point. Therefore, this input impedance is averaged from the lower limit at 2 GHz up to 8 GHz to provide a comparison with the quasi-static conformal mapping results.



Fig. 16 Real-valued input impedance of a two-arm Archimedean spiral antenna embedded in substrate of $\varepsilon_r = 10.2$ and height h = 1.27 mm with different metallization ratio.

Fig. 17 shows that the conformal mapping results of PCPW model are very close to the average input impedance of Archimedean spiral antenna embedded in a high dielectric constant substrate of $\varepsilon_r = 10.2$. Notice that the discrepancies are observed but the behavior of impedance curves is still quite fitting, though the average values also take the frequency- dependent behavior into account.



Fig. 17 Average input impedance from 2 GHz to 8 GHz of a two-arm Archimedean spiral antenna embedded in substrate of $\varepsilon_r = 10.2$ and varied height *h* simulated by HFSS (dash line with triangular markers) compared with the characteristic impedances of PCPW model obtained by conformal mapping (solid line).

CHAPTER III

DESIGN PROCESS I:

STRIPLINE-FED ARCHIMEDEAN SPIRAL ANTENNA

Center-fed spiral antennas have been studied extensively (e.g., [18], [32], and [125]) to support the needs of complex communication and sensing system as both arrays and stand-alone radiators. The balanced excitation of the center-fed spiral from a coaxial cable often requires wideband baluns or impedance transformers. These are just two of the major design challenges that can impact the desired operation, and numerous techniques have been investigated for this purpose (e.g. [51], [58]-[61]). Many of the resulting feed structures are electromagnetically functional but perpendicular to the plane of the spiral. These may not be desirable for structurally functionalized antennas and for the on-chip integration of the antenna with other components.

The proposed inward-fed two-arm Archimedean spiral antenna retains the desired planar form factor of the gap-fed spiral using a multi-layer topology. This is accomplished by structurally integrating the balun and feed network into one arm of the antenna using a stripline arrangement. This concept is based on [61], but the proposed design now operates over a considerably larger bandwidth, resides on electrically thinner substrate, and has a tapered impedance transformer which is now conformal to the outer-most winding of the spiral.

An inward-fed two-arm Archimedean spiral antenna is proposed which integrates a stripline feed network into one of the spiral arms. A broadband stripline tapered impedance transformer conformal to the spiral's winding provides a novel matching network between the input impedance of the spiral and the characteristic impedance of the strip-line. The Dyson-style balun at the spiral's center converts the guided stripline mode to the radiating spiral mode. The transformation from a gap-fed design to a stripline-fed design is provided to illustrate the design process. Measured and simulated results for a probe-fed design operating from 2 GHz to over 20 GHz are provided to illustrate the synthesis and performance of a demonstration antenna.

3.1 Synthesis of the Antenna Topology

Three designs of a two-arm Archimedean spiral antenna are presented to illustrate the transformation from a gap-fed design into a stripline-fed design. A stripline-fed design without an impedance transformer is provided as an intermediate step to better illustrate the synthesis process. The antenna is designed to have nominally high input resistance by initially choosing a slot-to-metal ratio less than unity; this value is desired to be greater than the measurement system's impedance Z_0 .

3.1.1 Gap-Fed Spiral Antenna

Fig. 18 shows the two-arm, two-turn, gap-fed Archimedean spiral antenna with arm width w and inter-arm separation s. The spiral is embedded in a dielectric slab of ε_r and thickness h_{sub} to represent the height of a stripline stack using two identical substrates of height 0.5 h_{sub} . The antenna resides at $z = 0.5 h_{sub}$ to account for dielectric loading in the stripline configuration. The triangular geometry in the upper-right of Fig.

1 has a base r_c that tapers linearly into the gap width g. The base-side connects to the spiral through a quarter-circle section of radius r_c . The inner and outer radii r_{in} and r_{out} are used to determine the lower-bound on operating frequency.



Fig. 18 Geometry of the two-arm gap-fed spiral antenna (left) at $z = 0.5 h_{sub}$, inset of gap feed region with circular and triangular transitions (upper-right), and cross-section of the substrate-embedded antenna (lower-right).

Fig. 19 summarizes the simulated [123] results for a design with $r_c = w = 5.0$ mm, g = 0.2 mm, and s = 1.0 mm using an RT Duroid® 5880 substrate with $\varepsilon_r = 2.2$ and h_{sub} = 1.016 mm (two 0.508 mm thick substrates). Quasi-frequency independent behavior (similar to an ideal spiral) is observed from 2 GHz to over 20 GHz. The non-zero reactance across this bandwidth is attributed in part to the circular and triangular sections at the center and embedding the gap-fed structure in a dielectric slab.



Fig. 19 Simulated input impedance Z_{in} and VSWR of the gap-fed spiral in Fig. 18.

The average input resistance across the stable impedance bandwidth is 83.24 Ω . This value is used to determine the closely-matched stripline characteristic impedance Z_I = 90 Ω and results in a VSWR < 1.5 across the aforementioned bandwidth. This is not a surprising result given the freedom to choose the reference impedance but the striplinebased antenna is designed around Z_I and its choice is very relevant to the synthesis of the antenna. The radiation patterns were examined and remained typical for spirals with similar architectures.

3.1.2 Stripline-Fed Spiral Antenna I (No Taper)

Fig. 20 illustrates the cross-section of the stripline-fed spiral antenna. This is synthesized by transforming one arm of the gap-fed spiral into a finite-ground stripline feed structure. The top and bottom grounds are formed by moving *Arm 1* from its substrate-embedded position to z = 0 mm (the bottom of the stack) and creating a copy at

 $z = h_{sub}$ (the top of the stack). These act as top and bottom grounds in the finite-ground stripline feed structure (guiding mode) and collectively act as *Arm 1* of the spiral antenna (radiating mode).



Fig. 20 Stripline-fed spiral antenna: (top) nominal cross-section of the stripline-fed topology, (bottom-left) stripline grounds formed by *Arm 1* at z = 0 and h_{sub} ; (bottom-center) stripline signal line (*Arm 2*) at z = 0.5 h_{sub} with no taper and $w_n = w_1$ and (bottom-right) stripline signal line (*Arm 2*) at z = 0.5 h_{sub} with curvilinear taper between w_0 and w_1 .

A signal line of width $w_1 = 0.34$ mm at z = 0.5 h_{sub} completes the transformation of *Arm 1* to a stripline feed network. It is both centered and conformal to the spiral winding of the top and bottom grounds, and extends through the feed gap until it connects with *Arm 2*. Fig. 20 shows the new *Arm 2*. The width w_1 corresponds to the characteristic impedance Z_1 and the width w_n refers to the outer-edge of *Arm 1*, which is unchanged in this configuration. The triangular sections at the center of *Arms 1* and 2 perform the mode-converter between the guided (stripline) and radiating (spiral) modes similar to the operation in [18]. Fig. 21 shows the overlay of all three metal layers stripline antenna with Arm 1 hatched semi-transparent (in grey) to show the section of Arm 2 that resides between the two Arm 1 layers.



Fig. 21 Stripline-fed spiral antenna without a tapered impedance transformer (left) and with a single-turn tapered stripline impedance transformer (right).

Fig. 22 shows the simulated results for the stripline-fed spiral antenna with no taper in the signal line. It is excited at the outer edge of *Arm 1*. The VSWR uses reference impedance Z_1 and remains below 1.5 across the band of interest. Minor scalloping of the radiation pattern occurs (not shown) as a result of the transformation to a stripline spiral, but the structure behaves very similar to the gap-fed design. This suggests that the stripline and spiral modes are sufficiently isolated.

3.1.3 Stripline-Fed Spiral Antenna II (Taper)

Fig. 23 summarizes the simulated results for the tapered *Arm 2* topology in Fig. 20 and Fig. 21. The tapered stripline impedance transformer matches Z_1 to $Z_0 = 50 \Omega$ (of the measurement system). The taper transitions inward from $w_n = w_0 = 0.76$ mm at the perimeter of *Arm 1* along one full winding of the spiral into w_1 . It should be noted that the width w_0 of the tapered line shown in Fig. 20 and Fig. 21 has been arbitrarily increased to better illustrate the profile of the taper. This value is limited in practice to provide efficient operation of the stripline mode.



Fig. 22 Simulated input impedance Z_{in} and VSWR of the stripline-fed spiral antenna using the signal layer in Fig. 20 without an impedance transformer.

The stripline taper length is determined according to band theory [32], such that the single outer winding can provide approximately one wavelength of transition. The circumference using the outer radius $r_{out} \sim r_{in} + 4s + 3w$ (with $r_{in} \sim r_c + 0.5g$) corresponds to a free-space wavelength at 1.98 GHz, which results in a length that facilitates lowreflections [126] across the bandwidth of the spiral (above 2 GHz and above). Minor scalloping of the radiation pattern (not shown) also occurs in this configuration, but the VSWR (Fig. 23) remains below 1.5 and has behavior similar to the initial stripline-fed topology.



Fig. 23 Simulated input impedance Z_{in} and VSWR of the stripline-fed spiral antenna using the signal layer in Fig. 20 with an impedance transformer.

3.1.4 Summary of Gap-Fed and Stripline-Fed Spiral Antennas

The impedances in Fig. 19, Fig. 22, and Fig. 23 for the gap-fed, stripline-fed, and stripline-fed with taper, respectively, all demonstrate similar behavior across the frequency independent range. The VSWRs have been overlaid in Fig. 24. This verifies to a degree that the antenna's impedance is not significantly changed with the transformation from a gap-fed spiral into a stripline-fed spiral (both with and without impedance tapers).



Fig. 24 Comparison of simulated VSWR of the gap-fed and stripline spirals.

3.2 Fabricated Antenna and Experiments

The stripline-fed spiral with the tapered design of *Arm 2* was selected for fabrication. The model in Fig. 25 shows a stepped stripline-to-microstrip transition (from w_0 to $w_{ms} = 1.58$ mm) and SMA probe-launch that were added to enable accurate and repeatable measurements. It includes a finite-ground stripline of width w_0 and length $L_A = 22.5$ mm and two metal vias of radius 0.615 mm at the step. These connect the upper portion of *Arm 1* to the linearly tapered section of the microstrip ground plane at z = 0 mm (with d = 7.5 mm and $L_B = 8$ mm) to the larger ground of length $L_C = 7$ mm. This transition was not optimized.

Fig. 26 shows the fabricated antenna and transition in Fig. 25 using seven #10-32 Nylon bolts and nuts mechanically fasten the two substrate layers. Initial discrepancies between measured and simulated results were observed (VSWR > 2). This was attributed to a thin air gap near the center (where bolts were undesirable) so a thin interfacial layer (< 0.1 mm) of Polydimethylsiloxane (PDMS) [127] ($\varepsilon_r = 2.8$ and tan $\delta = 0.015$) was applied and compressed during curing to fix this mismatch.



Fig. 25 The microstrip feed of the antenna with built-in impedance transformer.



Fig. 26 Fabricated stripline-fed antenna (with taper) and microstrip transition.

Fig. 27 shows the VSWR of the measured and simulated designs (with and without the probe feed). These results are in good agreement and suggest that the impedance is not significantly impacted by the transformation into a stripline structure. Fig. 28 shows the measured and simulated radiation patterns of the fabricated antenna in Fig. 26 at 2.5 GHz, 5 GHz, 10 GHz and 20 GHz for the *xz* and *yz* elevation cut-planes.
The mean μ and standard deviation σ of the broadside axial ratio (AR) in dB from 2 GHz to 20 GHz are (μ , σ) = (2.88, 1.35) and (5.12, 2.21) for the gap-fed and striplinefed spiral, respectively. The AR improves to (μ , σ) = (2.17, 0.73) and (4.84, 2.19) from 5 GHz to 20 GHz for both the gap-fed and stripline spiral, respectively. Measurements of the AR across the band were not available. The radiation pattern at 10 GHz and 20 GHz illustrate the impact of the asymmetry created by the microstrip transition and the excitation of unbalanced spiral modes that deteriorate the performance of the balanced spiral mode. Misalignment of the two substrates during fabrication also contributes to this error, but the overall agreement between simulated and measured data is acceptable. The major deviations between these results are most pronounced in the radiation patterns at frequencies above 10 GHz and the AR across the band.



Fig. 27 VSWR of the measured and simulated stripline-fed spiral antenna with impedance taper after substrate-bonding with PDMS.



Fig. 28 Measured and simulated elevation (xz and yz) radiation patterns at (a) 2.5 GHz, (b) 5 GHz, (c) 10GHz, and (d) 20GHz for the stripline-fed spiral antenna with impedance taper.

CHAPTER IV

ANALYSIS OF

A STRIPLINE ARCHIMEDEAN SPIRAL ANTENNA

There have been many investigations of the radiation characteristics of spiral antennas. Over the years, a lot of researches have been focused on examining the properties of Archimedean spiral antennas in various configurations, such as multi-arm spiral [35], [128], three-arm spiral [51], single-arm spiral [36], [38], [41], two-arm eccentric spiral [129], spirals above a plane reflector in free space [34], [130], spiral on semi-infinite electric materials [37], spirals backed by a cavity [33], [131], spirals on a dielectric substrate backed by a conducting plane reflector [39], and spirals backed by an electromagnetic band-gap (EBG) reflector [132]. In the above mentioned works, the antenna designs were proposed and their radiation properties were studied. However, there are no rules on how to design these spiral antennas accurately and therefore one has to resort on numerical method to predict antenna properties and then capable to optimize antenna performance before fabrications.

The development of analytical solutions to have a comprehensive understanding on spiral antennas [20], [31] has been limited mostly because of their complex curving structures. Therefore, a number of numerical methods and CAD tools have been developed to model very complex shapes of spiral antennas with supporting materials. However, typical frequency-domain techniques such as the finite element method (FEM) [45], [131] and the method of moments (MoM) [37], [43], and time-domain techniques such as finite difference time domain (FDTD) [39], [44] are still very time consuming for broad-band applications to be employed as an iterative and optimizing design tool. Since both of spiral antenna and its feeding element affect the impedance matching, it is also difficult to have both of them designed and analyzed simultaneously. To reduce development costs and design round-times on numerical or experimental means, the design rules and design process is necessary.

This chapter is based upon the design of stripline-fed Archimedean spiral antenna (or simply stripline Archimedean spiral antenna) presented in previous chapter. The stripline Archimedean spiral antenna has one arm configuration in stripline structure and a simpler feed system than the conventional center-fed or gap-fed spiral antenna as discussed. The stripline Archimedean spiral antenna does not require a balun circuit while it uses the layout similar to Dyson balun design [18] to convert the propagating stripline mode into radiating spiral mode. This spiral design without a balun circuit and with an integrated impedance- matching is beneficial, in particular, for fabrication of planar antenna connected with systems and an antenna array composed of spiral. The design process in previous chapter presented a concept that the stripline Archimedean spiral antenna has similar input impedance as its original form of planar gap-fed Archimedean spiral antenna with chosen geometry parameters and dielectric substrate, but how to design those geometry parameters of a stripline Archimedean spiral antenna on different substrate is absent. Following this analysis, the main purpose of this dissertation is to develop an easier design method for stripline-fed Archimedean spiral antenna at the design stage.

4.1 Summary of Stripline-Fed Archimedean Spiral Antenna

Fig. 29(a) shows the three-layer design and coordinate of a stripline Archimedean spiral antenna with separated height h between each layer. The expressions describing the Archimedean curves forming the spiral arms are

$$r = a\theta + r_{in} \tag{42}$$

where *r* is the radius of the spiral curve, θ is the winding angle in radian, *a* is the spiral growth rate, and r_{in} is the inner radius of the spiral. The two arms are truncated at outer radius

$$r_{out} = 2\pi N a + r_{in} \tag{43}$$

where *N* is the number of turns of the spiral antenna. The antenna is supported by Rogers RT/duroid 5880 substrate ($\varepsilon_r = 2.2$) with thickness of 2h = 1.016 mm (40 mil) as shown in Fig. 29(b). Notice that the input point has been moved to the outer radius by feeding with stripline transmission line.

The key component of this design is the Dyson balun structure in the central region of spiral, where is the dash square part in Fig. 29(b) zoomed in Fig. 30. The Dyson balun is formed by rapidly tapering the finite stripline grounds of dimension W_g and extruding the inner conductor of stripline width W_{ic} through the gap connecting the spiral arm. It consists of a triangular geometry with a base-dimension r_c and gap width g that tapers into the feed gap width g_{feed} and a quarter-circle section of radius r_c that connects the triangular gap-section to the spiral arms. The basic spiral dimension is

defined by the arm width W and the arm spacing S. The spiral growth rate is then determined by $a = 2(W+S)/2\pi$.

The antenna configuration parameters in [47] is summarized here: $r_c = W = W_g = 5.0 \text{ mm}$, S = 1.0 mm, $W_{ic} = 0.34 \text{ mm}$, $g_{feed} = 0.2 \text{ mm}$, and g = 1.2 mm. These remain the same in the following analysis. The simulation in previous work is performed in an ideal case with negligible metal thickness.

4.2 Analysis 1- Impedance Properties

The commercial numerical software HFSS based on finite element method (FEM) [123] is adopted to analyze a stripline Archimedean spiral antenna. The performance of stripline Archimedean spiral antenna is affected by the arm width, growth rate, dielectric loading, and feeding structure. Due to the very complex antenna geometry of spiral with so many parameters, the use of full wave solver to optimize antenna design becomes burdensome. Therefore, a design process not taking all of the parameters into consideration at the same time is necessary to reduce design time and to achieve the desired specifications. In this section, the stripline Archimedean spiral antenna is disintegrated to examine its operation and then synthesized back to small spiral step by step.



Fig. 29 Layouts of a stripline-fed Archimedean spiral antenna, (a) separated three layer designs and (b) implementation with substrate stack-up of height 2h.



Fig. 30 Geometry of the central region of a stripline Archimedean spiral antenna.

For any antenna design, the impedance matching is a key factor to radiate power effectively. To better understand the impedance properties, the stripline implementation of Dyson balun design shown in Fig. 30 is unwrapped into straight transmission line first while remain the dimension of triangular gap-section in the spiral center. The thick black dash line shown in Fig. 31 represents the unwrapped spiral arm, which is then split into both sides with the middle perpendicular plane modeled as a perfect magnetic conductor (PMC) wall. The three-port network is obtained with port10f stripline structure as well as port2 and port3 of elevated slot line structure, of which cross sectional view and denotation of parameters are also shown in Fig. 31. The parameters are the same as those denoted in Fig. 30, except the metal thickness t also included considering its effect on characteristic impedance in the next section. Note that the inner conductors of stripline between the upper ground and bottom ground at port2 and port3 are removed since the

field distribution of stripline mode is assumed to be confined between two ground plates and has no effect on elevated slot line modes.



Fig. 31 A three-port network remaining the dimension of triangular gap-section of Dyson balun design, where port1 shows a stripline structure and port2 shows an elevated slot line structure.

In order to examine the operation of Dyson balun of our original design, the triangular section is kept the same as that of stripline Archimedean spiral antenna but the arm spacing *S* of three-port network is designed to have the dimension of gap width *g*, i.e. S = g = 1.2 mm. The simulated characteristic impedance of each port is shown in Fig. 32 indicating that the stripline structure has a characteristic impedance of 83 Ω and the elevated slot line structure has a characteristic impedance of 157 Ω . If port2 and port3

are re-combined into a straight transmission line as the thick black dash line shown in Fig. 31, the characteristic impedance will be paralleled and the port of parallel elevated slot line structure is going to have a characteristic impedance of 78.5 Ω (black line in Fig. 32), which is quite close to that of the stripline mode. In other words, the reason why the power of stripline mode can be transmitted through the Dyson balun is because of the impedance-matching between two transmission lines of stripline and parallel elevated slot line.

Fig. 33 shows the simulated S-parameters to verify the power transmission due to impedance matching. The S_{11} shows low reflection at port1 of stripline mode and the S_{21} shows the power transmission from port1 to port2 about 3dB implying that the power is split equally from stripline mode into two elevated slot line modes. Specifically, the operation of Dyson balun is basically a power splitter and a mode converter, converting the stripline mode into two elevated slot line modes equally.



Fig. 32 Real part of characteristic impedance of stripline mode (red), elevated slot line mode (blue), and parallel elevated slot line mode (black) of the three port network.



Fig. 33 Simulated S_{11} (green), S_{22} (blue), and S_{21} (red) of the three-port network, where 3dB line means the power is split equally from port1 to port2 and port3, and 6dB line means the reflection from port2 due to the impedance mismatch between port2 and parallel impedance of port1 and port3.

4.3 Analysis 2- Common Mode Radiation

The unwrapped spiral arms shown in Fig. 34 are designed in order to verify the power transmission through the entire Dyson balun design. The triangular gap-section and quarter-circle section of Dyson balun is remained the same as those of stripline Archimedean spiral antenna (shown in Fig. 30). The unwrapped spiral arms consist of a parallel elevated slot line port and a stripline port with the same geometry parameters described in Section 4.1, while the arm spacing *S* is still given as S = g = 1.2 mm and the metal thickness is t = 0.

Fig. 35 shows the simulated S-parameters of unwrapped spiral arms. The S_{11} still shows no reflection at stripline port due to the impedance-matching between parallel elevated slot line port and stripline port. However, the S_{21} shows the power cannot be transmitted from port1 to port2 at 15.5 GHz. The power loss at 15.5 GHz can be explained by the electric field distribution shown in Fig. 36, where the two elevated slot line modes are in-phase at 15.5 GHz while they are not in-phase at other frequency points. This phase difference is due to the path difference between the outer elevated slot and inner elevated slot around the Dyson balun. When the two elevated slot line modes go in-phase, then the radiation so-called common mode radiation occurs inducing the power loss into free space. From the cross section view of port2, a clear radiation field is observed (not shown) supporting the common mode radiation.



Fig. 34 The unwrapped spiral arms with port1 of stirpline mode and port2 of parallel elevated slot line mode; the two dashed square parts show the cross sectional view of parallel elevated slot line structure with its dimensional parameters and the remaining Dyson balun design as in the central region of stripline Archimedean spiral antenna, respectively.



Fig. 35 Simulated S_{11} (green) and S_{21} (red) of the unwrapped spiral arms, where the lowest S_{21} is found at 15.5 GHz.



Fig. 36 Magnitude of electric field distribution at (a) 10 GHz, (b) 15.5 GHz, and (c) 20 GHz.

4.4 Analysis 3- Spiral Radiating Mode

To investigate the common mode radiation on stripline Archimedean spiral antenna, the unwrapped spiral arms in Fig. 34 is further designed to wind the two straight transmission lines in a small spiral turns of N = 0.5. The structure is shown in Fig. 37. It is easy to observe that there exists path difference between the inner and outer

elevated slot lines, which causes phase difference between these two propagating modes although they are excited simultaneously at the gap-section of Dyson balun.



Fig. 37 The straight transmission lines of parallel elevated slotline mode and stripline mode are wound in a small spiral shape.



Fig. 38 Simulated S_{11} (green) and S_{21} (red) of structure shown in Fig. 9, where the power loss arises at frequency of 5 GHz and 14.5 GHz.



Fig. 39 Simulated magnitude of electric field distribution at (a) 5 GHz and (b) 14.5 GHz.

The simulated S-parameters are shown in Fig. 38, where the S_{21} shows the power cannot be transmitted from stripline port to parallel elevated slot line port at frequency of 5 GHz and 14.5 GHz, respectively, because the two slot line modes are in-phase at these two frequency points (as shown in Fig. 39) and thus power is radiated. Furthermore, one can image that when the spiral turn *N* is increased, the two elevated slot line modes having path difference may go in-phase somewhere on spiral arm then producing the common mode radiation. Therefore, it is easy to conclude that the so-called spiral radiating mode is actually the radiation from in-phase elevated slot line mode (stripline Archimedean spiral antenna) or can be extended to a common slot line mode for general gap-fed planar two-arm Archimedean spiral antenna.

Here should be noted that the presented antenna in [47] had an arm spacing of S = 1 mm, which is different to that used in this analysis, i.e. S = 1.2 mm. The effect of smaller arm spacing produces the higher per-unit-length capacitance of transmission line and decrease the characteristic impedance of parallel slot line mode. As mentioned in Fig. 32, it can be imaged that the characteristic impedance of parallel port with S = 1 mm will be lower and cause the impedance more mismatching. This is because of the work in [47] started from a chosen geometry parameters and proved the functionality of stripline Archimedean spiral antenna but did not optimize.

CHAPTER V

DESIGN PROCESS II OF

A STRIPLINE ARCHIMEDEAN SPIRAL ANTENNA

This chapter is based on the analysis on impedance properties and radiation mechanism of a stripline Archimedean spiral antenna presented in previous chapter. The stripline implementation of Dyson balun acts as a mode converter and power splitter. Power can be transmitted to spiral radiating modes because of impedance-matching between two different transmission line structures, i.e. parallel elevated slot line and stripline. The radiation of spiral antenna occurs when the two elevated slot line modes are in-phase; therefore, the spiral radiating mode can be decomposed and explained by a common elevated slot line mode (for stripline Archimedean spiral antenna) or a common slot line mode (for gap-fed planar Archimedean spiral antenna). Based on this understanding, a design process is proposed simply by the concept of impedance matching with a design example. A compact (45 mm \times 40 mm \times 1.016 mm) bidirectional stripline Archimedean spiral antenna operating from 2 GHz to over 20 GHz is presented using this design process with a full-wave numerical simulation tool to optimize the geometry of spiral center. The measurement of the final design shows a good agreement with the simulated results.

5.1 Design Procedure 1 – Stripline Design

As was pointed out previously, the power able to transmit from the stripline guiding mode to spiral radiating mode is consequent to impedance matching between stripline and parallel elevated slot line, employing stripline based Dyson balun as a mode converter and a power splitter. The radiation of stripline Archimedean spiral antenna occurs when the two elevated slot line modes are propagating in-phase. Following the above analysis, an easier design method for stripline-fed Archimedean spiral antenna is developed in this section. The antenna design can be achieved simply by impedance matching of stripline mode and parallel elevated slot line mode. For simplicity, the design procedure will be described by using an example on a chosen substrate of Rogers RT/duroid 5880 of thickness h = 0.508 mm (20 mil) and a desired characteristic impedance of 75 Ω . This, however, does not limit our method to the chosen dielectric and substrate thickness; it can be extended to design any characteristic impedance of spiral antenna on various substrate stack-up.

For convenience in parametric design, the stripline structure is designed first. The metal thickness of t = 0.035 mm is also taken into consideration since it is going to decrease the characteristic impedance. A design equation with non-negligible metal thickness and infinite ground plates can be found in [133]. With the desired impedance $Z_0 = 75 \ \Omega$ and the chosen substrate stack-up, the conductor width of stripline is determined as $W_{ic} = 0.36$ mm.



Fig. 40 Simulated characteristic impedances of stripline with finite ground plates effect.

However, the stripline with infinite ground plates cannot be applied in stripline Archimedean spiral antenna design and there is no design equation including the finite ground effect on stripline. The numerical solver is used to determine the finite width W_g of ground plates. The simulated frequency response of characteristic impedance is shown in Fig. 40. As expected, the frequency-dependent impedance is observed because of the field distribution of stripline no longer confined within the substrate due to the finite ground; hence the field residing outside the substrate reduces the per-unit-length capacitance and then increases the characteristic impedance of stripline. In order to confine the field distribution of stripline mode inside the two ground plates without any effect on the spiral radiating mode as well as to keep the antenna design smaller, the finite ground width is determined at its saturated value $W_g = 4$ mm.

5.2 Design Procedure 2 – Parallel Elevated Slot Line Design

In the second step, the parallel elevated slot line structure is designed to possess the same characteristic impedance as stripline to transmit all the power from stripline mode into two elevated slot line modes. The upper-left of Fig. 41 shows the top view of stripline Archimedean spiral antenna design with spiral turns N = 4, where the cross section view of spiral arm is cut along the dash-dot line and displayed in the upper-right of Fig. 41. From the cross section view, it looks like a periodic structure and a section of periodic region is extracted in the lower-right of Fig. 41. When the spiral antenna is operating in its radiation region, the two elevated slot line modes are in-phase; hence assumed the symmetric field distribution of these two propagating modes, the perfect magnetic conductor (PMC) walls can be placed at the middle plane of stripline ground plate. Notice that the inner conductor of stripline is removed due to the stripline mode designed to well-confined in the ground plates with little coupling to the elevated slot line modes.

In this design example, the width of spiral arm *W* is chosen to be equal to W_g for simplicity and thus the only parameter left is the arm spacing *S*. Since there is no design equation for this transmission line structure, the numerical software is applied to obtain its characteristic impedance. Fig. 42 shows the simulated characteristic impedance of parallel elevated slot line structure modeling with PMC on the boundary. The frequency-dependent behaviors are observed at higher arm spacing. Eventually, the arm spacing is determined as *S* = 0.96 mm.



Fig. 41 Configuration of parallel elevated slot line: (upper-left) top view of stripline Archimedean spiral antenna design with N = 4, (upper-right) cross section view of spiral arms, which is similar to a periodic structure, and (lower-right) the periodic region of parallel elevated slot line structure.



Fig. 42 Simulated characteristic impedance of parallel elevated slot line.



Fig. 43 Electric field distribution in a section of parallel elevated slot line structure when common mode radiation occurs and its partial capacitance approximation.

The characteristic impedance of parallel elevated slot line mode can also be approximated and evaluated by conformal mapping analysis. For a quasi-TEM propagation, the characteristic impedance Z_0 can be completely determined by effective per unit length (*P.U.L.*) capacitances [134] as

$$Z_0 = \frac{\sqrt{\mathcal{E}_{reff}}}{cC} \tag{44}$$

where *c* is the speed of light, *C* is the *P.U.L.* capacitances that propagating mode sees, $\varepsilon_{reff} = C/C_0$ is effective dielectric constant, and C_0 is the *P.U.L.* capacitance contributed by the field distribution in free space. Fig. 43 shows the conjectured electric field distribution of common elevated slot line mode. Due to the symmetry of field distribution, a set of PMC can be placed at the middle plane of the metal strip and facilitate the analysis. The *P.U.L.* capacitances of this propagating mode can be approximated by partial capacitance technique as $C = C_0 + C_1$, where C_1 is the capacitance contributed by dielectric substrate with modeling the air-dielectric interface as a PMC.



Fig. 44 (a) A single cell of the periodic structure with no dielectric material present and (b) approximated PCPW structure to evaluate C_0 .



Fig. 45 (a) A single cell of the periodic structure with dielectric material, (b) approximated PCPW structure to evaluate C_I , and (c) equivalent mapping cell.

Fig. 44(a) shows a single cell of the periodic structure taken for conformalmapping analysis. This elevated slot line structure can be approximated by a slot line structure (Fig. 44(b)) with a diagonal distance d as slot width and a modified metal width W_m given by

$$d = \sqrt{S^2 + h^2} \tag{45}$$

$$W_m = \frac{W \times d}{S} \tag{46}$$

The capacitance C_a is essentially equal to quarter the capacitance of a periodic coplanar waveguide (PCPW) structure reported in [134]. It turns out that

$$C_a = \frac{\varepsilon_0}{2} \frac{K(k_0)}{K(k_0')} \tag{47}$$

$$k_0 = \sin\left(\frac{\pi}{2}\frac{W_m}{W_m + d}\right) \tag{48}$$

where $K(k_0)$ is the complete elliptic integral of first kind and $k_0' = (1 - k_0^2)^{1/2}$ is the complementary modulus of elliptic integral. The total *P.U.L.* capacitance C_0 is given by

$$C_0 = 4C_a \tag{49}$$

Fig. 45(a) shows one section of periodic structure with the dielectric material and the conjectured electric field contributing to the *P.U.L.* capacitance C_1 . It can be approximated by a slot line structure shown in Fig. 45(b). Notice that the field distribution above (below) the diagonal distance *d* in Fig. 45(a) is approximated by the field distribution in the upper (lower) half plane of Fig. 45(b), respectively. Assumed the field distribution in Fig. 45(b) is symmetric, a perfect electric conductor (PEC) wall can be placed in the middle of diagonal distance *d* as shown in Fig. 45(c). According to the equivalent series and parallel circuits, the *P.U.L.* capacitance C_d of Fig. 45(b) is equal to that of Fig. 45(c). Following the derivation in [134], we have

$$C_{d} = \varepsilon_{0} \left(\varepsilon_{r} - 1 \right) \frac{K(k_{1}')}{K(k_{1})}$$
(50)

$$k_{1} = \sqrt{\frac{1 - \mathrm{sn}^{2} \left[\frac{W_{m}}{W_{m} + d} K(k), k\right]}{1 - k^{2} \mathrm{sn}^{2} \left[\frac{W_{m}}{W_{m} + d} K(k), k\right]}}$$
(51)

where sn(x,k) is the Jacobian elliptic function with variable *x* and modulus *k* determined by (52)-(53).

$$k^{2} = 16q \prod_{n=1}^{\infty} \left(\frac{1+q^{2n}}{1+q^{2n-1}} \right)^{8}$$
(52)

$$q = \exp\left(-\pi \frac{2h}{W_m + d}\right) \tag{53}$$

The total *P.U.L.* capacitance C_1 is then given by

$$C_1 = 4C_d \tag{54}$$

The effective dielectric constant $\varepsilon_{reff} = C/C_0$ and the characteristic impedance Z_0 in (44) for the common slot line modes in parallel elevated slot line structure can then be derived using (55).

$$C = C_0 + C_1 \tag{55}$$

Fig. 46 shows the characteristic impedance of parallel elevated slot line approximated by PCPW model and obtained by conformal mapping has a very good agreement with numerical simulation.

It is noteworthy here that the arm spacing S = 0.2 mm can be chosen to have the impedance match with the system impedance of 50 Ω with no additional impedance matching circuit design. However, this arm spacing is too narrow to be fabricated by milling machine or chemical etching in our lab. In addition, for our chosen substrate

stack-up, the inner conductor width of stripline will be changed to $W_{ic} = 0.76$ mm to have $Z_0 = 50 \ \Omega$ and thus the wider ground plate is needed for stripline to confined the field distribution as well as the larger antenna design due to the wider spiral arm width. This demonstrates the iterative design procedure is needed to design stripline Archimedean spiral antenna. While compared to traditional design method considering all the parameters of spiral at the same time, this design procedure has great advantages in the design phase. Furthermore, to achieve a 50 Ω spiral antenna without needs of matching circuit, a substrate of higher dielectric constant can be used and will be demonstrated in Section 5.5.



Fig. 46 Characteristic impedance of approximated PCPW structure calculated by conformal mapping (red) compared with the real-valued characteristic impedances of parallel elevated slot line structure obtained by HFSS (blue). Both structures are embedded in a substrate of $\varepsilon_r = 2.2$ and h = 0.508mm with metal width W = 4mm.

5.3 Design Procedure 3 – Dyson Balun Design

After design the spiral parameters, the stripline based Dyson balun is implemented to connect these two transmission lines together. In order to examine its functionality, the two transmission line designs are wound into spiral configuration. Referred to Fig. 30, the radius r_c of quarter circle region is chosen to be equal to W and thus the last two parameters to be designed are the gap width g and the feed gap width g_{feed} . Since there is no design formula for Dyson balun, the parametric studies are performed on a stripline Archimedean spiral antenna with a small spiral turns N = 2, as shown in the sub-figure of Fig. 47. The simulated VSWR is shown in Fig. 47. In order to determine which set of parameters is the best design, the average VSWR from 2.5GHz to 20 GHz is calculated. It is found that the lowest average VSWR is about 1.24 with the design of g = 1.2 mm and $g_{feed} = 0.2$ mm.

It is noted that the spiral curve is not consistent with the quarter circular curve, as shown in Fig. 30, inducing somewhat impedance mismatch in the spiral center. Thus, the design concept for Dyson balun is to keep the arm spacing at this region not vary far away to S.

In the final, the parameters for the final prototype stripline Archimedean spiral antenna design are summarized in the following: $r_c = W = W_g = 4.0$ mm, S = 0.96 mm, $W_{ic} = 0.36$ mm, $g_{feed} = 0.2$ mm, and g = 1.2 mm.



Fig. 47 VSWR of entire stripline Archimedean spiral antenna with a small spiral turns to find out the optimized Dyson balun design.

5.4 Fabrication and Measurements

5.4.1 Integrated Impedance Transformer Design

To verify the proposed stripline-fed Archimedean spiral antenna through experiment, a tapered stripline of length about one full winding of spiral arm is employed to transform the antenna impedance (75 Ω) to a standard impedance (50 Ω). Fig. 48 shows the integrated configuration of the prototype antenna designed and a stripline impedance transformer. The tapered stripline is centered and conformal to the spiral winding of the ground layer and tapers from $W_{ic} = 0.36$ mm to $W_z = 0.76$ mm.



Fig. 48 Stripline Archimedean spiral antenna with a single-turn tapered stripline impedance transformer.

Fig. 49 shows the simulated VSWR and axial ratio (AR) of the stripline Archimedean spiral antenna integrated with impedance transformer. The VSWR is below 1.5 over the frequency bands from 2.5 GHz to over 20 GHz, which proves that the impedance transformer inside the antenna structure does not affect the antenna performance.

The axial ratio shown in Fig. 49 is calculated at the zenith (*z*-axis) of antenna design. A good circular polarization antenna should have AR below 3dB, while an AR less than 10 dB and greater than 3dB is defined as elliptical polarization [29]. The simulated AR shows that the stripline Archimedean spiral antenna has AR below 5 dB over the frequency range 6 to 20 GHz with several frequency bands having AR below 3dB, the overall operation of which is close to circular polarization. The axial ratio can be easily improved by increasing the number of spiral turns N at the expense of wider bandwidth and larger antenna dimensions.



Fig. 49 Simulated VSWR and axial ratio (AR) for stripline Archimedean spiral antenna with a single-turn tapered stripline impedance transformer.

5.4.2 Transition Design

Several extra dimensions were added to enable accurate and repeatable impedance and radiation pattern measurements. Fig. 50 shows a stripline-to-microstrip and then to-coax (not shown here) transition design, which was not optimized but provided a low reflection over the antenna bandwidth. The finite-ground stripline arm of the antenna was extended a straight length of $W_d / 2 - L_{ms} = 25$ mm, where W_d is the width of square substrate and L_{ms} is the length of the microstrip region created to solder the edge-fed SMA probe available for measurement. The dimension of 50 Ω microstrip is $W_{ms} = 1.5$ mm with larger ground width of $W_{msg} = 20$ mm to support the field distribution of microstrip mode. Two metallic vias of radius 0.615 mm at the step join the stripline grounds in a tapered section of stripline ground plane of which dimensions tapers from W_g to $W_{ge} = 8$ mm over a length $L_{ext} = 5$ mm.



Fig. 50 Transition design to enable measurement.

Fig. 51 shows the simulated VSWR and AR of the stripline Archimedean spiral antenna with transition design. The VSWR is still below 1.5 over wide frequency bands because of the impedance matching design between stripline and microstrip. However, the AR degrades a lot due to the unwanted radiation from this transition region. The degradation of AR due to this mode transition cannot be mitigated by increasing the spiral turns and can only be compensated by the optimization of transition design. Nevertheless, if the antenna is connected to system directly by stripline the degradation of AR can be mitigated.



Fig. 51 Simulated VSWR and axial ratio (AR) for stripline Archimedean spiral antenna with a single-turn tapered stripline impedance transformer.

5.4.3 Measurements Results

Fig. 52 shows the fabricated antenna by using two Rogers RT/duroid 5880 substrates of thickness 0.508 mm (20 mil) to implement stripline structure and its simulation model with the SMA connector. Seven #10-32 Nylon fasteners are used for mechanical alignment and compression with a thin layer of PDMS of $\varepsilon_r = 3.1$ and tan = 0.05 to bond the two 20-mil substrates together, as shown in subfigure (b) of Fig. 53.

The measured VSWR is shown in Fig. 53, the discrepancy is observed due to the non-ideal fabrication compared to simulation model but it still remains below 2 over very wide frequency range from 2 GHz to 20 GHz. Basically, the simulated and measured VSWR are in good agreement.



Fig. 52 Photo of fabricated antenna and the simulation model with probe effect.



Fig. 53 Measured and simulated VSWR. Sub-figure: (a) ideal embedded stripline conducted in simulation, and (b) elimination of air-gap effect by PDMS, where g_a is the air gap thickness between two substrates.

Fig. 54 shows the measured and simulated radiation patterns of the fabricated antenna at 3 GHz, 5 GHz, 10 GHz, and 15 GHz each in the xz and yz elevation cutplanes. The little discrepancies between measurement and simulation are caused by the non-ideal fabrication. In addition to that, they can also be explained by the supporting material presented around the antenna during the measurement process and spurious radiations caused by the feed cable.

After the measured radiation pattern is obtained, the calculation of AR can be performed on the time-phase difference between the two orthogonal components of field [118]. For a circular polarization antenna, the radiation field can be decomposed into two components: a right-handed circular polarization amplitude component E_R and a lefthanded circular polarization amplitude component E_L , which can be obtained from the radiation field components in spherical coordinate system (r, θ , φ) by

$$E_{R} = \frac{1}{\sqrt{2}} \left(E_{\theta} + j E_{\varphi} \right) \tag{56}$$

$$E_{L} = \frac{1}{\sqrt{2}} \left(E_{\theta} - j E_{\varphi} \right) \tag{57}$$

where both E_{θ} and E_{φ} are complex values including phase information. The axial ratio is then given by [135]

$$AR = \left| \frac{\rho + 1}{\rho - 1} \right| \tag{58}$$

where $\rho = E_R / E_L$.

Fig. 55 shows the simulated and measured AR. Note that the measured AR is obtained from 2.6 GHz to 18 GHz since the probe horn antennas used at the anechoic

chamber of the Texas A&M University only cover these frequency bands. The fabricated antenna cannot reveal a good circular polarization antenna due to several reasons as discussed before. Furthermore, the discrepancies can also be explained by the measurement method we adopted; for testing nearly circularly polarized antennas the rotating-source method is better than calculation from the orthogonal components of electric field since it obtains the axial ratio directly.


Fig. 54 Measured and simulated elevation (xz and yz) radiation patterns at (a) 3 GHz, (b) 5 GHz, (c) 10 GHz, and (d) 15 GHz.



Fig. 55 Measured and simulated AR.

5.5 Further Designs and Discussions

The design flow chart for stripline Archimedean spiral antenna was proposed in Fig. 56. There are some considerations have to be noted. For a chosen substrate, there is limitation on desired impedance. If the desired impedance is too high, the inner conductor width W_{ic} of stripline will be too narrow to be fabricated. If the desired impedance is too small, the arm spacing *S* of slot line will also be too small to be fabricated. Therefore, the characteristic impedance of stripline Archimedean spiral antenna cannot be chosen too high or too small depending on the supporting substrate used. However, as our design examples show, the impedance transformer can be easily integrated into the antenna structure of stripline Archimedean spiral antenna to achieve impedance matching to standard system impedance of 50 Ω no matter what substrate used.

To further validate the proposed design procedure, two additional stripline Archimedean spiral antenna designs on Rogers RT/duroid 6010 substrate ($\varepsilon_r = 10.2$) with different thickness are presented in Fig. 57, where the structural parameters are also summarized. In Fig. 57(a), the design is performed with a desired impedance of 40 Ω on a substrate of thickness h = 0.635 mm and Fig. 57(b) shows the design with desired impedance of 50 Ω on a substrate of thickness h = 0.635 mm. Both designs do not integrate with impedance transformer and have similar behaviors on VSWR below 2 over very wide frequency range from about 2.5 GHz to over 20 GHz.

More designs are conducted through extensive simulations on various substrate stack-up with higher dielectric constant and thicker substrate height but not shown in this paper. With higher dielectric constant and thinner substrate, the dimension of stripline design can be reduced and thus the antenna size; contrarily, with a thicker substrate, the dimension of stripline design have to be enlarged to confine the stripline mode and thus the antenna size is increased. Adding spiral turns can also improve the bandwidth, VSWR and AR (not shown here) at expense of larger antenna size.



Fig. 56 Design flow chart for the stripline-fed Archimedean spiral antennas.



Fig. 57 Further design examples of stripline Archimedean spiral antennas on (a) Rogers RT/duroid 6010 substrate of thickness h = 0.635 mm (25 mil) and (b) Rogers RT/duroid 6010 substrate of thickness h = 0.254 mm (10 mil).

CHAPTER VI

APPLICATIONS: DEVELOPMENT OF A DISC-SHAPED UAV

Unpiloted aerial vehicles (UAVs) and other remotely-operated or autonomous systems have played increasingly important roles in military and homeland security applications. They have also provided very useful platforms to monitor environmental conditions and survey the impact of natural disasters. Wireless connectivity remains one of the key features that enhance their operational functionality, but it also provides a challenging situation when throughput from multiple video, communication, and other high data-rate sensors require a substantial bandwidth. The variety of form factors and sizes of UAVs can inadvertently play a role in limiting this functionality since they have traditionally been based on fixed-wing or rotary-wing aircraft. Morphing-wing, flapping, and flying-wing configurations are adding to the diversity of UAV structures, but in many of these smaller platforms (such as those which can be hand-launched) the requirements for aerodynamic functionality can place limitation on the real estate available for antenna integration and restrict the performance of the wireless systems. This chapter will provide an overview of work on the development of an aerodynamically-functional broadband antenna-based UAV design in which an inwardfed stripline spiral antenna performs entirely or as part of the chassis in a disc-shaped flying-wing UAV.

A conformal mapping analysis which has been developed in Chapter II to accurately predict the input impedance of the spiral when it is operating in the lowest order balanced spiral mode is applied here to design the spiral antenna. The concept of two in-phase traveling guiding modes inspired the design of snail-type stripline-based topology. Next, the design of the flying wing structure is provided, which includes a brief overview on its construction and aerodynamic performance. The antenna performance, aerodynamic operation, and potential for collaborative beam-forming in volumetric swarms are discussed.

6.1 Snail-Type Stripline Spiral Antenna

The conformal mapping analysis derived in Chapter II can be used in the design process to synthesize an antenna with a desired value of characteristic impedance. This has been done for a desired characteristic impedance of 90 Ω by generating a family of design curves using a substrate with $\varepsilon_r = 2.2$ and h = 0.508 mm. The resulting antenna is designed with W = 4 mm and S = 1 mm (for 92 Ω); this value is set to be greater than the impedance Z_0 of the measurement system and is used to determine the closely-matched stripline feed structure. Fig. 58 illustrates the cross-section of the stripline spiral antenna. This is synthesized by transforming one arm of the gap-fed spiral into a finite-ground stripline feed structure. The top and bottom grounds are formed by moving Arm 1 from its substrate-embedded position z = 0 to z = -h (the bottom of the stack) and creating a copy at z = h (the top of the stack); these act as top and bottom grounds in the finiteground stripline feed structure. A signal line of width $W_I = 0.34$ mm at z = 0 completes the transformation of Arm 1 to a stripline feed network. It is centered and conformal to the spiral winding of the top and bottom grounds, and extends through the feed gap until it connects with Arm 2, as illustrated in Fig. 58 (right). The width W_1 corresponds to the impedance $Z_1 = 90 \ \Omega$ and the triangular sections at the center of Arms 1 and 2 perform the mode-converter between the guided (stripline) and radiating (spiral) modes.



Fig. 58 Stripline spiral antenna: (middle) nominal cross-section of the stripline-fed topology, (left) stripline grounds formed by Arm 1 at z = -h and h, and (right) stripline signal line (Arm 2) at z = 0.

Previous analysis for spiral antenna has shown that radiation occurs when the two slot-line modes are in-phase. However, when two slot-line modes propagate to the end of spiral arm, one slot-line mode stops and another slot-line mode keep propagating. The purpose of "snail-type" antenna shown in Fig. 59 (middle) is to extend this radiating mode of the spiral without significantly increasing the spiral dimensions. The snail-type antenna has similar size as spiral antenna with better VSWR, bandwidth, and axial ratio (not shown here). Fig. 59 (right) shows the design of stripline impedance transformer inside the spiral antenna structure matches Z_I to $Z_0 = 50 \Omega$ of the measurement system \Box . The taper transitions inward from $W_0 = 0.76$ mm at the perimeter of Arm 1 along one full winding of the spiral into W_I .



Fig. 59 Overlay of stripline spiral antenna (left), stripline snail-type antenna without a tapered impedance transformer (middle), and stripline snail-type antenna with a one-turn tapered impedance transformer (right).

The desired bandwidth for the UAV application extends from the lower end of the UHF band (2 MHz) into the mid-range of the VHF bands (20 GHz). Achieving this lower bound can be accomplished by extending number of turns to increase the outer radius of the spiral, but the authors' fabrication capabilities for this are currently limited to a 10"x10" square of substrate which can be chemically etched. Thus, a bolt-in modular design is considered here and a stripline snail-type spiral with N = 10 has been designed extending to lower frequency of 400 MHz. Fig. 60 shows the fabricated antenna with a stripline-to-microstrip and then to-coax transition design, which was not optimized but provided a low reflection over the antenna bandwidth. Fig. 61 shows the measured VSWR remaining below 2 over very wide frequency range from 400 MHz to over 20 GHz. Fig. 62 shows the comparison of measured and simulated results below 2 GHz. The discrepancy and oscillation are observed due to the non-ideal fabrication compared to simulation model but it still shows a good agreement. Fig. 63 shows a sample of the simulated and measured radiation patterns in S-band.



Fig. 60 Fabricated stripline snail-type antenna and microstrip transition.



Fig. 61 Measured VSWR from 0.1 GHz to 20 GHz.



Fig. 62 Measured VSWR compared with simulated VSWR up to 2 GHz.



Fig. 63 Sample of measured and simulated radiation patterns in S-Band for the snail-type stripline spiral antenna.

6.2 Disc-Shaped Flying Wing Design

The construction of the disc-shaped aerodynamic platform represents the next step in the prototype UAV development. The dimensions for this structure are intentionally set to be larger than the modular bolt-in design of the snail-type spiral so additional components can be placed on the platform without detuning or interfering with the operation of the spiral. Modifications to structurally functionalized antenna designs are used in subsequent designs to prevent this detuning and increase the size and bandwidth. For this design, however, the structural body consists of a Coroplast disk (ε_r \sim 1.25) which has a 12 in radius (cut from a 24 in x 24 in sheet of stock material). An irregular convex pentagon Coroplast tail fin (approx. 6 in. long and 5 in. tall) is fastened to the body on its narrow edge using a 0.375 in. thick basswood mount on the back of the disc along its centerline; this structure helps to control undesired rolling during flight. Two 6 in. cuts are then made parallel and 1 in. from the centerline on both sides of the tail fin then continued outward and perpendicular to the centerline to the edge of the disc on both sides. These body modifications form the two elevons for yaw, pitch, and roll control while maintaining the overall disc shape. Nylon hinges and mounting hardware are used to attach these ailerons to the main body. Fig. 64 shows a notional diagram of chassis, control surfaces, and several other elements of the design. This includes the the motor mount (also created from 0.375 in. thick basswood) and carbon fiber composite push-rods, which are used to avoid placing metallic structures above the antenna when connecting to the control horns on the elevons.



Fig. 64 Chassis and control surfaces of the UAV

Fig. 65 shows the final design with the mounted snail-type stripline spiral antenna; it connects directly to the six-channel ISM band radio (2.4 GHz) through an SMA-to-UNCC connector to wirelessly control the UAV. A 40 A brushless DC motor connects attaches directly to the front edge of the disc via the basswood motor mount and powers a 10x6 true-pitch tapering-tip maple propeller. All other components (electronic speed controller, servos, 900 MHz video system, etc.) attach directly to the body and are powered by a 2100 mAh lithium-polymer (LiPo) battery. The motor was selected to produce an adequate amount of thrust for the weight of the final UAV design. There is very little lift generated at a 0° (horizontal) angle of attack due to the flattened body shape so most of the lift is achieved by powering the UAV to a higher angle of attack where smaller pressures are produced on the top side. However, the UAV is

capable of hovering in place because most of the lift is provided by the thrust of the propeller.



Fig. 65 Completed flying disc with modularly integrated antenna.

Most of the weight in this configuration arises from the motor and the LiPo battery, so the front of the battery is located about two inches from the leading edge and centered behind the motor mount. This yields a center of gravity 6 inches behind the motor mount, or a 25% chord length from leading edge. These dimensions are assigned based on a trade study performed for similar RC planes and the battery is mounted on the bottom face to increase roll stabilization. All UAV components are mounted above the body with the exception of the battery and motor. The servos are placed about 6 inches from the front of the UAV and at equal distances of 5 inches from the centerline. The receiver and electronic speed control are mounted between the motor mount and spiral antenna slightly left of center in order to counter the clockwise torque induced by the propeller and motor. Fig. 66 and Fig. 67 show the measured VSWR and radiation patterns of the spiral after mounting all of the aforementioned components onto the disc. Based on these results it can be concluded that the operation of the antenna is maintained. Simulations of this completed structure were not available due to the numerical complexity introduced by the additional components placed on the structure.



Fig. 66 Measured VSWR of the snail-type spiral antenna after mounting onto the UAV.



Fig. 67 Sample of measured radiation patterns in S-Band for the snail-type stripline spiral antenna after mounting onto the UAV.

CHAPTER VII ANALYTIC ANALYSIS OF MULTILAYERED PERIODIC COPLANAR WAVEGUIDES

Coplanar waveguides (CPW) have been widely used in the design of monolithic microwave integrated circuits (MMIC) [64], [65] and printed circuit board (PCB) [66]. CPW offers several advantages over microstrip transmission line [63] including lower dispersion and easily access to ground. The ideal configuration for CPW consists of a conductor strip of width W on top of an infinitely thick dielectric substrate with two semi-infinite ground conductors on both sides of a spacing S, as shown in Fig. 68(a). The first analytic formula for this configuration is given by Wen [67] using conformal mapping to estimate quasi-TEM wave properties. In actual implementation, the CPW structures are neither of infinite substrate [90] nor infinite lateral extent [80], as shown in Fig. 68(b), where the ground strip width and the substrate height are denoted as W_g and h, respectively. For the sake of maximizing the circuit density, the ground plane width should be as small as possible, but truncating lateral ground planes and placing CPW close to each other also increase line-to-line coupling [64] and change the characteristic impedance of CPW. The coupling effects of coupled CPW and parallel CPW shown in Fig. 68(c) are discussed in [89] and [64], respectively. Nowadays, since most systems use signaling interfaces in which large numbers of transmission lines including CPW are routed in parallel through packages, connectors, and interconnect inside integrated chips,

the coupling can play an important role in determining the performance of the system and the investigation on parallel CPW is not enough.



Fig. 68 Configuration of coplanar waveguides, (a) CPW with infinite lateral ground plate, (b) CPW with finite lateral ground plate, (c) parallel CPW, and (d) periodic CPW.

Many systems currently use signaling interfaces in which large numbers of transmission lines (including CPW) are routed in parallel through packages, connectors, and interconnect inside integrated chips. In order to maximize the circuit density, the ground conductor width should be as small as possible, but truncating lateral ground planes and placing CPW close to each other also increase line-to-line coupling [64]. This can play an important role in determining the performance of the system.

This paper analyzes a circumstance in which several CPW transmission lines are parallel due to highly demand on miniaturization of circuit design. The analysis of characteristic impedance, propagation delay, and coupling effect of a target line in a multiple transmission line environment (as an equivalent single transmission line) is a common treatment during the initial design of a full system, and followed here. For simplification and to investigate the properties of several parallel CPWs, the configuration called periodic coplanar waveguide (PCPW) with identical dimension in each periodic cell as shown in Fig. 69 is a suitable model to analyze. The configuration of PCPW consists of a conductor strip of width *W* on top of dielectric substrate of ε_r and thickness *h* with two lateral ground conductors of width W_g on both sides of a spacing *S*. This work is different from the periodic CPW structures in [136]-[139], which could be described as periodic loaded CPWs since the properties of CPW are modified by periodically loaded circuit elements.

Two contrasting propagating modes will be examined in PCPW structure. The first mode corresponds to the common mode where all the signal strips are switching inphase, as illustrated in Fig. 69 (a). The other mode is called the differential mode (Fig. 69 (b)) where adjacent signals are switching out-of-phase. In this paper, the conformal mapping method is applied to develop quasi-static design formulas for quickly evaluating the even- and odd-mode characteristic impedance and effective dielectric constant of a CPW transmission line in PCPW structures. These formulas are useful for computer-aided design (CAD) tools. The results calculated by the proposed expressions are in excellent agreement with those obtained by full-wave simulation and experimental data.



Fig. 69 Cross-sectional view of a PCPW in (a) even (common) mode and (b) odd (differential) mode.

7.1 Conformal Mapping Analysis of PCPW

Fig. 70(a) shows the periodic section of the PCPW used for the analysis, where the boundary plane between two adjacent sections can be modeled as perfect magnetic walls (PMC) for even mode propagation or as perfect electric walls (PEC) for odd mode propagation. The length of periodic section is denoted as $L_p = 2S + W + W_g$. For the sake of clarity, a single-layered substrate PCPW (as shown in Fig. 70(a)) is sufficient to explain conformal mapping steps, while the theory presented below is applicable to the general case of multilayered structures with or without conductor-backed plates. The assumption for quasi-TEM wave propagation requires the dimensions of PCPW $L_p \ll \lambda_g/2$ (λ_g is guided wavelength), which is usually valid for the application of packaging or microwave circuit design. To simplify analysis without losing generality, the metal strips have zero thickness with perfect conductivity and the dielectric layers are lossless.

In the quasi-TEM approximation, the characteristic impedance Z_0 and effective dielectric constant ε_{reff} can be completely determined from the effective per-unit-length capacitance *C* when the dielectric material is present and the per-unit-length capacitance C_0 with no dielectric material [119]. The partial capacitance technique [78], [80], [104], which models the air/dielectric and dielectric/dielectric interfaces as PMC, is used to evaluate the effective per-unit-length capacitance *C* of PCPW in (59), where C_1 represents the per-unit-length capacitance when the partial charges reside only inside the dielectric layer. In the case of a multilayered substrate, the limitation of partial capacitance approach is discussed in [105].

$$C = C_0 + C_1 \tag{59}$$

7.1.1 Common Mode (Even Mode)

For the common mode propagation, the plane of boundary of periodic section is PMC. Fig. 70(b) shows the conformal mapping steps for evaluation of C_{e0} . The transformation (60) maps the *z*-plane of PCPW onto the upper half *t*-plane and then transformation (61) maps the upper half *t*-plane onto a rectangle in the *w*-plane, where t_c = $t(z_c)$, $t_d = t(z_d)$, and $z_c = S + W/2$ and $z_d = W/2$ are the coordinates of *z*-plane.

$$t = -\cos\left(\frac{2\pi z}{L_p}\right) \tag{60}$$

$$w = \int^{t} \frac{dt'}{\sqrt{(t'^2 - 1)(t' - t_c)(t' - t_d)}}$$
(61)

The per-unit-length capacitance of air-filled PCPW is given by (62), where $K(k_{e0})$ is the complete elliptic integral of first kind with the modulus k_{e0} and the complementary modulus k_{e0} ' expressed in (63) and (64), respectively.

$$C_{e0} = 4\varepsilon_0 \frac{K(k_{e0})}{K(k_{e0}')}$$
(62)

$$k_{e0} = \cot\left(\frac{\pi}{2}\frac{2S+W}{L_p}\right)\tan\left(\frac{\pi}{2}\frac{W}{L_p}\right)$$
(63)

$$k_{e0}' = \sqrt{1 - k_{e0}^2} \tag{64}$$

Fig. 70(c) shows the conformal mapping steps for evaluation of C_{e1} . The transformation (65) maps the *z*-plane of PCPW onto the upper half *t*-plane, where $sn(u,r_1)$ is the Jacobian elliptic function with a variable *u* and a modulus r_1 determined by (66) and (67), and $K_1 = K(r_1)$ can be evaluated by (68).

$$t = \operatorname{sn}^{2} \left(K_{1} \frac{2z}{L_{p}}, r_{1} \right)$$
(65)

$$q_1 = \exp\left(-\pi \frac{2h_1}{L_p}\right) \tag{66}$$

$$r_{1} = 4\sqrt{q_{1}} \prod_{n=1}^{\infty} \left(\frac{1+q_{1}^{2n}}{1+q_{1}^{2n-1}} \right)^{4}$$
(67)

$$K_1 = \frac{\pi}{2} + 2\pi \sum_{s=1}^{\infty} \frac{q_1^s}{1 + q_1^{2s}}$$
(68)

Then the transformation (69) maps the upper half t-plane onto a rectangle in the w-plane.

$$w = \int^{t} \frac{dt'}{\sqrt{t'(t'-1)(t'-t_c)(t'-t_d)}}$$
(69)

The per-unit-length capacitance C_{e1} is given by (70), where the modulus k_{e1} and the complementary modulus k_{e1} ' is given by (71) and (72), respectively.

$$C_{e1} = 2\varepsilon_0 \left(\varepsilon_{r1} - 1\right) \frac{K(k_{e1})}{K(k_{e1}')}$$
(70)

$$k_{e1} = \frac{\operatorname{sn}\left(K_{1} \frac{W}{L_{p}}, r_{1}\right)}{\operatorname{sn}\left(K_{1} \frac{2S + W}{L_{p}}, r_{1}\right)} \sqrt{\frac{1 - \operatorname{sn}^{2}\left(K_{1} \frac{2S + W}{L_{p}}, r_{1}\right)}{1 - \operatorname{sn}^{2}\left(K_{1} \frac{W}{L_{p}}, r_{1}\right)}}$$
(71)
$$k_{e1}' = \sqrt{1 - k_{e1}^{2}}$$
(72)

The total per-unit-length capacitance is the sum of all partial capacitances and can be expressed as (73), where the effective dielectric constant is given in (74).

$$C_{e} = 4\varepsilon_{0}\varepsilon_{reff}^{\text{even}} \frac{K(k_{e0})}{K(k_{e0}')}$$
(73)

$$\varepsilon_{reff}^{\text{even}} = 1 + \frac{1}{2} (\varepsilon_{r1} - 1) \frac{K(k_{e1})}{K(k_{e1}')} \frac{K(k_{e0}')}{K(k_{e0})}$$
(74)

The common-mode characteristic impedance is

$$Z_0^{\text{even}} = \frac{30\pi}{\sqrt{\varepsilon_{reff}}} \frac{K(k_{e0}')}{K(k_{e0})}$$
(75)

7.1.2 Differential Mode (Odd Mode)

For the differential mode propagation, the plane of boundary of periodic section is PEC. As shown in Fig. 70(b), the transformation from *z*-plane to *t*-plane is the same as (60), while the transformation (76) maps the upper half *t*-plane onto a rectangle in the *w*-plane.

$$w = \int^{t} \frac{dt'}{\sqrt{(t'+1)(t'-t_c)(t'-t_d)}}$$
(76)



Fig. 70 Conformal mapping for evaluation of per-unit-length capacitance, (a) periodic section of PCPW, (b) transformation steps for C_0 , and (c) transformation steps for C_1 .

The per-unit-length capacitance of air-filled PCPW is given by (77) with the modulus k_{o0} and the complementary modulus k_{o0} ' expressed in (78) and (79), respectively.

$$C_{o0} = 4\varepsilon_0 \frac{K(k_{o0})}{K(k_{o0}')}$$
(77)

$$k_{o0} = \sin\left(\frac{\pi}{2}\frac{W}{L_p}\right) / \sin\left(\frac{\pi}{2}\frac{2S+W}{L_p}\right)$$
(78)

$$k_{o0}' = \sqrt{1 - k_{o0}^2} \tag{79}$$

For the calculation of C_{o1} as shown in Fig. 70(c), the transformation in (65)-(68) maps *z*-plane to *t*-plane and then the transformation (80) maps the upper half *t*-plane onto a rectangle in the *w*-plane.

$$w = \int^{t} \frac{dt'}{\sqrt{t'(t'-t_{a})(t'-t_{c})(t'-t_{d})}}$$
(80)

The per-unit-length capacitance C_{o1} is given by (81)-(83).

$$C_{o1} = 2\varepsilon_0 \left(\varepsilon_{r1} - 1\right) \frac{K(k_{o1})}{K(k_{o1}')}$$
(81)

$$k_{o1} = \frac{\operatorname{sn}\left(K_{1}\frac{W}{L_{p}}, r_{1}\right)}{\operatorname{sn}\left(K_{1}\frac{2S+W}{L_{p}}, r_{1}\right)} \sqrt{\frac{1 - r_{1}^{2}\operatorname{sn}^{2}\left(K_{1}\frac{2S+W}{L_{p}}, r_{1}\right)}{1 - r_{1}^{2}\operatorname{sn}^{2}\left(K_{1}\frac{W}{L_{p}}, r_{1}\right)}}$$
(82)
$$k_{o1}' = \sqrt{1 - k_{o1}^{2}}$$
(83)

The total per-unit-length capacitance for differential mode propagation is expressed as (84), where the effective dielectric constant ε_{reff} is given in (85).

$$C_{o} = 4\varepsilon_{0}\varepsilon_{reff}^{\text{odd}} \frac{K(k_{o0})}{K(k_{o0}')}$$
(84)

$$\varepsilon_{reff}^{\text{odd}} = 1 + \frac{1}{2} \left(\varepsilon_{r1} - 1 \right) \frac{K(k_{o1})}{K(k_{o1}')} \frac{K(k_{o0}')}{K(k_{o0})}$$
(85)

The differential-mode characteristic impedance is

$$Z_0^{\text{odd}} = \frac{30\pi}{\sqrt{\varepsilon_{reff}}} \frac{K(k_{o0}')}{K(k_{o0})}$$
(86)

The coupling coefficient can be evaluated by the expression

$$C_{\varrho} = \frac{Z_0^{\text{even}} - Z_0^{\text{odd}}}{Z_0^{\text{even}} + Z_0^{\text{odd}}}$$

$$\tag{87}$$

7.1.3 Multilayer Extension

Fig. 71 shows a PCPW structure embedded in a set of infinite number of dielectric layers. The parallel partial capacitance approximation can be easily extended to multilayer cases as demonstrated in [104]. The limitation of partial capacitance technique is discussed in [105], which states that the parallel partial capacitance (PPC) approach is valid only when the dielectric constant is decreasing layer by layer away from the PCPW structure. When the dielectric constant is increasing away from the PCPW structure, the series partial capacitance (SPC) approach can be used to evaluate the quasi-TEM properties. Despite this limitation, the analysis in this section is still applicable to most applications.

For the common mode, the total per-unit-length capacitance and the characteristic impedance are shown in (73) and (75), respectively, with the effective dielectric constant redefined in (88)-(93).

$$\varepsilon_{reff}^{\text{even}} = 1 + \sum_{i} \xi_{ei} \varepsilon^{(i)}, \quad i = \begin{cases} t1, \dots, tm; & \text{for top layers} \\ b1, \dots, bn; & \text{for bottom layers} \end{cases}$$
(88)

$$\xi_{ei} = \frac{1}{2} \frac{K(k_{ei})}{K(k_{ei}')} \frac{K(k_{e0}')}{K(k_{e0})}$$
(89)

$$k_{ei} = \frac{\operatorname{sn}\left(K_{i} \frac{W}{L_{p}}, r_{i}\right)}{\operatorname{sn}\left(K_{i} \frac{2S+W}{L_{p}}, r_{i}\right)} \sqrt{\frac{1-\operatorname{sn}^{2}\left(K_{i} \frac{2S+W}{L_{p}}, r_{i}\right)}{1-\operatorname{sn}^{2}\left(K_{i} \frac{W}{L_{p}}, r_{i}\right)}}$$
(90)

$$r_{i} = 4\sqrt{q_{i}} \prod_{n=1}^{\infty} \left(\frac{1+q_{i}^{2n}}{1+q_{i}^{2n-1}}\right)^{4}, \quad q_{i} = \exp\left(-\pi \frac{2h^{(i)}}{L_{p}}\right)$$
(91)

$$\varepsilon^{(i)} = \varepsilon_{ri} - \varepsilon_{r(i+1)}, \quad \text{with} \begin{cases} \varepsilon_{rt(m+1)} = 1\\ \varepsilon_{rb(n+1)} = 1 \end{cases}$$
(92)

$$h^{(i)} = \begin{cases} \sum_{i=t1}^{tm} h_i, & \text{for top layers} \\ \sum_{i=b1}^{bn} h_i, & \text{for bottom layers} \end{cases}$$
(93)



Fig. 71 PCPW in multilayered dielectric structure.

For the differential mode, the total per-unit-length capacitance and the characteristic impedance are shown in (84) and (86), respectively, with the effective dielectric constant redefined in (94)-(96).

$$\varepsilon_{reff}^{\text{odd}} = 1 + \sum_{i} \xi_{oi} \varepsilon^{(i)}, \quad i = \begin{cases} t1, \cdots, tm; & \text{for top layers} \\ b1, \cdots, bn; & \text{for bottom layers} \end{cases}$$
(94)

$$\xi_{oi} = \frac{1}{2} \frac{K(k_{oi})}{K(k_{oi}')} \frac{K(k_{o0}')}{K(k_{o0})}$$
(95)

$$k_{oi} = \frac{\operatorname{sn}\left(K_{i} \frac{W}{L_{p}}, r_{i}\right)}{\operatorname{sn}\left(K_{i} \frac{2S+W}{L_{p}}, r_{i}\right)} \sqrt{\frac{1 - r_{i}^{2} \operatorname{sn}^{2}\left(K_{i} \frac{2S+W}{L_{p}}, r_{i}\right)}{1 - r_{i}^{2} \operatorname{sn}^{2}\left(K_{i} \frac{W}{L_{p}}, r_{i}\right)}}$$
(96)

When the dielectric constant is increasing layer by layer away from the PCPW structure, the series partial capacitance (SPC) approach can be used to evaluate the quasi-TEM properties of PCPW.

7.2 Validation and Discussions

This section will present the calculated results by conformal mapping analysis to verify the derived expressions as well as to investigate the properties of PCPW. Although, to the author's best knowledge, there is no discussion on periodic coplanar waveguides and odd mode propagation in parallel CPW structure, comprehensive comparisons with the results of full-wave analysis and available experimental data in literatures will be presented in this section and demonstrate that the derived formulas are accurate for most of the application range of physical dimensions and available dielectric materials.

It is worth to note that all functions used in the analysis (e.g. elliptic integrals of first kind) are available in most mathematical software packages. The infinite series expansion functions of (67) and (68) converge quickly with n = 200 and s = 1000, respectively, with the computation time less than 1s in modern personal computer. When the modulus k is close to 1 or 0, the calculation of Jacobian elliptic function sn(u,k) can be simplified. For $2h_1/L_p < 0.1$ ($k \rightarrow 1$), the function is given by

$$\operatorname{sn}(u,k) = \tanh u \tag{97}$$

For $2h_1/L_p > 100 \ (k \rightarrow 0)$, the approximation is given by

$$\operatorname{sn}(u,k) = \sin u \tag{98}$$

Otherwise, the Jacobian elliptic function can be evaluated by

$$\operatorname{sn}(u,k) = \frac{2\pi}{kK(k)} \sum_{s=0}^{\infty} \frac{q_1^s}{1+q_1^{2s}}$$
(99)

7.2.1 Analysis of PCPW with Infinite Periodicity

Fig. 72 shows the characteristic impedance of single layered PCPW operating in even mode. The parameters used in this computation are $\varepsilon_{rb1} = 13$, $h_{b1} = 1.5$ mm, and W + 2S = 3mm. As expected, the agreement are excellent between the calculated results with $W_g = 10(W + 2S)$ and those from [140] using full wave analysis with assumption of infinite lateral ground. It is intuitive to see that the curves are dependent on the lateral ground width W_g . In this case, the PCPW with $W_g > 2(W + 2S)$ can be considered as an isolated CPW with infinite lateral grounds, where (W + 2S) can be denoted as the dimension of isolated CPW. As each CPW is placed closer than the dimension of isolated CPW in periodic structure, e.g. $W_g < (W + 2S)$, the coupling between transmission lines results in impedance change.



Fig. 72 Characteristic impedance of single layered PCPW with $\varepsilon_{rb1} = 13$, $h_{b1} = 1.5$ mm, and W + 2S = 3mm.

The experimental data for single layered CPW can be found in [67] with very thick substrate and in [90] with finite substrate thickness. Fig. 73 shows the comparison of calculated characteristic impedance obtained in this work with these experimental results. The parameters used in this calculation are listed in TABLE II with an assumption of large lateral ground of $W_g = 10(W + 2S)$. It is seen that our conformal mapping results agree with corresponding measured results within 2.6%. The little discrepancy observed in the Case A of thin substrate is due to modeling air/dielectric

interface as PMC. This can be explained by less field concentration in the substrate and partial field distribution in air resulting in a less accurate model of PMC on interface.

Case	А	В	С	D	Е
E _{rb1}	10	10	10	16	130
<i>W</i> +2 <i>S</i> [mm]	2	2	2	2	2
h_{b1} [mm]	S	3 <i>S</i>	∞	∞	∞

TABLE IIPARAMETERS OF PCPW FOR CALCULATION IN FIG. 73.



Fig. 73 Measured and calculated characteristic impedance of single layered PCPW with infinite periodicity.

Fig. 74 shows the characteristic impedance of PCPW on two- layered substrate, compared with the results taken from [141] using numerical spectrum domain method and from [65] using analytical conformal mapping method. The calculations are carried out with the parameters listed in TABLE III and large lateral ground width of $W_g = 10(W$

+ 2S). It shows that the analytic formulas derived in this paper can also be used in multilayered isolated CPW structures.

Case	E _{rb1}	ε_{rb2}	h_{b1} [μ m]	h_{b2} [μ m]	<i>S</i> [μm]
Α	12.9	3.78	200	∞	20
В	20	10	200	∞	100
С	12.9	10	200	∞	200

TABLE III PARAMETERS OF PCPW FOR CALCULATION IN FIG. 74.



Fig. 74 Characteristic impedance of two-layered PCPW with infinite periodicity.

7.2.2 Analysis of PCPW with Equal Strip Width

In this section, the PCPWs with fixed periodicity of W + S = 2.5 mm and equal strip width of $W_g = W$ are analyzed. The field distribution now has dependence on both substrate height h and the metallization ratio χ defined by (100). Specifically, for a fixed metallization ratio χ , smaller values of (W+S)/h correspond to smaller conductor spacing S compared to h and an increased field distribution within the substrate; this increases the total per-unit-length capacitance and decreases the characteristic impedance of the propagating modes.

$$\chi = \frac{W}{W+S} \tag{100}$$

The characteristic impedance in even- and odd-modes are calculated and compared with the simulated data from COMSOL Multiphysics [142], a proven commercial solver based on finite element method. The corresponding PCPW CAD model consists of nine unit cells of CPW. A range of commercially available substrate of various heights and dielectric constants, i.e. Roger Duroid 5880 and 6010 [143], is used to validate the proposed analytical formulas. The simulated characteristic impedance are obtained by (101), where per-unit-length capacitance *C* and C_0 are extracted from the center cell with/without dielectric material present, respectively, and *c* is the speed of light.

$$Z_0 = \frac{1}{c\sqrt{C \times C_0}} \tag{101}$$

TABLE IV PARAMETERS OF PCPW FOR CALCULATION IN FIG. 75.

Case	\mathcal{E}_{rt1}	E _{rb1}	h_{tl} [mil]	h_{b1} [mil]	W+S [mm]
Α	2.2	10.2	20	125	2.5
В	2.2	10.2	62	25	2.5



Fig. 75 Even- and odd-mode characteristic impedance of PCPW embedded in dielectric materials versus metallization ratio obtained by conformal mapping (CM) and finite element method (FEM).

Fig. 75 shows the even- and odd-mode impedance of PCPW embedded in two dielectric materials of different heights. The calculations are carried out with the parameters listed in TABLE IV. Fig. 76 shows the even- and odd-mode impedance of PCPW on top of two-layered substrate with the parameters listed in TABLE V. Overall, the conformal mapping results agree with the numerical simulations within an error of 3.5%, which is also acceptable for a numerical error. In the Case C of Fig. 76, the discrepancy is observed in the thin substrate due to the approximation of modeling dielectric/air interface as a PMC, while for higher metallization ratio, the higher field concentration inside the dielectric materials resulting in a more accurate PMC model on interface. In general, the discrepancy in odd-mode is larger than that in even-mode because of PEC boundary between two unit cells rendering more field distribution out of substrate.

As seen in these analyses, both even- and odd-mode impedance and the difference between them increase as the metallization ratio decreases. Fig. 77 shows the coupling coefficients decreases as metallization ratio χ increases.

Case h_{b1} [mil] h_{b2} [mil] W+S [mm] \mathcal{E}_{rb1} \mathcal{E}_{rb2} 2.5 10.2 2.2 15 31 С 10.2 2.2 62 62 D 2.5

PARAMETERS OF PCPW FOR CALCULATION IN FIG. 76.

TABLE V



Fig. 76 Even- and odd-mode characteristic impedance of PCPW residing on two-layer substrate of finite thickness obtained by conformal mapping (CM) and finite element method (FEM).


Fig. 77 Coupling coefficients as a function of metallization ratio.

7.2.3 Analysis of PCPW with Unequal Strip Width

In this section, a more general case is considered, in which all the dimensional parameters are normalized by the dimension of isolated CPW since the calculated impedances are not dependent on the size of W + 2S. This can be seen in the expression of modulus k_{eo} , k_{e1} , k_{o0} , and k_{o1} in (63), (71), (78), and (82), respectively. Note that the value of W/(W+2S) can also be described as the metallization ratio of an isolated CPW.

Fig. 78 and Fig. 79 show the even- and odd-mode characteristic impedance of PCPW on top of GaAs ($\varepsilon_{rb1} = 12.9$). For a given W_g , the increase in W increases the perunit-length capacitance and therefore reduces both the even- and odd-mode characteristic impedance. For a given W, the increase in W_g increases the unit cell marginally. This increases the coupling surface of the conductor in the case of even mode and thus increases the per-unit-length capacitance resulting in the reduction of line impedance, while in the case of odd mode, the line impedance is slightly increased because the distance of coupling to the PEC boundary of a unit cell are also increased, incurring reduced per-unit-length capacitance. For a large W_g , e.g. $W_g/(W+2S)=3$, the even-mode impedances coincide with the odd-mode ones and can be considered as the characteristic impedance of an isolated CPW with infinite lateral ground.

For a given W and W_g , the reduction of h_{b1} decreases the effective dielectric constants of the structure hence increases both the even- and odd-mode characteristic impedance of PCPW. It is worth to notice that both even- and odd-mode characteristic impedance reach saturation values for a thick substrate, where the field distribution can be considered as well-confined inside the substrate and not experienced the dielectric boundary. The saturated substrate thickness h_{b1} for characteristic impedance is dependent on each parameter, but as a rule of thumb, the PCPW can be assumed on top of infinite layer when the normalized substrate height of $h_{b1}/(2S + W)$ is larger than 1.



Fig. 78 Even-mode characteristic impedance of PCPW on top of GaAs versus $h_{bl}/(2S + W)$.



Fig. 79 Odd-mode characteristic impedance of PCPW on top of GaAs versus $h_{b1}/(2S + W)$.

7.3 Design Curves

A closed-form analytical solution has been given for obtaining the quasi-TEM properties of periodic coplanar waveguide (PCPW) structure, which is a simplified case for discussion of paralleling several CPW transmission lines. The presence of the other CPWs at both sides of one CPW induces the coupling and alters the transmission line properties. The two contrasting cases of even- and odd-mode are discussed in this paper. The analysis has led to accurate formulas for effective dielectric constants, per-unit-length capacitance, and characteristic impedances of one periodic cell in PCPW for these two modes. The calculated results generated by the proposed formulas are in excellent agreement with those results obtained by full-wave analysis and experimental data.

The derived formulas provide intuitive physical insight and are useful for miniaturized circuit application where the distance of acceptable coupling without altering transmission line impedance. The substrate thickness that can be considered as infinite substrate for characteristic impedance as well as per-unit-length capacitance is determined. This model can also be extended for analysis of guarding ground effects in circuit interconnects. Using this model, it would be possible for designers to find optimum line impedance for system performance and to reduce crosstalk in coupled CPW transmission lines. An example of design contour plot is also given in Fig. 80.



Fig. 80 Design curves for the even-mode characteristic impedance as a function of W_g and h_{b1} for a PCPW with metallization ratio of 0.5 on top of GaAs.

CHAPTER VIII

CONCLUSIONS

This work has covered many different aspects on the spiral antenna. It begins by modifying the original band theory of spiral antennas, developing a periodic coplanar waveguide (PCPW) model, and then deriving the analytical formulas using conformal mapping for evaluating the characteristic impedance of spiral antenna operating in radiation region. The derivation provides the preliminary work in support of the aperchassis (a radiating structure with structural functionality and volumes/surfaces for internalized subsystems). The stripline Archimedean spiral antenna design provides an internalized feed network and leaves surfaces and volumes to integrate subsystems to enable multifunctionality and thereby aid in the mitigation of size and/or weight constraints. Measured and simulated results in the first and second design example demonstrated the basic design and operation of the structure. Future and ongoing work will examine the structure in more detail. In addition to this, the effects of structural deformations are also under investigation.

8.1 Summary

In Chapter II, a simple method and analytical closed-form expressions for the characteristic impedance of a periodic coplanar waveguide (PCPW) and an Archimedean spiral antenna have been obtained using conformal mapping techniques. These expressions are quasi-static providing an accurate prediction for the impedance at lowest

operating frequency with no substrate and the band-average impedance with present embedded substrate. The comparison between the calculated results and simulated results using numerical software verifies the accuracy of the expressions. This verification suggested that this analytical expression could evaluate the input impedance of a spiral antenna a lot faster. This is the first time the quasi-static analysis performed on spiral antenna to predict its input impedance operating in its radiation region and gives physical insight when the spiral arm width is non-negligible and non-selfcomplementary. This method can be easily extended to multi-layers structures.

In Chapter III, the design and fabrication of a stripline-fed two-arm Archimedean spiral antenna has been proposed using a simple concept, remaining the original impedance properties. The design transforms one arm of the planar gap-fed spiral into a stripline feed network, and improvements over prior work [61] on this topology include a tapered impedance transformer which remains conformal to a single turn of the spiral and an improved impedance bandwidth. Measured and simulated results showed good agreement for a fabricated version of the antenna, from which it can be concluded that the transformation from of the gap-fed spiral into the stripline-fed spiral does not significantly degrade the impedance performance of the spiral. Fig. 81 summarizes the design flow chart for design process I.



Fig. 81 The design flow chart for design process I.

In Chapter IV and V, the radiation mechanism and impedance properties of how stripline Archimedean spiral antenna radiates has been thoroughly examined. A design method of stripline Archimedean spiral antenna has been proposed for broadband operation with circular polarization based on a complete analysis. The antenna design can be achieved easily by impedance matching of stripline mode and parallel elevated slot line mode. The characteristic impedance of parallel elevated slot line structure is analytically derived by approximated PCPW model and conformal mapping methods. Measured and simulated results of design example were in good agreement for a fabricated version of the antenna. A design flow chart is drawn to summarize the design procedure after carrying out a design study. Two additional antennas have also been successfully designed for validating the proposed design procedure on arbitrary substrate stack-up.

In Chapter VI, one of the applications of stripline Archimedean spiral antenna is presented. The analytical solution derived by conformal mapping analysis is applied to accurately predict the input impedance of the spiral antenna when it is operating in the lowest order balanced spiral mode. According to the design process I developed in Chapter III, this analytical formula can help a stripline Archimedean spiral antenna design without the time-consuming numerical analysis. Based on the concept of two inphase traveling guiding modes incurring the radiation, the transformation of the spiral design into a snail-type stripline-based topology is proposed with better performance than the traditional spiral antenna. Next, the design of the flying wing structure is provided, which includes a brief overview on its construction and aerodynamic performance. The antenna performance, aerodynamic operation, and potential for collaborative beam-forming in volumetric swarms are discussed.

In Chapter VII, a closed-form analytical solution has been derived for obtaining the quasi-TEM properties of periodic CPW (PCPW) transmission lines. It has been shown that the presence of the other CPWs at both sides of one CPW induces the coupling between two CPWs and modifies the transmission line properties, especially for miniaturized circuit application where the distance of minimum coupling needed to be determined. Analytical results generated by the proposed formulas are in excellent agreement with those results obtained by full-wave analysis and experimental data. These formulas are accurate and easy to program in computers.

8.2 Contributions

- The first physical insight into the operation and radiation mechanism of an Archimedean spiral antenna with non-negligible metal strip width is proposed.
- The first comprehensive analysis based on physical speculation of periodic coplanar waveguide (PCPW) model and conformal mapping method predicts the characteristic impedance of spiral antenna when operating in its radiation region accurately. This close-form analytical formula is a lot more accurate than the results calculated from commercial numerical software package, HFSS.
- It is the first time that the input impedance of an Archimedean spiral antenna can be known based on its geometric parameters prior to time-consuming numerical calculation.
- A novel compact multifunctional antenna, a stripline-fed Archimedean spiral antenna, is designed, fabricated, and demonstrated, which has very broadband VSWR, axial ratio (AR), and broadside radiation pattern from 2 GHz to over 20 GHz.

- A stripline Archimedean spiral antenna is a totally planar spiral antenna design without the need of complex wideband balun circuits and matching circuit compared to the traditional vertically-fed spiral design, which is a lot beneficial to array placement and fabrication.
- The first simple design process is proposed for novel spiral antenna without the aid of numerical software, based on a very simple concept of impedance matching.
- The stripline structure can easily provide isolation on antenna properties and integrate circuit elements inside the antenna.
- The first stripline Archimedean snail-type antenna is designed for the better performance to the stripline Archimedean spiral antenna.
- The first analytical solution for periodic coplanar waveguides is derived. The minimum ground width without effect on the properties of adjacent coplanar waveguides can be decided by this work.

8.3 Future Work

Measured and simulated results in the first and second design example demonstrated the basic design and operation of the stripline Archimedean spiral antennas. Future and ongoing work will examine the structure in more detail with developing the transmission line model for the Archimedean spiral antenna. In addition to this, the effects of structural deformations and the integration of microwave circuits are also under investigation. Moreover, the application of conformal mapping analysis can be extended to discuss the effect of guarding traces on microstrip and microstrip coupled capacitors.

8.3.1 Transmission Line Model for Archimedean Spiral Antenna

The transmission line mode is first proposed in 2007 to account for the impedance properties of spiral antenna [57]. From the band theory, the active region of spiral antenna is like a circular loop antenna. Thus, the spiral antenna can be modeled as a cascade set of loop antenna with their radiation resistance to account for the radiation in the transmission line model, as shown in Fig. 82. For each circular loop antenna, we have the radiation resistance as

$$R_{rad} \approx 5120 \left(\frac{\pi r}{\lambda}\right)^4$$
 (102)

The P.U.L. resistance in transmission line model is obtained by

$$R = \frac{R_{rad}}{2\pi r} \approx 2560\pi^3 \frac{r^3}{\lambda^4}$$
(103)

If we unwrap the spiral as shown in Fig. 83, each section of transmission line can be represented a circular loop antenna at different frequency, where the characteristic impedance and the propagation constant are

$$Z_0 = \sqrt{\frac{R + j\omega L}{j\omega C}}$$
(104)

$$\gamma = \sqrt{j\omega RC - \omega^2 LC} \tag{105}$$

The input impedance of each section of transmission line is

$$Z_{x+\Delta x} = Z_0 \frac{Z_x + Z_0 \tanh(\gamma \Delta x)}{Z_0 + Z_x \tanh(\gamma \Delta x)}$$
(106)

After cascading, Fig. 84 shows the comparison of input impedance calculated by transmission line mode and HFSS. However, Lee *et. al* did not explain where the inductance and capacitance comes from. Now, the conformal mapping derivation can be applied to complete this transmission line model.



Fig. 82 Decompose the spiral antenna into small current loop antenna.



Fig. 83 Cascade each current loop antenna as a transmission line model.



Fig. 84 Comparison of transmission line model and numerical results.

8.3.2 Integrated Filter Design

To demonstrate that the circuit elements can be put inside the antenna structure without impacts on antenna performance, an edge-coupled microwave bandpass filter as shown in Fig. 85 is designed in stripline stack-up for this purpose. Fig. 86 shows the frequency response is at 9.75 GHz. In order to put the filter inside the spiral structure, the filter design is modified to occupy less footprint. Fig. 87 shows the modified filter has the same frequency response as the original one.

Fig. 88 shows the filter is conformal integrated into the antenna structure and Fig.89 shows the simulated VSWR of the antenna structure in Fig. 88.



Fig. 85 Edge-coupled stripline bandpass filter design.



Fig. 86 Frequency response of the filter design.



Fig. 87 The filter design in order to be conformal integrated into the stripline-fed Archimedean spiral antenna.



Fig. 88 The spiral conformal filter integrated into the antenna structure.



Fig. 89 The simulated VSWR of the stripline spiral antenna integrated with spiral filter.

8.3.3 Conformal Mapping Applications

The conformal mapping analysis can also be applied on several problems. The first one is microstrip coupled capacitance. The second one is to discuss the effects of guarding traces on parallel microstrips for package and PCB design. We can provide the minimum width and spacing of guarding traces analytically for package and PCB design instead of empirical design rules. The third one is to evaluate the capacitance of interdigital capacitor.

REFERENCES

- [1] M. A. Peyrot-Solis, G. M. Galvan-Tejada, and H. Jardon-Aguilar, "State of the art in ultra-wideband antennas," in *Int. Conf. Elect. Electron. Eng.*, 2005, pp. 101-105.
- [2] Z. N. Chen, "UWB antennas with enhanced performances (invited)," in *Int. Conf. Microw. Millimeter Wave Tech., ICMMT* 2008, pp. 387-390.
- [3] J. W. Wu, H. M. Hsiao, J. H. Lu, and S. H. Chang, "Dual broadband design of rectangular slot antenna for 2.4 and 5 GHz wireless communication," *Electron. Lett.*, vol. 40, pp. 1461-1463, Nov. 2004.
- [4] Y. Kim, S. Keely, J. Ghosh, and H. Ling, "Application of artificial neural networks to broadband antenna design based on a parametric frequency model," *IEEE Trans. Antennas Propag.*, vol. 55, pp. 669-674, Mar. 2007.
- [5] H. Yi, K. Chan, and B. Cheeseman, "Review of broadband antenna measurements," in *Europ. Conf. Antennas Propag., EuCAP*, 2006, pp. 1-4.
- [6] P. E. Mayes, "Frequency-independent antennas and broad-band derivatives thereof," *Proc. IEEE*, vol. 80, pp. 103-112, Jan. 1992.
- [7] I. Hertl and M. Strycek, "UWB antennas for ground penetrating radar application," in *19th Int. Conf. Applied Electromagn. Commun., ICECom*, 2007, pp. 1-4.
- [8] P. V. Nikitin and K. V. S. Rao, "Helical antenna for handheld UHF RFID reader," in *IEEE Int. Conf. RFID*, 2010, pp. 166-173.
- [9] X. L. Bao, M. J. Ammann, and P. McEvoy, "Microstrip-fed wideband circularly polarized printed antenna," *IEEE Trans. Antennas Propag.*, vol. 58, pp. 3150-3156, Oct. 2010.
- [10] Y.-X. Guo, K.-W. Khoo, and L. C. Ong, "Wideband circularly polarized patch antenna using broadband baluns," *IEEE Trans. Antennas Propag.*, vol. 56, pp. 319-326, Feb. 2008.
- [11] C. Lin, F.-S. Zhang, Y.-C. Jiao, F. Zhang, and X. Xue, "A three-fed microstrip antenna for wideband circular polarization," *IEEE Antennas Wireless Propag. Lett.*, vol. 9, pp. 359-362, Apr. 2010.
- [12] R. Bawer and J. Wolfe, "The spiral antenna," in *IRE Int. Conv. Rec.*, 1960, pp. 84-95.

- [13] V. Rumsey, "Frequency independent antennas," in *IRE Int. Conv. Rec.*, 1957, pp. 114-118.
- [14] J. D. Dyson, R. Bawer, P. E. Mayes, and J. I. Wolfe, "A note on the difference between equiangular and Archimedes spiral antennas (correspondence)," *IRE Trans. Microw. Theory Tech.*, vol. 9, pp. 203-205, Mar. 1961.
- [15] P. Lacko, "Archimedean-spiral and log-spiral antenna comparison," *Proc. SPIE*, vol. 4742, p. 230, 2002.
- [16] D. Paolino, "Reduced-size spiral antenna design using dielectric overlay loading for use in ground penetrating radar and design of alternative antennas using Vivaldi radiators," *Proc. SPIE*, vol. 4742, p. 218, 2002.
- [17] V. Rumsey, "A solution to the equiangular spiral antenna problem," *IRE Trans. Antennas Propag.*, vol. 7, pp. 117-117, Dec. 1959.
- [18] J. Dyson, "The equiangular spiral antenna," *IRE Trans. Antennas Propag.*, vol. 7, pp. 181-187, Apr. 1959.
- [19] J. Dyson, "The unidirectional equiangular spiral antenna," *IRE Trans. Antennas Propag.*, vol. 7, pp. 329-334, Oct. 1959.
- [20] B. Cheo, V. Rumsey, and W. Welch, "A solution to the frequency-independent antenna problem," *IRE Trans. Antennas Propag.*, vol. 9, pp. 527-534, Nov. 1961.
- [21] R. Sivan-Sussman, "Various modes of the equiangular spiral antenna," *IEEE Trans. Antennas Propag.*, vol. 11, pp. 533-539, Sep. 1963.
- [22] J. J. H. Wang and V. K. Tripp, "Design of multioctave spiral-mode microstrip antennas," *IEEE Trans. Antennas Propag.*, vol. 39, pp. 332-335, Mar 1991.
- [23] S. M. Wentworth and S. M. Rao, "Analysis of equiangular spiral antennas," *Int. J. Microw. Millimeter-Wave Comput.-Aided Eng.*, vol. 6, pp. 92-99, 1996.
- [24] Z. Lou and J.-M. Jin, "Modeling and simulation of broad-band antennas using the time-domain finite element method," *IEEE Trans. Antennas Propag.*, vol. 53, pp. 4099-4110, Dec. 2005.
- [25] M. McFadden and W. R. Scott, "Analysis of the equiangular spiral antenna on a dielectric substrate," *IEEE Trans. Antennas Propag.*, vol. 55, pp. 3163-3171, Nov. 2007.
- [26] H. Nakano, K. Kikkawa, Y. Iitsuka, and J. Yamauchi, "Equiangular spiral antenna backed by a shallow cavity with absorbing strips," *IEEE Trans. Antennas Propag.*, vol. 56, pp. 2742-2747, Aug. 2008.

- [27] H. Nakano, K. Kikkawa, N. Kondo, Y. Iitsuka, and J. Yamauchi, "Low-profile equiangular spiral antenna backed by an EBG reflector," *IEEE Trans. Antennas Propag.*, vol. 57, pp. 1309-1318, May 2009.
- [28] S.-G. Mao, J.-C. Yeh, and S.-L. Chen, "Ultrawideband circularly polarized spiral antenna using integrated balun with application to time-domain target detection," *IEEE Trans. Antennas Propag.*, vol. 57, pp. 1914-1920, July 2009.
- [29] T. W. Eubanks and K. Chang, "A compact parallel-plane perpendicular-current feed for a modified equiangular spiral antenna," *IEEE Trans. Antennas Propag.*, vol. 58, pp. 2193-2202, July 2010.
- [30] E. M. Turner, "Spiral slot antenna," United States Patent, 1958.
- [31] W. Curtis, "Spiral antennas," *IRE Trans. Antennas Propag.*, vol. 8, pp. 298-306, May 1960.
- [32] J. Kaiser, "The Archimedean two-wire spiral antenna," *IRE Trans. Antennas Propag.*, vol. 8, pp. 312-323, May 1960.
- [33] C. Fumeaux, D. Baumann, and R. Vahldieck, "Finite-volume time-domain analysis of a cavity-backed Archimedean spiral antenna," *IEEE Trans. Antennas Propag.*, vol. 54, pp. 844-851, Mar. 2006.
- [34] H. Nakano, K. Nogami, S. Arai, H. Mimaki, and J. Yamauchi, "A spiral antenna backed by a conducting plane reflector," *IEEE Trans. Antennas Propag.*, vol. 34, pp. 791-796, Jun. 1986.
- [35] N. J. Champagne, II, J. T. Williams, R. M. Sharpe, S. U. Hwu, and D. R. Wilton, "Numerical modeling of impedance loaded multi-arm Archimedean spiral antennas," *IEEE Trans. Antennas Propag.*, vol. 40, pp. 102-108, Jan. 1992.
- [36] H. Nakano, Y. Shinma, and J. Yamauchi, "A monofilar spiral antenna and its array above a ground plane-formation of a circularly polarized tilted fan beam," *IEEE Trans. Antennas Propag.*, vol. 45, pp. 1506-1511, Oct. 1997.
- [37] H. Nakano, K. Hirose, I. Ohshima, and J. Yamauchi, "An integral equation and its application to spiral antennas on semi-infinite dielectric materials," *IEEE Trans. Antennas Propag.*, vol. 46, pp. 267-274, Feb. 1998.
- [38] R.-L. Li and H. Nakano, "Numerical analysis of arbitrarily shaped probe-excited single-arm printed wire antennas," *IEEE Trans. Antennas Propag.*, vol. 46, pp. 1307-1317, Sep. 1998.

- [40] H. Nakano, M. Ikeda, K. Hitosugi, and J. Yamauchi, "A spiral antenna sandwiched by dielectric layers," *IEEE Trans. Antennas Propag.*, vol. 52, pp. 1417-1423, Jun. 2004.
- [41] H. Nakano, R. Satake, and J. Yamauchi, "Extremely low-profile, single-arm, wideband spiral antenna radiating a circularly polarized wave," *IEEE Trans. Antennas Propag.*, vol. 58, pp. 1511-1520, May 2010.
- [42] H. Nakano, T. Igarashi, H. Oyanagi, Y. Iitsuka, and J. Yamauchi, "Unbalancedmode spiral antenna backed by an extremely shallow cavity," *IEEE Trans. Antennas Propag.*, vol. 57, pp. 1625-1633, Jun. 2009.
- [43] H. Nakano, *Helical and Spiral Antennas : A Numerical Approach*. New York: Wiley, 1987.
- [44] C. W. Penney and R. J. Luebbers, "Input impedance, radiation pattern, and radar cross section of spiral antennas using FDTD," *IEEE Trans. Antennas Propag.*, vol. 42, pp. 1328-1332, Sep. 1994.
- [45] R. Li and G. Ni, "Numerical analysis of 4-arm Archimedian printed spiral antenna," *IEEE Trans. Magn.*, vol. 33, pp. 1512-1515, Mar. 1997.
- [46] D. S. Filipovic, A. U. Bhobe, and T. P. Cencich, "Low-profile broadband dualmode four-arm slot spiral antenna with dual Dyson balun feed," *IEE Proc. Microw. Antennas Propag.*, pp. 527-533, Dec. 2005.
- [47] T.-K. Chen and G. H. Huff, "Stripline-fed Archimedean spiral antenna," *IEEE Antennas Wireless Propag. Lett.*, vol. 10, pp. 346-349, May 2011.
- [48] D. S. Filipovic and J. L. Volakis, "A flush-mounted multifunctional slot aperture (combo-antenna) for automotive applications," *IEEE Trans. Antennas Propag.*, vol. 52, pp. 563-571, Feb. 2004.
- [49] M. N. Afsar, W. Yong, and R. Cheung, "Analysis and measurement of a broadband spiral antenna," *IEEE Antennas Propag. Mag.*, vol. 46, pp. 59-64, Feb. 2004.
- [50] E. Gschwendtner and W. Wiesbeck, "Ultra-broadband car antennas for communications and navigation applications," *IEEE Trans. Antennas Propag.*, vol. 51, pp. 2020-2027, Aug. 2003.

- [52] T. Ozdemir, J. L. Volakis, and M. W. Nurnberger, "Analysis of thin multioctave cavity-backed slot spiral antennas," *IEE Proc. Microw. Antennas Propag.*, vol. 146, pp. 447-454, 1999.
- [53] J. M. Bell and M. F. Iskander, "A low-profile Archimedean spiral antenna using an EBG ground plane," *IEEE Antennas Wireless Propag. Lett.*, vol. 3, pp. 223-226, Dec. 2004.
- [54] L. Schreider, X. Begaud, M. Soiron, B. Perpere, and C. Renard, "Broadband Archimedean spiral antenna above a loaded electromagnetic band gap substrate," *IET Microw. Antennas Propag.*, vol. 1, pp. 212-216, Feb. 2007.
- [55] B. A. Kramer, M. Lee, C.-C. Chen, and J. L. Volakis, "Design and performance of an ultrawide-band ceramic-loaded slot spiral," *IEEE Trans. Antennas Propag.*, vol. 53, pp. 2193-2199, July 2005.
- [56] B. A. Kramer, S. Koulouridis, C. C. Chen, and J. L. Volakis, "A novel reflective surface for an UHF spiral antenna," *IEEE Antennas Wireless Propag. Lett.*, vol. 5, pp. 32-34, Dec. 2006.
- [57] M. Lee, B. A. Kramer, C.-C. Chen, and J. L. Volakis, "Distributed lumped loads and lossy transmission line model for wideband spiral antenna miniaturization and characterization," *IEEE Trans. Antennas Propag.*, vol. 55, pp. 2671-2678, Oct. 2007.
- [58] P. C. Werntz and W. L. Stutzman, "Design, analysis and construction of an Archimedean spiral antenna and feed structure," in *IEEE Southeastcon* '89. *Proc. Energy Informat. Technol.*, 1989, pp. 308-313 vol.1.
- [59] K. Louertani, R. Guinvarc'h, N. Ribiere-Tharaud, and M. Darces, "Design of a spiral antenna with coplanar feeding solution," in *Proc. IEEE AP-S Int. Symp.*, 2008, pp. 1-4.
- [60] J. A. Kasemodel, C.-C. Chen, I. J. Gupta, and J. L. Volakis, "Miniature continuous coverage wideband GPS antenna array," in *Proc. IEEE AP-S Int. Symp.*, 2008, pp. 1-4.
- [61] G. H. Huff and T. L. Roach, "Stripline-based spiral antennas with integrated feed structure, impedance transformer, and dyson-style balun," in *Proc. IEEE AP-S Int. Symp.*, 2007, pp. 2698-2701.

- [63] R. W. Jackson, "Considerations in the use of coplanar waveguide for millimeterwave integrated circuits," *IEEE Trans. Microw. Theory Tech.*, vol. 34, pp. 1450-1456, Dec. 1986.
- [64] G. Ghione and C. U. Naldi, "Coplanar waveguides for MMIC applications: effect of upper shielding, conductor backing, finite-extent ground planes, and line-toline coupling," *IEEE Trans. Microw. Theory Tech.*, vol. 35, pp. 260-267, Mar. 1987.
- [65] S. S. Bedair and I. Wolff, "Fast, accurate and simple approximate analytic formulas for calculating the parameters of supported coplanar waveguides for (M)MIC's," *IEEE Trans. Microw. Theory Tech.*, vol. 40, pp. 41-48, Jan. 1992.
- [66] W. Deal, "Coplanar waveguide basics for MMIC and PCB design," IEEE Microw. Mag., vol. 9, pp. 120-133, Aug. 2008.
- [67] C. P. Wen, "Coplanar waveguide: a surface strip transmission line suitable for nonreciprocal gyromagnetic device applications," *IEEE Trans. Microw. Theory Tech.*, vol. 17, pp. 1087-1090, May 1969.
- [68] O. Ramer, "Integrated optic electrooptic modulator electrode analysis," *IEEE J. Quantum Electron.*, vol. 18, pp. 386-392, Mar. 1982.
- [69] J. Kessler and R. Dill, "Influence of buffer layers within YBa/sub 2/Cu/sub 3/O/sub 7-x/ coplanar waveguide structure," *IEEE Microw. Guided Wave Lett.*, vol. 2, pp. 6-7, Jan. 1992.
- [70] N. Newman and W. G. Lyons, "High-temperature superconducting microwave devices: Fundamental issues in materials, physics, and engineering," J. Supercond., vol. 6, pp. 119-160, 1993.
- [71] C. Wang, C. Lai, and T. Ma, "Novel uniplanar synthesized coplanar waveguide and the application to miniaturized rat-race coupler," in *IEEE MTT-S Int. Microw. Symp. Dig.*, 2010, pp. 1-1.
- [72] K. K. M. Cheng and J. K. A. Everard, "A new technique for the quasi-TEM analysis of conductor-backed coplanar waveguide structures," *IEEE Trans. Microw. Theory Tech.*, vol. 41, pp. 1589-1592, 1993.
- [73] G. Ghione and C. Naldi, "Parameters of coplanar waveguides with lower ground plane," *Electron. Lett.*, vol. 19, pp. 734-735, Sep. 1983.

- [75] S. Gevorgian, L. J. P. Linner, and E. L. Kollberg, "CAD models for shielded multilayered CPW," *IEEE Trans. Microw. Theory Tech.*, vol. 43, pp. 772-779, Apr. 1995.
- [76] S. S. Gevorgian, T. Martinsson, P. L. J. Linner, and E. L. Kollberg, "CAD models for multilayered substrate interdigital capacitors," *IEEE Trans. Microw. Theory Tech.*, vol. 44, pp. 896-904, Jun 1996.
- [77] N. H. Zhu, E. Y. B. Pun, and P. S. Chung, "Field distributions in supported coplanar lines using conformal mapping techniques," *IEEE Trans. Microw. Theory Tech.*, vol. 44, pp. 1493-1496, Aug. 1996.
- [78] E. Carlsson and S. Gevorgian, "Conformal mapping of the field and charge distributions in multilayered substrate CPWs," *IEEE Trans. Microw. Theory Tech.*, vol. 47, pp. 1544-1552, Aug. 1999.
- [79] J.-W. Lee, I.-P. Hong, T.-H. Yoo, and H.-K. Park, "Quasi-static analysis of conductor-backed coupled coplanar waveguide," *Electron. Lett.*, vol. 34, pp. 1861-1862, Sep. 1998.
- [80] C. Veyres and V. Fouad Hanna, "Extension of the application of conformal mapping techniques to coplanar lines with finite dimensions," *Int. J. Electron.*, vol. 48, p. 47, 1980.
- [81] S. Visan, O. Picon, and V. F. Hanna, "3D characterization of air bridges and via holes in conductor-backed coplanar waveguides for MMIC applications," in *IEEE MTT-S Int. Microw. Symp. Dig.*, 1993, pp. 709-712 vol.2.
- [82] Z. Jinchang, T. Yu, C. Yan, G. Bo, and T. Ling, "The analysis of electromagnetic characteristics of coplanar waveguide in different models using Finite-Difference Time-Domain method," in *Int. Conf. Microw. Millimeter Wave Technol.*, *ICMMT*, 2008, pp. 730-733.
- [83] Z. Ma and E. Yamashita, "Comparative studies of discontinuities in single and double layered conductor-backed coplanar waveguides," in *IEEE MTT-S Int. Microw. Symp. Dig.*, 1996, pp. 1803-1806 vol.3.
- [84] D. A. Thompson and R. L. Rogers, "The interdigital coplanar waveguide: a new low-impedance micromachinable planar structure," *IEEE Microw. Guided Wave Lett.*, vol. 8, pp. 257-259, Jul. 1998.

- [86] J. B. Knorr and K. Kuchler, "Analysis of coupled slots and coplanar strips on dielectric substrate," *IEEE Trans. Microw. Theory Tech.*, vol. 23, pp. 541-548, Jul. 1975.
- [87] E. Yamashita and K. Atsuki, "Analysis of microstrip-like transmission lines by nonuniform discretization of integral equations," *IEEE Trans. Microw. Theory Tech.*, vol. 24, pp. 195-200, Apr. 1976.
- [88] T. Hatsuda, "Computation of coplanar-type strip-line characteristics by relaxation method and its application to microwave circuits," *IEEE Trans. Microw. Theory Tech.*, vol. 23, pp. 795-802, Oct. 1975.
- [89] C. P. Wen, "Coplanar-waveguide directional couplers," *IEEE Trans. Microw. Theory Tech.*, vol. 18, pp. 318-322, Jun. 1970.
- [90] M. E. Davis, E. W. Williams, and A. C. Celestini, "Finite-boundary corrections to the coplanar waveguide analysis (short papers)," *IEEE Trans. Microw. Theory Tech.*, vol. 21, pp. 594-596, Sep. 1973.
- [91] V. Fouad hanna, "Finite boundary corrections to coplanar stripline analysis," *Electron. Lett.*, vol. 16, pp. 604-606, July 1980.
- [92] M. Gillick, I. D. Robertson, and J. S. Joshi, "Direct analytical solution for the electric field distribution at the conductor surfaces of coplanar waveguides," *IEEE Trans. Microw. Theory Tech.*, vol. 41, pp. 129-135, Jan. 1993.
- [93] S. S. Bedair and I. Wolff, "Fast and accurate analytic formulas for calculating the parameters of a general broadside-coupled coplanar waveguide for (M)MIC applications," *IEEE Trans. Microw. Theory Tech.*, vol. 37, pp. 843-850, May 1989.
- [94] K. K. M. Cheng, "Analysis and synthesis of coplanar coupled lines on substrates of finite thicknesses," *IEEE Trans. Microw. Theory Tech.*, vol. 44, pp. 636-639, Apr. 1996.
- [95] G. Ghionc and M. Goano, "A closed-form CAD-oriented model for the high-frequency conductor attenuation of symmetrical coupled coplanar waveguides," *IEEE Trans. Microw. Theory Tech.*, vol. 45, pp. 1065-1070, Jul 1997.

- [96] C. H. Wu and S. Uysal, "A new systematic and efficient method of analysis for conductor-backed coplanar-waveguide directional couplers," *IEEE Trans. Microw. Theory Tech.*, vol. 47, pp. 1127-1131, Jul. 1999.
- [97] K. K. M. Cheng, "Effect of conductor backing on the line-to-line coupling between parallel coplanar lines," *IEEE Trans. Microw. Theory Tech.*, vol. 45, pp. 1132-1134, Jul 1997.
- [98] K. K. M. Cheng, "Characteristic parameters of symmetrical triple coupled CPW lines," *Electron. Lett.*, vol. 33, pp. 685-687, Apr 1997.
- [99] Z. Xilang, W. Yu, and W. Shu, "Concise formulas of quasistatic parameter for parallel coupling coplanar lines," *Electron. Lett.*, vol. 34, pp. 1671-1672, Aug 1998.
- [100] V. Fouad Hanna and D. Thebault, "Analysis of asymmetrical coplanar waveguides," *Int. J. Electron.*, vol. 50, p. 221, 1981.
- [101] V. F. Hanna and D. Thebault, "Theoretical and experimental investigation of asymmetric coplanar waveguides (short papers)," *IEEE Trans. Microw. Theory Tech.*, vol. 32, pp. 1649-1651, Dec 1984.
- [102] S. S. Gevorgian and I. G. Mironenko, "Asymmetric coplanar-strip transmission lines for MMIC and integrated optic applications," *Electron. Lett.*, vol. 26, pp. 1916-1918, Oct. 1990.
- [103] F. Shao-Jun and W. Bai-Suo, "Analysis of asymmetric coplanar waveguide with conductor backing," *IEEE Trans. Microw. Theory Tech.*, vol. 47, pp. 238-240, Feb 1999.
- [104] C. Erli and S. Y. Chou, "Characteristics of coplanar transmission lines on multilayer substrates: modeling and experiments," *IEEE Trans. Microw. Theory Tech.*, vol. 45, pp. 939-945, Jun. 1997.
- [105] G. Ghione and M. Goano, "Revisiting the partial-capacitance approach to the analysis of coplanar transmission lines on multilayered substrates," *IEEE Trans. Microw. Theory Tech.*, vol. 51, pp. 2007-2014, Sept. 2003.
- [106] J. Svacina, "A simple quasi-static determination of basic parameters of multilayer microstrip and coplanar waveguide," *IEEE Microw. Guided Wave Lett.*, vol. 2, pp. 385-387, Oct 1992.
- [107] Z. Ning Hua, E. Yue Bun Pun, and L. Jia Xu, "Analytical formulas for calculating the effective dielectric constants of coplanar lines for OIC applications," *Microw. Optical Technol. Lett.*, vol. 9, pp. 229-232, 1995.

- [108] A. N. Sytchev and A. Y. Shevtsov, "Design of slablines on multilayer dielectrics," in *Int. Conf. Microw. Millimeter Wave Technol. Proc.*, *ICMMT*, 1998, pp. 581-586.
- [109] M. Y. Frankel, S. Gupta, J. A. Valdmanis, and G. A. Mourou, "Terahertz attenuation and dispersion characteristics of coplanar transmission lines," *IEEE Trans. Microw. Theory Tech.*, vol. 39, pp. 910-916, Jun 1991.
- [110] W. Heinrich, "Quasi-TEM description of MMIC coplanar lines including conductor-loss effects," *IEEE Trans. Microw. Theory Tech.*, vol. 41, pp. 45-52, Jan 1993.
- [111] G. Ghione, "A CAD-oriented analytical model for the losses of general asymmetric coplanar lines in hybrid and monolithic MICs," *IEEE Trans. Microw. Theory Tech.*, vol. 41, pp. 1499-1510, Sep 1993.
- [112] C. Liu, Y. Lu, C. Du, J. Cui, and X. Shen, "The broadband spiral antenna design based on hybrid backed-cavity," *IEEE Trans. Antennas Propag.*, vol. 58, pp. 1876-1882, June 2010.
- [113] M. Kim, H. Choo, and I. Park, "Two-arm microstrip spiral antenna for multibeam pattern control," *Electron. Lett.*, vol. 41, pp. 627-629, May 2005.
- [114] Y. Mushiake, "Self-complementary antennas," *IEEE Antennas Propag. Mag.*, vol. 34, pp. 23-29, Dec. 1992.
- [115] H. G. Booker, "Slot aerials and their relation to complementary wire aerials (Babinet's principle)," J. Inst. Elect. Eng. - Part IIIA: Radiolocation, vol. 93, pp. 620-626, 1946.
- [116] Y. Mushiake, "A report on Japanese development of antennas: from the Yagi-Uda antenna to self-complementary antennas," *IEEE Antennas Propag. Mag.*, vol. 46, pp. 47-60, 2004.
- [117] W. L. Stutzman and G. A. Thiele, *Antenna Theory and Design*, 2nd ed.: Wiley, 1997.
- [118] C. A. Balanis, *Antenna Theory: Analysis and Design*, 3rd ed. Hoboken, New Jersey: Wiley-Interscience, 2005.
- [119] R. E. Collin, *Foundations for Microwave Engineering*, 2 ed. New York: Wiley-IEEE Press, 2000.
- [120] T. A. Driscoll and L. N. Trefethen, *Schwarz–Christoffel Mapping*. Cambridge, U.K.: Cambridge Univ. Press, 2002.

- [121] I. S. Gradshten and . . . yzhik, *Table of Integrals, Series and Products*. Oxford, UK: Academic Press, 2007.
- [122] Z. Nehari, *Conformal Mapping*. New York: McGraw-Hill, 1952.
- [123] *HFSS*, v12.0 ed.: Ansoft Corporation, Pittsburgh, PA, 2010.
- [124] J. A. Huffman and T. Cencich, "Modal impedances of planar, noncomplementary, N-fold symmetric antenna structures," *IEEE Antennas Propag. Mag.*, vol. 47, pp. 110-116, Feb. 2005.
- [125] R. Bawer and J. J. Wolfe, "A printed circuit balun for use with spiral antennas," *IRE Trans. Microw. Theory Tech.*, vol. 8, pp. 319-325, May 1960.
- [126] R. E. Collin, "The optimum tapered transmission line matching section," *IRE Proc.*, vol. 44, pp. 539-548, 1956.
- [127] Sylgard 184: Dow Corning Corporation, Midland, MD.
- [128] R. G. Corzine and J. A. Mosko, *Four-Arm Spiral Antennas*. Norwood, MA: Artech House, 1990.
- [129] R. T. Gloutak, Jr. and N. G. Alexopoulous, "Two-arm eccentric spiral antenna," *IEEE Trans. Antennas Propag.*, vol. 45, pp. 723-730, Apr. 1997.
- [130] S. C. Wu, "Analysis and design of conductor-backed square Archimedean spiral antennas," *Electromagn.*, vol. 14, pp. 305 318, 1994.
- [131] T. Ozdemir, J. L. Volakis, and M. W. Nurnberger, "Analysis of thin multioctave cavity-backed slot spiral antennas," *Proc. IEE Microw. Antennas Propag.*, vol. 146, pp. 447-454, Dec. 1999.
- [132] H. Nakano, K. Hitosugi, N. Tatsuzawa, D. Togashi, H. Mimaki, and J. Yamauchi, "Effects on the radiation characteristics of using a corrugated reflector with a helical antenna and an electromagnetic band-gap reflector with a spiral antenna," *IEEE Trans. Antennas Propag.*, vol. 53, pp. 191-199, Jan. 2005.
- [133] K. Chang, I. Bahl, and V. Nair, *RF and Microwave Circuit and Component Design for Wireless Systems*. New York: John Wiley & Sons, 2002.
- [134] T.-K. Chen and G. H. Huff, "On the input impedance of two-arm gap-fed Archimedean spiral antennas in free space," *IEEE Trans. Antennas Propag.*, submitted for publication.

- [135] C. Igwe, "Axial ratio of an antenna illuminated by an imperfectly circularly polarized source," *IEEE Trans. Antennas Propag.*, vol. 35, pp. 339-342, Mar. 1987.
- [136] Y. Fukuoka and T. Itoh, "Slow-wave propagation on MIS periodic coplanar waveguide," *Electron. Lett.*, vol. 19, pp. 37-38, Jan. 1983.
- [137] D. Kaiser, M. Block, U. Lackmann, and D. Jager, "Variable phase shift of spatially periodic proton-bombarded Schottky coplanar lines," *Electron. Lett.*, vol. 25, pp. 1135-1136, Aug. 1989.
- [138] J. Sor, Y. Qian, and T. Itoh, "Miniature low-loss CPW periodic structures for filter applications," *IEEE Trans. Microw. Theory Tech.*, vol. 49, pp. 2336-2341, Dec. 2001.
- [139] Z. Lei, "Guided-wave characteristics of periodic coplanar waveguides with inductive loading unit-length transmission parameters," *IEEE Trans. Microw. Theory Tech.*, vol. 51, pp. 2133-2138, Oct. 2003.
- [140] C.-N. Chang, Y.-C. Wong, and C. H. Chen, "Full-wave analysis of coplanar waveguides by variational conformal mapping technique," *IEEE Trans. Microw. Theory Tech.*, vol. 38, pp. 1339-1344, Sep. 1990.
- [141] R. H. Jansen, "Hybrid mode analysis of end effects of planar microwave and millimetrewave transmission lines," *IEE Proc. H Microw. Optics and Antennas*, vol. 128, pp. 77-86, 1981.
- [142] COMSOL Multiphysics, v4.2a ed., 2011.
- [143] *RT/Duroid*[®] *laminates*: Rogers Corporation, Rogers, CT.

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