Studies on planar helical slow-wave structures for traveling-wave tube applications

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**STUDIES ON PLANAR HELICAL SLOW-WAVE STRUCTURES
FOR TRAVELING-WAVE TUBE APPLICATIONS**

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ABSTRACT

This thesis presents extensive studies on a new planar helix slow-wave structure (SWS) which consists of a planar helix with straight-edge connections (PH-SEC). The PH-SEC has been studied in the context of traveling-wave tube (TWT) applications. Unlike the conventional circular helix, the PH-SEC can be fabricated easily using printed circuit techniques or microfabrication techniques. In addition, by changing the aspect ratio, the PH-SEC can become suitable for interaction with a sheet electron beam which can offer many advantages for high frequency TWTs. To avoid oscillations at high power levels, the PH-SEC has also been modified into a rectangular ring-bar slow-wave structure with straight-edge connections (RRB-SEC); this modification is similar to that of the circular helix into the circular ring-bar SWS.

For the PH-SEC, dispersion characteristics for several practical modifications to the basic structure have been examined. These modifications comprise a vacuum tunnel, metal shield and multilayer dielectric substrates. A modified effective dielectric constant (MEDC) method has been proposed to obtain the dispersion characteristics for the different possible configurations. Further, coupling impedance for the different configurations has been calculated using the corresponding two-dimensional approximations. Effects of variations in the aspect ratio, metal shield distance and dielectric constant of the substrates on the phase velocity and the coupling impedance have been studied.

The PH-SEC structure incorporating coplanar waveguide (CPW) feed has been designed and fabricated for printed circuit fabrication and microfabrication. Effects of dimensional parameters have been studied. Several PH-SECs with band edge frequency less than 10 GHz have been fabricated using printed circuit techniques. The measured results for these structures validate the analytical results obtained using the MEDC method. A microfabrication process, involving several UV-LIGA steps, has been proposed and demonstrated to produce high-aspect-ratio PH-SEC structures at W-band (75 - 110 GHz). On-wafer measurements have been carried out on a number of microfabricated SWSs. The cold-test parameters

(dispersion characteristics and coupling impedance) of the SWSs have also been obtained using simulations, and the effects of fabrication, such as surface roughness, have been accounted for by estimating effective conductivity of different parts of the microfabricated structures.

It is shown that, compared to the PH-SEC, its modification, the RRB-SEC enhances the coupling impedance for the fundamental forward-wave while reducing the coupling impedance for the backward-wave. Detailed results for the phase velocity and the coupling impedance of the RRB-SEC have been presented to show the effects of structure dimensions. The RRB-SEC incorporating the CPW feed has also been designed and fabricated for W-band using the microfabrication process developed for the PH-SEC.

A low power square aspect ratio PH-SEC incorporating a sever has been simulated with a circular cross-section electron beam at W-band using 3D CST Particle Studio. A simplified coupler, similar to the CPW, has been used in the Particle-In-Cell (PIC) solver. The linear and non-linear amplification of the input signal has been examined. The input and output couplers based on W-band (WR-10) rectangular waveguide have been designed for high power application. It is shown that the structure with the rectangular waveguide couplers leads to a ‘cleaner’ output compared to that with the CPW feed. A square aspect ratio RRB-SEC has also been designed with rectangular waveguide couplers.

The studies reported in this thesis have potential applications in printed-circuit TWTs at low frequencies ($<18\text{GHz}$) and microfabricated TWTs at millimeter wave frequencies (30 - 300 GHz). The proposed microfabrication process can also be scaled to fabricate planar helical SWSs at terahertz frequencies (0.3 - 3 THz).

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LIST OF ACRONYMS AND SYMBOLS

ACRONYM	FULL EXPRESSION
2D	Two dimensional
3D	Three dimensional
AFM	Atomic force microscope
APBN	Anisotropic pyrolytic boron nitride
BWO	Backward wave oscillator
CAD	Computer aided design
CPW	Coplanar waveguide
CST	Computer simulation technology
CVD	Chemical vapor deposition
DC	Direct current
DFT	Discrete Fourier transform
DRIE	Deep reactive ion etching
ECM	Electronic countermeasure
EDC	Effective dielectric constant
EDM	Electrical discharge machining
FDTD	Finite difference time domain
FEM	Finite element method
FIT	Finite integral technique
HFSS	High frequency structure simulator
LIGA	Lithography, electroplating, and micro-molding
MCPA	Multichannel power amplifier
MEDC	Modified effective dielectric constant
MEMS	Micromachined electromechanical systems
MIC	Microwave Integrated Circuit
MMIC	Monolithic Microwave Integrated Circuit
MPM	Microwave power module
MWS	Microwave studio
PBA	Perfect boundary approximation
PCB	Printed circuit board
PEC	Perfect electric conductor
PH-SEC	Planar helix with straight-edge connections
PIC	Particle-in-cell
PPM	Periodic-permanent-magnet
PTH	Plated-through-hole
RF	Radiofrequency
RIE	Reactive ion etching

RMS	Root-mean-square
RRB-SEC	Rectangular ring-bar with straight-edge connections
SEM	Scanning electron microscope
SWS	Slow-wave structure
TE	Transverse electric
TLM	Transmission-line-matrix
TM	Transverse magnetic
TST	Thin sheet technique
TWT	Traveling-wave tube
UAV	Unmanned aerial vehicle
UC	Unidirectionally conducting
UV	Ultraviolet
VED	Vacuum electron device

SYMBOL	MEANING
β	Phase constant
β_0	Phase constant of the fundamental mode
β_n	Phase constant of the n^{th} space harmonic
ε	Permittivity of a medium
ε_0	Permittivity of free-space
ε_r	Relative permittivity of a medium
ε_{eff}	Effective dielectric constant
γ	Radial decay coefficient
λ_g	Guide wavelength
μ	Permeability of a medium
μ_0	Permeability of free space
μ_r	Relative permeability of a medium
σ	Conductivity
δ_s	Skin depth
ω	Angular frequency
η_{cir}	Circuit efficiency
η_e	Electronic efficiency
Ψ	Pitch angle
Ψ_{eff}	Effective pitch angle for PH-SEC
c	Speed of light in free space
$E_z \{0\}$	On-axis longitudinal electric field intensity
$E_{z,\text{av}}$	Average longitudinal electric field intensity
f	Frequency
f_{bump}	Fraction of the surface area covered by the rough surface
I_0	Zeroth-order modified Bessel function of the first kind

I_1	First-order modified Bessel function of the first kind
I_{beam}	Electron beam current
k	Wave number in a medium
k_0	Wave number in free space
k_n	Wave number in region n
k_{xn}, u, v, w	Transverse decay coefficients
K_0	Zeroth-order modified Bessel function of the second kind
K_1	First-order modified Bessel function of the second kind
K_c	Coupling impedance (or interaction impedance)
n	Integer index
P	Power propagating through the structure
$P_{a,rough}$	Power absorption due to rough surface
$P_{a,smooth}$	Power absorption due to smooth surface
P_c	Cross section perimeter
P_{in}	Radiofrequency input power
P_{out}	Radiofrequency output power
r, θ, z	Circular cylindrical coordinates
$R_{max, TE}$	Resistivity for protrusions due to foreign objects
S_{11}	Reflection coefficient
S_{21}	Transmission coefficient
v_p	Phase velocity
v_g	Group velocity
V_z	Longitudinal voltage
V_{beam}	Electron beam voltage
W	Energy stored per unit axial length
x, y, z	Cartesian coordinates
Z_0	Characteristic impedance
Au	Gold
BeO	Beryllium oxide
Cu	Copper
GaN	Gallium nitride
Si	Silicon
SiC	Silicon carbide
SiO_2	Silicon dioxide
Ti	Titanium

CHAPTER 1

INTRODUCTION

1.1 Background

Microwave (0.3 - 30GHz) and millimeter-wave (30 - 300GHz) high power technologies are crucial for modern society. These technologies have applications in civilian (e.g., cellular communications, satellite communications, broadcast media transmission and domestic microwave oven), military (e.g., radar systems, electronic countermeasures (ECM) and high-power microwave weapons), scientific (e.g., radio astronomy, spectroscopy and deep space communications) as well as industrial (e.g., testing instruments, material processing and industrial plasmas) arenas [1]. The power levels for the above mentioned applications range from watts to megawatts. For such power levels, vacuum electron devices (VEDs), including traveling-wave tubes (TWTs), klystrons, magnetrons, etc, remain the preferred technology for high frequency applications even today.

In general, vacuum electronics and solid-state sources have complementary roles in modern systems. However, for high power and high frequency applications, solid-state sources encounter significant limitations. All linear beam microwave sources operate on the principle of converting the kinetic energy of an electron stream to microwave power. In solid-state devices, the electron stream moves in a solid semiconducting medium. This reduces the mobility of the electrons, restricting the operation to moderate frequencies. Further, the collisions of the electron stream inside the solid medium convert a significant portion of its kinetic energy into heat; this reduces the efficiency of operation. In addition, the maximum operating temperature and the maximum electric field inside a semiconducting medium are also restricted. On the other hand, in VEDs, the electron stream moves collisionlessly through an evacuated region; the spent electron beam can be collected in a suitable collector. Thus, in the context of high power and high frequency sources, vacuum electronics can offer significant advantages compared to

solid-state electronics. For example, communication satellites routinely use TWT amplifiers which are a common type of VED [1]-[2].

1.1.1 Traveling-Wave Tube (TWT)

TWTs amplify electromagnetic power to levels ranging from a few watts to a few mega-watts. These provide the largest bandwidth among all high power VEDs. TWTs have advantages of ultra-wide bandwidth, high power, high gain, high efficiency, ultra small size and low weight. It is a superior technology choice for applications such as communication satellite transmitters, high-power radar systems and ECM systems. TWTs exist for frequencies ranging from below 1 GHz to over 100 GHz, and are being developed for THz frequency; average power ranges from watts to hundreds of kilowatts, and efficiencies can range from 30-70% [3].

Figure 1.1: Simplified schematic diagram of a traveling-wave tube (TWT).

Figure 1.1 shows a simplified schematic diagram of a TWT. It consists of an electron gun, a circular helix slow-wave structure (SWS), a collector and an array of focusing magnets. The entire device from the electron gun to the collector is enclosed in a vacuum envelope. Two primary elements of a TWT are an electron beam and a traveling electromagnetic wave. The emitted electrons from the cathode of the electron gun are pushed by electric force due to the biasing of electrodes and

co-propagate with the traveling electromagnetic wave. The spent electron beam is recovered in a collector. The array of focusing magnets can be solenoid, permanent magnets or periodic-permanent-magnets (PPM). The magnetic field is oriented along the direction of flow of the electron beam and it is used to force the electrons to travel within the helix SWS. The electromagnetic wave is guided by a waveguide structure, which is usually a SWS. As shown in Fig. 1.1, the electron beam is placed within or near the SWS so that it can interact with the electric field of the electromagnetic wave in the interaction region. For this interaction, it is essential that the electrons and the electromagnetic wave have similar velocities. This situation is referred to as ‘velocity synchronism’. Once the velocities of the electron beam and the electromagnetic wave are synchronized, a net kinetic energy transfer from the electron beam to the electromagnetic wave and amplification of the electromagnetic wave can take place.

1.1.2 Slow-Wave Structures (SWSs) for TWTs

A key component of the TWTs is the SWS that slows down the phase velocity of the electromagnetic wave, enabling velocity synchronism. The most common SWS is the circular helix [4]-[5], as shown in Fig. 1.2 (a), because of its un-matched capability for strong electron-wave interaction over large bandwidths [3]. Helix TWTs can have bandwidths in excess of two octaves. But the high RF losses and low coupling impedance of a single helix may limit its usage to moderate power levels. Contra-wound circular helix [6], as shown in Fig. 1.2 (b), having two helices with identical radius wound with a reversed pitch angle and coinciding with each other, was developed to improve the power handling capability as well as to reduce interaction with the backward wave of the single helix. As shown in Fig. 1.2 (c), the practical form of a contra-wound helix is realized by a circular ring-bar structure [7]-[8]. Contra-wound circular helix and circular ring-bar structures have limited bandwidths of 10 % to 20 %. Multifilar circular helix [9]-[10], as shown in Fig. 1.2 (d), having more than one helix with the same pitch angle and close spacing with each other, was suggested for the advantages of high coupling impedance and

Figure 1.2: Circular helix and helix-derived slow-wave structures (SWSs): (a) circular helix, (b) cross-wound circular helix, (c) circular ring-bar, and (d) two-filar circular helix.

greater bandwidth over the unifilar helix. The physical structure of a multifilar circular helix is more closely related to the sheath-helix model [11] which ignores the effects of periodicity.

1.1.2.1 Planar SWSs

Interest has been growing in the recent years in high frequency applications of TWTs. As frequency increases, the conventional circular helix or helix-derived structures scale down to very small dimensions which impose difficulty in fabrication. Karp [12]-[13] introduced a planar SWS, as shown in Fig. 1.3, which was realized by an array of transverse slots on the broad wall of a ridge waveguide. The electron beam flows parallel to the slotted wall and experiences an alternate strong and weak field as it passes through the slots. The critical part of such a circuit is the slotted wall which can be realized using photo-etching. Pierce [11] analyzed Karp's SWS by replacing the slotted wall with infinitesimally thin perfectly conducting wires. The approximate structure is identical to a unidirectionally

Figure 1.3: Karp's planar SWS.

Figure 1.4: A pair of unidirectionally conducting (UC) screen.

conducting (UC) screen which, similar to the sheath-helix, ignores the effects of periodicity.

Similar objective of printed-circuit fabrication motivated the studies of planar helix [14]-[23]. The planar helix, as shown in Fig. 1.4, consists of a pair of UC screens that extend infinitely in the y - and z -direction. The top and bottom UC screens are oriented in y' and y'' directions, respectively, and are spaced apart from each other in the x -direction. The symmetry of the structure along the z -axis suggests that the directions of phase propagation and the average Poynting vector averaged over the cross-section will be in the z -direction.

The planar helix structure studied in [16]-[23] is infinite in width along the y -direction. To confine the planar helix structure in the transverse direction, it was

Figure 1.5: Confined planar helix using a pair of metal plates.

suggested in [21] that for one of the modes, the edges of the UC screens can be connected. A structure proposed in [24] for backward-wave oscillators is very similar to the planar helix but with transverse confinement using a pair of metal plates. A confined planar helix [25], as shown in Fig. 1.5, very similar to the structure proposed in [24], was constructed with a microstrip line feed. The top and bottom conducting screens can be realized by printing finite width metal strip arrays on the top and bottom of a dielectric substrate using printed circuit techniques. Although the confined planar helix can be realized easily, the structure exhibits a narrow range of frequencies [24].

Another possibility of transverse confinement of planar helix is using vertically inclined metal strips [26]-[28]. Such a structure leads to the so-called rectangular helix. The rectangular helix, as shown in Fig. 1.6, is realized by winding a metal strip in a rectangular shape with a certain pitch angle. The rectangular helix can be practically useful since the structure is confined in width. However, the vertically inclined metal strips are difficult to realize in practice.

In the context of high frequency TWTs, the output power of the conventional cylindrical beam devices falls quadratically with increasing frequency. On the other hand, the output power of sheet beam devices falls linearly with increasing frequency [29]. In addition, sheet beam devices also offer higher beam current capacity, decreased beam voltage, lower magnetic field requirement and increased

Figure 1.6: Rectangular helix.

bandwidth. Many studies have focused on formation of sheet beam and its transportation [30]-[33]. Planar SWSs are more suitable to interact with sheet beams. A flattened helix [34] is a flattened version of the circular helix by winding the helix on a dielectric slab or on two dielectric rods with an inner metal shield. However the structure of the flattened helix is not suitable for microfabrication. Other planar structures reported in [35]-[39] may also be suitable for interaction with a sheet electron beam; however, the helix-based structures offer larger operating bandwidths since their dispersion characteristics can be tailored by the design of the supports and the envelop of the structure [40].

1.2 Motivation

Research on miniaturized TWTs has steadily grown in the recent past due to the advancement of microfabrication techniques. Semiconductor device fabrication involves steps like photolithography and chemical etching for pattern transfer of two-dimensional integrated circuitries. Microfabrication adapts the semiconductor device fabrication technologies to fabricate three-dimensional micromachined electromechanical systems (MEMS), such as pressure sensors, accelerometers and gyros [41]. Microfabrication has also been applied to fabricate radiofrequency (RF), biomedical, and optical devices. Various techniques, such as lithography,

electroplating, and micro-molding (LIGA), deep reactive ion-etching (DRIE), electrical discharge machining (EDM) etc., have been established for producing three-dimensional structures suitable for high frequency VEDs [42].

Many applications like detection of concealed weapons, chemical threat detection, unmanned aerial vehicles (UAV) and combat identification require high frequency and small size amplifiers. Miniaturized TWTs can have widespread applications not only in wide-band communications, imaging radar, spectroscopy etc. but also in unexplored scientific, industrial and medical applications [43]-[47]. Currently, several factors remain as key challenges for fabricating micron-size structures for VEDs, such as bonding of two halves of the circuits, reducing the surface roughness, reducing the processing time, packaging of complete devices, beam generation and beam transportation. Research in VEDs for easy fabrication and low-cost mass production while exhibiting good thermal dissipation remains crucial.

The need for miniaturization of TWTs and the challenges in the fabrication of SWSs for such TWTs provide the motivation for the research reported in this thesis.

1.2.1 The Proposed SWS

A new planar helix SWS which uses straight-edge connections (PH-SEC) at the conjunction ends of a planar helix is proposed in this thesis. Figure 1.7 shows the PH-SEC SWS immersed in free space. The top and bottom strips are inclined at

Figure 1.7: Proposed SWS for TWT.

Figure 1.8: Plan view of idealized (a) circular helix, (b) rectangular helix and (c) PH-SEC in free space. (d) Unfolded view of the corresponding helices.

angle of $\pm\Psi_l$ with respect to the y -axis. The straight-edge connections subtend zero angle, $\Psi_2 = 0$, with respect to the x -axis. For practical applications, the structure could be supported by dielectric substrates.

The PH-SEC can be considered to be derived from a pair of parallel UC screens [18]-[19], or from the rectangular helix [26]. The PH-SEC retains the broad bandwidth feature of the conventional circular helix. Unlike the rectangular or the circular helix, the straight-edge connections can be realized easily using printed circuit and microfabrication techniques. In addition, the PH-SEC can have a rectangular cross-sectional geometry which can be suitable for interaction with a sheet electron beam.

As shown in the plan view in Fig. 1.8 (a), the circular helix with a certain radius a_i can be considered to be obtained by winding a conducting wire or tape in a helical fashion on a circular cylinder. Similarly, a general rectangular helix, as shown in Fig. 1.8 (b), can be obtained by replacing the circular cylinder by a rectangular cylinder of cross-section width $2b$ and height $2a$. In the rectangular helix, the connections between any two opposite sides require inclined conductors. The PH-SEC, as shown in Fig. 1.8 (c), is also obtained using a rectangular cylinder;

however, in this case, the inclined conductors in the vertical direction are replaced by straight connections. Due to the straight-edge connections in the PH-SEC, the parts of length $2a$ are not visible in the 2D views.

The unfolded views of the circular helix, general rectangular helix and PH-SEC are shown in Fig. 1.8 (d). As seen in Fig. 1.8 (d), for a fixed period L and fixed cross section perimeter P_c , the pitch angle for PH-SEC, Ψ_1 , is greater than the pitch angle for both circular and general rectangular helices, Ψ , since the straight-edge connections have 0° pitch angles. Since the pitch angle is related to phase velocity, it can be seen that the phase velocity of the PH-SEC is less than or equal to that for a general rectangular helix with identical cross-sectional perimeter and equal inclination of the parts with length $2b$.

1.3 Objectives

As mentioned in Section 1.1, the circular helix and the other helix-derived structures pose fabrication difficulty especially for micron size features. In this context, the idea of straight-edge connections for planar helix appears very attractive due to easier fabrication. The objective of the work in this thesis is to study and develop a PH-SEC SWS which is suitable for TWT applications. This objective includes the following goals:

1. The primary target is to understand the propagation characteristics of the proposed PH-SEC structure. A mathematical model based on the field theory approach needs to be formulated for determining the dispersion characteristics of the PH-SEC. This study is important for understanding the effects of basic dimensions, such as the width, height and period of the PH-SEC, on the dispersion characteristics. More importantly, the study is expected to explain the important effects of the straight-edge connections.
2. The next goal is to modify the basic PH-SEC structure for practical usage. For TWT application, the PH-SEC may be supported by dielectric substrates for convenience in fabrication and/or conduction of heat generated in the structure. In addition, the dispersion characteristics of the PH-SEC can be tailored by

manipulating the extent of dielectric loading. In this context, one needs to study the PH-SEC in the presence of dielectric loading. The effect of multilayer dielectric substrates on dispersion characteristics should be determined using the abovementioned mathematical model. In addition, a model for determining the coupling impedance of the PH-SEC should be developed.

3. Simulation of the PH-SEC using a 3D electromagnetic simulator should be carried out to obtain accurate dispersion and impedance characteristics. The simulated results provide verification of the results calculated based on the theoretical model.
4. The proposed PH-SEC with a practical configuration and suitable input-output couplers needs to be designed and fabricated. Fabrication schemes for PH-SEC using printed circuit fabrication and microfabrication techniques need to be examined. This work is important to show the fabrication feasibility of the straight-edge connections.
5. A circular ring-bar structure, derived from the circular helix, was mentioned in Section 1.1.2. The ring-bar structure alleviates some of the problems that arise in the use of the circular helix. In a similar manner, another goal of this research is to examine a rectangular ring-bar structure with straight-edge connections (RRB-SEC), derived from the PH-SEC.
6. Using the above studies, to design the PH-SEC and RRB-SEC for TWT applications. Hot-test parameters such as output power, gain, operating bandwidth, and electronic efficiency should be estimated.

1.4 Major Contributions of the Thesis

A novel planar helix SWS using straight-edge connections (PH-SEC) has been proposed in this thesis. The PH-SEC has been studied in the context of TWT applications. Major contributions of the thesis may be identified as the following:

1. The proposed PH-SEC structure is suitable for printed circuit fabrication as well as micro-fabrication due to its planar configuration. It offers several advantages. First, printed circuit TWTs can be fabricated at a very low cost compared to the

conventional TWTs due to relatively fewer machined parts required [48]-[49]. Second, an advantage of a planar geometry is the possibility of use of a sheet electron beam. Finally, the PH-SEC does not entail an inherent limitation on its bandwidth since its guiding properties resemble those of a circular helix.

2. In practice, theoretical analysis can provide a good starting point for the various parameters of a structure while 3D electromagnetic simulations can be used for further optimization. In this context, an effective dielectric constant method has been proposed to study the dispersion characteristics and coupling impedance of the PH-SEC.
3. The inclined helix conductors of the PH-SEC can be deposited on dielectric substrates which provide excellent heat conductivity for cooling a tube based on the PH-SEC. In this context, the PH-SEC in the presence of multilayer dielectric substrate has been examined. Several configurations have been proposed for practical realization of PH-SEC. In addition, the study also provides guidelines for flattening the phase velocity versus frequency curves of the PH-SEC.
4. This thesis also proposes a wideband input / output feeding method for PH-SEC using a tapered coplanar waveguide (CPW). Over a decade of bandwidth has been demonstrated.
5. A microfabrication process has been proposed to fabricate the PH-SEC at W-band (75 GHz – 110 GHz). The proposed process can be scaled to produce PH-SEC with different aspect-ratios and smaller sizes for operation at higher frequencies. Surface roughness is considered to play an important role for RF power absorption in high frequency microfabricated structures. A method of estimating the increase in RF power absorption in the PH-SEC due to surface roughness is also proposed.
6. Under certain conditions, the circular-helix-based TWTs may suffer from oscillations [6]. To address this problem, a new structure, rectangular ring-bar with straight-edge connections (RRB-SEC) is proposed. Analogous to the circular ring-bar [7], the new structure has the potential to enable high power operation of planar TWTs operating at millimeter-wave and higher frequencies.
7. As a step towards design of a TWT based on the PH-SEC and RRB-SEC structures, hot-test parameters, such as output power, gain, and efficiency have been simulated for the structures on silicon substrate at W-band.

1.5 Organization of the Thesis

The thesis has been organized into seven chapters. Chapter 1 gives a brief introduction to TWTs and SWSs for TWT application. The motivation behind this research work is also explained by identifying various limitations of the previous research on the circular helix and planar helix SWSs. The idea of straight-edge connections for the planar helix and its advantages are presented in this chapter. This chapter also lists the significant contributions of the present work.

Chapter 2 presents an overview of TWTs. The operating principle of the TWTs is described in some detail. The cold-test parameters for SWSs for TWTs, namely, dispersion characteristics and coupling impedance, are described. The method of analysis to obtain the cold-test parameters is reviewed. A brief review of microfabricated SWSs is also included. The 3D electromagnetic simulation tools for TWT applications are also reviewed in this chapter.

Chapter 3 describes the method of analysis for the cold-test parameters for the PH-SEC. First, the dispersion characteristics of the PH-SEC immersed in free space are calculated using an effective dielectric constant (EDC) method. Next, the PH-SEC in the presence of multilayer dielectric substrates is proposed. Several practical configurations, such as metal shields and vacuum tunnel, are taken into consideration in the analysis. Based on the EDC method, a modified effective dielectric constant (MEDC) method is proposed to obtain the dispersion characteristics for the PH-SEC in the presence of multilayer dielectric substrates. In addition, the coupling impedance of the PH-SEC is approximated using its 2D equivalent circuit. Effects of variations in dimensional and material parameters are studied in detail with a view to obtaining a flat phase velocity versus frequency curve. The calculated results using EDC and MEDC methods are compared with the simulation results obtained using Computer Simulation Technology (CST) Microwave Studio eigenmode solver.

A study of possible realization of straight-edge connections using cylindrical via and tape via is presented in Chapter 4. The effects of thickness and width of the inclined metal strips on the phase velocity and coupling impedance are also presented in this chapter. The focus of the rest and the major part of this chapter is

the design and fabrication of the PH-SEC using printed circuit and microfabrication techniques. Two PH-SEC configurations proposed in Chapter 3, with band edge frequency less than 10 GHz, have been fabricated and measured. A wideband input / output coupler has been designed using a CPW. PH-SEC on silicon wafer has been fabricated for W-band operation. On-wafer cold-test measurements have been carried out on a number of fabricated structures. The parameters measured are return loss, attenuation and phase velocity and the results cover the frequency range from 70-100 GHz. Cold-test parameters of the PH-SEC have also been obtained using simulations. The effects of fabrication, such as surface roughness, have also been presented.

Chapter 5 presents a detailed study of the properties of the RRB-SEC structure with square and rectangular cross sections. The simulated dispersion characteristics and coupling impedance for the RRB-SEC are compared with those for the circular ring-bar, circular helix and PH-SEC with a view to assess the suitability of the RRB-SEC for high power applications. The effects of structure dimensions on the properties of the RRB-SEC are studied. Two RRB-SEC configurations that can be microfabricated on a silicon wafer are described. A W-band RRB-SEC for the simpler of the two configurations on silicon, with CPW feed, has been designed and microfabricated. The on-wafer cold-test measured S -parameters, covering the 80 - 110 GHz frequency range, are also presented. The surface roughness and contact resistance are also accounted for in the simulations.

The study of the hot-test parameters for a low power square PH-SEC on a thin silicon substrate with a trench in silicon beneath the structure for TWT amplifier operation at W-band is presented in Chapter 6. These results are obtained using CST 3D Particle-In-Cell (PIC) solver. The hot-test parameters obtained using a simplified CPW coupler and a rectangular waveguide coupler are compared. The linear and non-linear gain of the TWT are examined; the spectral purity and possibility of oscillations in the hot structure are also considered. A high power square RRB-SEC with a similar silicon configuration is also studied in this chapter.

Chapter 7 concludes this thesis by summarizing the important findings and results of this research. It also points out certain limitations of the PH-SEC SWS. Some suggestions for future explorations leading to possible improvements and extensions of the current work have also been mentioned in this chapter.

CHAPTER 2

OVERVIEW OF SLOW-WAVE STRUCTURES (SWSs) FOR TRAVELING-WAVE TUBES (TWTs)

2.1 Introduction

The application of high power vacuum electron devices (VEDs) has been steadily growing in the commercial satellite communication systems, broadcasting, and microwave ovens for military, industry, as well as home use. VEDs stand out from the solid state devices due to reliable performance at high power, high frequency, high efficiency and lower cost [50]-[51]. In recent years, the development of solid state power amplifier devices using wide band gap semiconductor materials, such as Gallium Nitride (GaN) and Silicon Carbide (SiC) [52], has improved their capability with respect to high power and high frequency operations. Therefore, the VEDs at lower microwave frequencies become less attractive. However, the solid state devices suffer from relatively low efficiency and smaller bandwidth [50].

The traveling-wave tube (TWT) provides the largest bandwidth among all high power VEDs. TWTs represent the large majority of all sales of such devices for all applications other than the domestic microwave oven. These serve as the final stage amplifiers in almost all communications satellite transmitters. These are also used as the high-power amplifier in high-power radars and electronic countermeasure (ECM) systems [1]. With their ultra-wide bandwidth, high power, high gain, high efficiency, and small size and weight, TWT-based microwave power modules (MPM) are the ideal technology for unmanned aerial vehicles (UAV) and tethered decoys [1]. The ultra-wide bandwidth of an MPM eliminates the need for multiple transmitters for multiple applications in several frequency bands.

TWTs are also suitable for application in modern high-speed terrestrial communication systems. High-data-rate communication systems such as personal

communication systems, local area networks and high definition broadcast systems require high capacity channels with increased output power and amplification bandwidth [53]. Such wide bandwidths and high power levels cannot be supported by solid-state power amplifiers. Development of miniaturized TWTs should make them even more attractive for the applications mentioned so far. In addition, there is a strong demand for high power sources for advanced radar and communication systems between 100 and 300 GHz [33]. These systems have applications in wide band communication, imaging radars, spectroscopy, and in many more unexplored areas of scientific, industrial and medical applications [29]. Going up in frequency, THz devices can be used for imaging tools for security purpose, medical and agriculture [29]. Miniaturized TWTs fabricated using techniques of microfabrication are very promising as microwave amplifiers for such high frequency systems. The compact size requirements for many of these high frequency sources require microfabricated slow-wave structures (SWSs).

A review of the SWSs for TWTs and of the relevant literature is presented in this chapter. In Section 2.2, various helix-related SWSs and the interaction of electromagnetic wave and electron beam is discussed. Several reference books [11], [54]-[58] have been used for gathering this information. Several methods of analysis for determining the dispersion characteristics and coupling impedance of the helical SWSs are also reviewed in this section. Section 2.3 reviews briefly the microfabricated SWSs reported in the literature for TWT applications. Section 2.4 focuses on the simulation tools for designing TWTs. This section explains the motivation for choosing the particular simulation tools used for the research work reported in this thesis.

2.2 Traveling Wave Interaction Theory

The use of electron beam interaction along an electromagnetic circuit in the context of vacuum tubes was proposed by Lindenblad [59]. He invented the e-beam-driven device, which was capable of amplifying a wide band of frequencies, in 1940. Helix TWT, in which the input signal is fed directly to the helix rather than modulation of electron gun configuration introduced by Lindenblad, was invented by R. Kompfner

Figure 2.1: Radiofrequency (RF) charge and electric field pattern for a helix slow-wave structure (SWS) [54]-[55].

[60]-[62] in 1947 in Britain. The first practical helix TWT was developed and analyzed by J.R. Pierce and L.M. Field [4]-[5], [11], [63]. Field theory, electron bunching and gain analysis were investigated subsequently by others [64]-[66]. Pierce provided important analysis for TWT parameters [11] for general SWSs. The analysis of helix using field theory approach has been given in [64].

A schematic diagram and several components of a TWT have been introduced in Section 1.1.1. These components include an electron beam and a SWS that supports an electromagnetic wave. The energy transfer between the beam and the wave can only take place when the propagation velocities of the electron beam and the electromagnetic wave are synchronized. Once the two velocities are synchronized, the electron beam is velocity modulated by the electromagnetic wave. To explain the bunching process, understanding the field distribution of the SWS will be useful. Figure 2.1 (a) shows the radiofrequency (RF) charge and the electric field pattern of a helix SWS [54]. The electric field lines extend from the positive charge to the negative charge with very strong field at the axis of the structure. The corresponding axial electric field is plotted in Fig. 2.2 [54].

The electron beam propagates at a speed similar to the axial velocity of the electromagnetic wave, through the center of the helix and experiences alternating

Figure 2.2: Axial electric field for velocity modulation [54]-[55].

accelerating and decelerating forces along the helix. The electrons are forced towards region **A** causing formation of bunches. The fields produced by the bunched electrons cause the electron charges on the helix to move away from the region **A** and towards the region **B**. The speed of electromagnetic wave must be slightly slower than the speed of electrons in order to produce a growing field in the helix [54]. Due to the increasing space charge forces the bunching of electron beam is restricted after a certain interaction length. Besides that, the speed of electrons also slows down due to the loss in kinetic energy. As a consequence, eventually the bunched beam moves out of phase with the growing signal and the amplitude of the signal drops [54]-[55].

In the following subsections, we first describe the fundamentals of SWS and then go on to describe a practical circular helix SWS for TWT applications. The method of analysis for the circular helix, planar helix, and the rectangular helix are also described. The modes of operation for the ring-bar SWSs are also reviewed.

2.2.1 Slow-Wave Structure (SWS)

A SWS is an electromagnetic waveguiding structure in which the electromagnetic waves can propagate with a phase velocity v_p smaller than the speed of light in free space c ($\approx 3 \times 10^8$). A SWS finds applications in phase shifters or delay lines since the guide wavelength is very much smaller than the wavelength of the fast wave structures, such as the rectangular or cylindrical waveguide. This helps in

Figure 2.3: Typical dispersion curve of a periodic structure.

minimizing the circuit dimensions. The SWSs also find application in TWTs since the phase velocity of the guided electromagnetic wave needs to be synchronized with the velocity of the electron beam which is typically $1 / 10$ of c . The slow-wave properties can be realized by introducing axial periodicity in the structure.

Understanding the dispersion characteristics of a periodic structure is crucial for SWS design. Figure 2.3 shows the typical dispersion curve (normalized wave number, kL , versus normalized phase constant, βL) of a periodic structure. The wave number is related to frequency, f , through

$$k = \omega \sqrt{\mu \varepsilon} \quad (2.1)$$

where $\omega = 2\pi f$, $\mu = \mu_0 \mu_r$, and $\varepsilon = \varepsilon_0 \varepsilon_r$. There exist different modes of propagation, indicated by solid lines, at different frequency bands. The frequency bands with some value of βL are called pass-bands, while the frequency bands with no value of βL correspond to stop-bands. The dispersion curve is useful for finding the phase velocity ($v_p = \omega / \beta$), and group velocity ($v_g = d\omega / d\beta$) of the wave. There are forward and backward waves which correspond to the positive and negative group velocity, respectively. There are multiple value of β for a given k , corresponding to the various space harmonics of the periodic structure. The phase constant of the n^{th} space harmonic is defined as

$$\beta_n = \beta_0 + \frac{2\pi n}{L} \quad (2.2)$$

where L is the period of the structure, β_0 is the phase constant of the fundamental mode, and n is a positive or negative integer or zero. The $\beta = k_0$ line indicates the phase constant corresponding to free space propagation. Therefore, the phase velocity is lower than the speed of light in free space for the lowest frequency band (Fig. 2.3). A beam line (dotted line in Fig. 2.3) indicates the ‘phase constant’ of the electron beam. The points of intersection between the dispersion curves and the beam line indicate the frequencies of possible beam wave interaction. For TWT operation as a forward wave amplifier, the group velocity must be in the same direction as the beam velocity. The frequencies for $\beta L = \pi$ are called π -point frequencies or band-edge frequencies.

The axial electric field of a periodic structure can be written as

$$E_z = \sum_{n=-\infty}^{\infty} A_n F_n(x, y) e^{j\beta_n z} \quad (2.3)$$

where A_n is the field amplitude of the n^{th} space harmonic, and $F_n(x, y)$ is a field function describing the variations in the transverse directions. The periodicity of the structure is accounted for in the phase term. The axial electric field is the sum of all the spatial harmonics. This expression is called the Floquet spatial harmonic expansion.

2.2.2 Circular Helix SWS

The circular helix SWS has been adopted in TWT applications extensively; it offers a very broad bandwidth and is relatively easy to construct. As shown in Fig. 2.4 (a), the helix is obtained by winding a tape conductor, with a width SW and a thickness ST , in a helical fashion on a circular cylinder with radius a_i . The dispersion characteristics of the helix SWS can be controlled through the pitch angle Ψ and / or period L . The pitch angle is related to the period and radius by

Figure 2.4: (a) Important dimensions of a circular helix. (b) Configuration of circular helix in a traveling-wave tube (TWT).

$$\Psi = \tan^{-1} \left(\frac{L}{2\pi a_i} \right) \quad (2.4)$$

Since the electromagnetic wave primarily follows the helix [67], the axial component of the phase velocity of an electromagnetic wave on the helix can be approximated as

$$v_p \approx c \sin(\Psi) \quad (2.5)$$

Figure 2.4 (b) shows the configuration of a circular helix in a TWT. In a practical device the helix is supported inside a metal tube with the help of several dielectric

rods. The dielectric medium in the vicinity of the helix affects the phase velocity as well as the strength of the axial electric field of the structure. The simplest support system consists of dielectric wedge bars which may be considered as a dielectric tube with portions radially cut out from it [40]. The supporting dielectric rods also serve as a heat removal channel for the heat generated in the helix. The dielectric rods should have excellent mechanical contact with the helix in order to reduce the thermal resistance. For efficient cooling of the helix, ceramic type of dielectric rods is used. However, the dielectric constants for ceramic materials are high which significantly impacts the strength of the axial electric field. Hence thin rectangle dielectric rods, as depicted in Fig. 2.4 (b), are used typically. The metal shield also serves as a vacuum envelope. Similar to the effect of dielectric loading, the phase velocity as well as the strength of the axial electric field of the structure can be affected by the distance between the helix and the metal shield. The helix is normally made of tungsten or molybdenum tape which provides sufficient mechanical strength and high melting point. Materials with high thermal conductivity, such as beryllium oxide (BeO) and anisotropic pyrolytic boron nitride (APBN), are normally used for the dielectric support rods.

There are both a forward and a backward wave traveling in a helix. For TWT amplifier operation, as depicted in Fig. 2.5, the electron beam interacts with the forward wave which propagates from the electron gun to the collector. The backward wave travels from the collector end to the gun end of the helix and carries power reflected from the load. These reflections are caused by the mismatch at the output RF coupler. The backward wave is undesired for a TWT amplifier because it can result in oscillation or variation in gain with respect to frequency. To prevent the feedback provided by the growth of backward wave, attenuators and severers are used in the TWTs. An attenuator can be achieved by coating the dielectric support rods with a lossy material such as carbon film. Both forward and backward waves are attenuated by the lossy material. However, the small loss in the forward wave mode due to the lossy material [68] can be compensated by increasing the beam wave interaction length. For high power tubes, one or several severers are used to suppress the backward wave. Figure 2.5 shows a two-section helix loaded with two severers. The severers prevent the reflected signals from the load end from reaching the input port. Since the velocity modulated beam is traveling across the severed section, the

Figure 2.5: Two-section helix containing sever [54]-[55].

forward wave continues to grow. For maximum efficiency, the sever region should be kept short and the gain of the input section should be kept low.

The RF input and output coupler design is crucial for feeding and tapping the RF signal to and from the helix efficiently over a wide frequency range. As a rule of thumb, the return loss of the input or output couplers should be less than -20 dB. This requires impedance matching between the helix and other waveguiding structures. Figure 2.6 shows the transition between the helix and a coaxial line. The impedance of the helix varies significantly with frequency. To match the impedance of the helix over a wide frequency range, as shown in Fig. 2.6 (a), the geometry of the helix is changed gradually into a coaxial line. Although this technique offers very broad matching bandwidth, the overall length of the tube is increased due to the tapered pitch spiral slot and tapered outer conductor. Hence, it increases the weight of the tube. Another technique, as shown in Fig. 2.6 (b), tapers the center conductor of the coaxial line for impedance matching. This technique reduces the taper length but the welded joint between the helix and the coaxial line is quite fragile. To resolve this issue, the center conductor of the coaxial line should be supported at the output window of the tube.

The dispersion characteristics and the coupling impedance of the helix are important parameters for TWT operation. These parameters are known as cold-test

Figure 2.6: Transition from helix to coaxial line: (a) tapered pitch spiral slot, and (b) tapered coaxial line [54]-[55].

parameters. There are two analytical techniques for determining the dispersion characteristics of the helix, namely, field and equivalent circuit analyses. For field analysis, sheath-helix [11] and tape-helix [69] models have been proposed in the literature. The coupling impedance is a measure of the strength of interaction of the electromagnetic wave in the SWS with an electron beam. In the following subsections, the sheath-helix model for the analysis of circular helix and the coupling impedance is reviewed since this model is utilized in the analysis of the PH-SEC in Chapter 3.

2.2.2.1 Sheath-Helix Model

The helix SWS can be replaced by an approximate sheath-helix model that can be analyzed as a simple boundary-value problem. The sheath-helix model replaces the actual helix by a circular cylindrical sheath. While the actual helix is made of a tape or wire with finite width and thickness, the sheath-helix assumes an infinitesimally thin cylindrical sheath with a radius equal to the mean radius of the actual helix. The sheath conductivities are anisotropic and are infinite and zero in directions parallel and perpendicular to the helix winding direction, respectively. The sheath-helix approximation is closer to the actual situation if the helix is wound with a small

pitch angle. A validity condition for the sheath-helix approximation is defined as:

$$\frac{\lambda_g}{L} \gg 1 \quad (2.6)$$

where λ_g is the guide wavelength. The guide wavelength is related to the phase constant and phase velocity as:

$$\lambda_g = \frac{2\pi}{\beta} = \frac{v_p}{f} \quad (2.7)$$

For relatively small pitch angles, such that $\sin \Psi \approx \tan \Psi$, another validity condition for sheath-helix approximation can be derived, namely, the free-space wavelength (c/f) is a lot greater than the cross-sectional perimeter of the helix, $2\pi a_i$, where a_i is the mean radius in the sheath-helix model, and much larger than the periodicity L (see equation (2.5)).

For simplicity, in the following the helix is assumed in the free space, without the dielectric support rods and metal envelope. The effect of dielectric support rods and metal envelope can be included by using a dielectric loading factor. Due to the anisotropic conductivity of the sheath-helix, one expects both transverse electric (TE) and transverse magnetic (TM) modes to be present. The electric and magnetic fields inside ($r < a_i$) and outside ($r > a_i$) the helical sheath, in cylindrical coordinates, may be written as [40]:

$$\begin{aligned} E_{zn} &= A_n I_0 \{\gamma_n r\} + B_n K_0 \{\gamma_n r\} \\ H_{zn} &= C_n I_0 \{\gamma_n r\} + D_n K_0 \{\gamma_n r\} \\ E_{\theta n} &= \left(-j\omega \frac{\mu_n}{\gamma_n} \right) [C_n I_1 \{\gamma_n r\} - D_n K_1 \{\gamma_n r\}] \\ H_{\theta n} &= \left(j\omega \frac{\epsilon_n}{\gamma_n} \right) [A_n I_1 \{\gamma_n r\} - B_n K_1 \{\gamma_n r\}] \end{aligned} \quad (2.8)$$

where the subscript n refers to the region of the structure; $n = 1$ and 2 refer to the region inside and outside the helix, respectively. A_n , B_n , C_n and D_n are field constants

for each region. For the helix in free space, both regions have $\epsilon_n = \epsilon_0$, $\mu_n = \mu_0$, and $\gamma_n = \gamma = (\beta^2 - \omega^2 \mu_0 \epsilon_0)^{1/2}$. The term $\exp j(\omega t - \beta z)$ is understood in the field expressions in (2.8). I_0 and K_0 are zeroth-order modified Bessel functions of the first and second kind, respectively, while I_1 and K_1 are first-order modified Bessel functions of the first and second kind, respectively. Due to the nature of the field distribution, B_1 , A_2 , D_1 , and C_2 are equal to zero.

There are four boundary conditions for the sheath-helix. Three of the boundary conditions are as follows:

$$\begin{aligned} E_{\theta 1} \cos \Psi + E_{z1} \sin \Psi &= 0 \\ E_{\theta 2} \cos \Psi + E_{z2} \sin \Psi &= 0 \\ H_{\theta 1} \cos \Psi + H_{z1} \sin \Psi &= H_{\theta 2} \cos \Psi + H_{z2} \sin \Psi \end{aligned} \quad (2.9)$$

The fourth boundary condition can be chosen from the following:

$$\begin{aligned} E_{z1} &= E_{z2} \\ E_{\theta 1} &= E_{\theta 2} \end{aligned} \quad (2.10)$$

After applying the four boundary conditions, one can obtain the characteristic equation of the helix as

$$\frac{k_0 \cot \Psi}{\gamma} = \left[\frac{I_0 \{ \gamma a_i \} K_0 \{ \gamma a_i \}}{I_1 \{ \gamma a_i \} K_1 \{ \gamma a_i \}} \right]^{\frac{1}{2}} \quad (2.11)$$

where the left hand side can be replaced by

$$\frac{k_0 \cot \Psi}{\gamma} = \frac{k_0 \cot \Psi}{\left(\beta^2 - k_0^2 \right)^{\frac{1}{2}}} \quad (2.12)$$

The dispersion curves of a helix with a particular mean radius a_i and pitch angle Ψ can be plotted using (2.11). However, one should also understand that the sheath-helix model only produces the dispersion curve for the fundamental mode since the

actual axial periodicity is not considered in the model. Thus the effects of spatial harmonics cannot be obtained through the sheath-helix model. More rigorous analysis methods such as the tape-helix model can be applied to find out this information. In addition, analyses considering the dielectric support rods and metal shield have been extensively discussed in the literature [70]-[75].

2.2.2.2 Coupling Impedance

The other cold-test parameter is coupling impedance, K_c , with units of Ω . The coupling impedance is different from the characteristic impedance, Z_0 , of a transmission line, and it is defined as:

$$K_c = \frac{|V_z|^2}{2P} \quad (2.13)$$

where V_z is the longitudinal voltage which may be found by taking the negative line integral of the axial electric field intensity over $\lambda_g / 4$ [56]. Thus

$$V_z = - \int_{z=\frac{\lambda_g}{4}}^0 E_z dz = - \int_{z=\frac{\lambda_g}{4}}^0 E_z \{0\} \sin(\beta z) dz = \frac{E_z \{0\}}{\beta} \quad (2.14)$$

Hence, (2.13) can be rewritten as

$$K_c = \frac{E_z \{0\}^2}{2\beta^2 P} \quad (2.15)$$

More often, the axial electric field intensity, $E_z \{0\}$, in (2.15) is replaced by the average electric field intensity, $E_{z,av}$, over the cross section of the electron beam. The term P is the power propagating through the structure and it is related to average Poynting vector as:

$$P = \frac{1}{2} \text{Re} \int_z (E \times H^*) (2\pi r) dr \quad (2.16)$$

The power propagating through the structure can also be related to group velocity v_g through the following

$$P = Wv_g \quad (2.17)$$

where W is the energy stored per unit axial length of the structure.

2.2.3 Ring-Bar SWS

Several helix-derived SWSs have been introduced in Chapter 1 for possible high-power helix-based TWTs. The helix-derived SWS such as the ring-bar SWS can have larger transverse dimensions for interacting with high voltage electron beam without giving rise to a backward mode. The ring-bar SWS, as shown in Fig. 2.7, consists of a metal ring with inner radius, thickness and width of a_i , ST and SW , respectively, and a crossover, with arc length of BW , to join the individual rings. A period of the ring-bar structure, L , consists of two rings and two crossovers arranged at 180° angular displacement.

The current paths at the crossovers of the ring-bar are illustrated in Fig. 2.8. Generally two possible current paths can be expected, ie. symmetric mode (or E mode) and anti-symmetric mode (or H mode). For the symmetric mode, E_z exists in the axial direction, and the current has net movement in the z -direction but not in the

Figure 2.7: Important dimensions of the ring-bar SWS.

Figure 2.8: Current paths at the crossovers of the ring-bar SWS: (a) symmetric mode and (b) anti-symmetric mode [7].

θ -direction. There is no effect if a conducting plane is placed normal to the z -axis and centered in the middle of the crossovers. Such a perfect reflector plane can be used to measure the phase velocity and impedance. For the anti-symmetric mode, there are equal current flows in the θ direction for each ring. Hence, the current only has net movement in the θ -direction but no net movement in the z -direction. The anti-symmetric mode cannot be used for interacting with a beam traveling in the axial direction since the axial electric field is zero [7].

As mentioned in Section 1.1.2, ring-bar structure is a practical form of contra-wound helix. Lopes and Motta analyzed the ring-bar structure using contra-wound helix model [76]. Figure 2.9 shows the comparison of dispersion characteristics for the fundamental forward wave mode (mode 0) and backward wave mode (mode -1) between the contra-wound helix and the single-tape helix. The ring-bar has larger variation of phase velocity with respect to frequency compared to that for the single-tape helix. Both structures are assumed to operate at a normalized frequency of around 0.4. The backward wave mode of the single-tape helix has a phase velocity much closer to that for the forward wave mode at the operating frequency which may result in producing oscillations. On the other hand, there is a significant difference in the phase velocity for the forward and backward wave modes for the ring-bar structure.

Ring-bar SWS has more thermal contact with the supporting dielectric material and hence it has a better performance with respect to heat removal from the helix. In addition, the coupling impedance of the ring-bar SWS is also higher,

Figure 2.9: Comparison of normalized phase velocity of the contra-wound helix and single-tape helix [76].

corresponding to a higher gain per wavelength of the device, than that of the simple helix. However, the operating bandwidth of the ring-bar SWS is limited to 10% ~ 20% [54] due to large variation in the phase velocity with respect to frequency.

2.2.4 Planar Helix SWS

As mentioned in section 1.1.2.1, the circular helix SWS is difficult to be fabricated using printed-circuit or microfabrication techniques. These techniques are important for low cost production, in particular for high frequency TWTs. On the other hand, planar SWSs are quite suitable for these fabrication techniques. In addition, a planar structure can be designed to accommodate a sheet electron beam, which offers certain advantages as the frequency of operation increases.

A pair of unidirectionally conducting (UC) screens has been proposed and studied in some detail in the past [16]-[20], [22]-[23]. It consists of a pair of parallel UC screens, as depicted in Fig. 1.4 in Chapter 1, with the top and bottom UC screens conducting in the y' and y'' directions while insulating in the perpendicular directions z' and z'' . These studies show that the guiding properties of a pair of UC

Figure 2.10: The planar helix resembles of a pair of unidirectionally conducting (UC) screens.

screens resemble those of the circular helix, ie. the guided wave slows down by a factor of $\sin(\Psi)$, where Ψ is the pitch angle of the helix or angle of the UC screens. Besides that, the two screens have current flow in directions oblique to the direction of wave propagation. Therefore, a pair of UC screens can be called a flattened helix or a planar helix. These studies also show that the planar helix has a great potential for TWT application since its planar geometry is suitable for fabrication using printed circuit or microfabrication.

In this section, the planar helix with a pair of UC screens printed on a dielectric medium [18], as shown in Fig. 2.10, is revisited. The dielectric medium has a thickness $2a$ and a dielectric constant ϵ_2 . The entire structure is immersed in a medium with a dielectric constant of ϵ_1 . The coordinate system used for the planar helix is cartesian coordinates (x, y, z) . It is assumed that the UC screens are infinite in the y direction and also infinite in the propagation direction (z). The y' and y'' directions make an angle $\pm \Psi_1$ with respect to the y -axis.

2.2.4.1 Field Expressions and Boundary Conditions

The planar helix supports hybrid modes, i.e., both E_z and H_z are simultaneously present. The symmetry of the structure suggests that the general solution can be

decomposed into transverse symmetric (even) and transverse anti-symmetric (odd) solutions. The term transverse symmetric implies that the transverse components E_x , E_y , H_x , and H_y are symmetric with respect to x while the longitudinal components E_z and H_z are anti-symmetric. Similarly, the transverse anti-symmetric solution implies that transverse components are anti-symmetric and the longitudinal components are symmetric. It is assumed that the field quantities have no variation in the y -direction, ie. $\partial / \partial y = 0$. The fields outside the UC screens must decay exponentially whereas the fields in between the UC screens can have hyperbolic variation with respect to the x -direction. Assuming the time dependence as $\exp(j\omega t)$ and that the wave is propagating in the $+z$ direction, the longitudinal components for the transverse anti-symmetric mode can be written as

$$E_{z1} = A_1 e^{-k_{x1}(x-a)} e^{-j\beta z} \quad (2.18a)$$

$$H_{z1} = B_1 e^{-k_{x1}(x-a)} e^{-j\beta z} \quad (2.18b)$$

$$E_{z2} = A_2 \cosh(k_{x2}x) e^{-j\beta z} \quad (2.18c)$$

$$H_{z2} = B_2 \cosh(k_{x2}x) e^{-j\beta z} \quad (2.18d)$$

where the subscripts 1 and 2 represent the regions $|x| > a$ and $|x| < a$, respectively. A_1 , A_2 , B_1 , and B_2 are the amplitude coefficients. The propagation constant β and the transverse decay coefficients k_{x1} and k_{x2} are related by the equations

$$k_{x1}^2 = \beta^2 - k_1^2, \quad k_{x2}^2 = \beta^2 - k_2^2 \quad (2.19)$$

where

$$k_1^2 = \omega^2 \mu_0 \epsilon_1, \quad k_2^2 = \omega^2 \mu_0 \epsilon_2 \quad (2.20)$$

The other transverse field components can be obtained by

$$E_{xn} = \frac{j}{k_{xn}^2} \left(\beta \frac{\partial E_{zn}}{\partial x} + \omega \mu \frac{\partial H_{zn}}{\partial y} \right)$$

$$\begin{aligned}
E_{yn} &= \frac{-j}{k_{xn}^2} \left(-\beta \frac{\partial E_{zn}}{\partial y} + \omega \mu \frac{\partial H_{zn}}{\partial x} \right) \\
H_{xn} &= \frac{-j}{k_{xn}^2} \left(\omega \varepsilon \frac{\partial E_{zn}}{\partial y} - \beta \frac{\partial H_{zn}}{\partial x} \right) \\
H_{yn} &= \frac{j}{k_{xn}^2} \left(\omega \varepsilon \frac{\partial E_{zn}}{\partial x} + \beta \frac{\partial H_{zn}}{\partial y} \right)
\end{aligned} \tag{2.21}$$

where the subscript n represents the region 1 or 2. All $\partial / \partial y$ terms can be ignored since there is no variation in the y -direction.

Following the sheath-helix model, the planar helix must satisfy the following boundary conditions:

$$\begin{aligned}
E_y' &= E_y \cos(\Psi_1) \pm E_z \sin(\Psi_1) = 0 \quad \text{on } x = \pm a \\
E_y'' &= E_y \cos(\Psi_1) \pm E_z \sin(\Psi_1) = 0 \quad \text{on } x = \pm a
\end{aligned} \tag{2.22a}$$

$$\begin{aligned}
E_z' &= \mp E_y \sin(\Psi_1) + E_z \cos(\Psi_1) \quad \text{is continuous across } x = \pm a \\
E_z'' &= \mp E_y \sin(\Psi_1) + E_z \cos(\Psi_1) \quad \text{is continuous across } x = \pm a
\end{aligned} \tag{2.22b}$$

$$\begin{aligned}
H_y' &= H_y \cos(\Psi_1) \pm H_z \sin(\Psi_1) \quad \text{is continuous across } x = \pm a \\
H_y'' &= H_y \cos(\Psi_1) \pm H_z \sin(\Psi_1) \quad \text{is continuous across } x = \pm a
\end{aligned} \tag{2.22c}$$

The conditions (2.22a) and (2.22b) imply that E_y and E_z are continuous across $x = \pm a$. After applying the boundary conditions, the characteristic equation for the transverse anti-symmetric mode is obtained as

$$\frac{\frac{k_1^2}{k_{x1}} + \frac{k_2^2}{k_{x2}} \tanh k_{x2}a}{k_{x1} + k_{x2} \coth k_{x2}a} = \tan^2 \Psi_1 \tag{2.23}$$

The characteristic equation for the planar helix is a transcendental equation. One can obtain the dispersion curves for the planar helix for given a , Ψ_1 , and dielectric properties of the mediums. For decaying fields outside the helix, k_{x1} must be real and positive. On the other hand, k_{x2} may be real or imaginary. For the transverse symmetric case, the ‘cosh’ function in (2.18c) and (2.18d) can be replaced by ‘sinh’

function. The transverse anti-symmetric mode should be chosen for TWT applications since in this mode the E_z component is non-zero at $x = 0$.

2.2.5 Rectangular Helix SWS

The planar helix assumes that the structure is infinite in the y direction. From the point of view of practical application, the structure needs to be confined in the transverse direction. One possibility for this is the rectangular helix, as shown in Fig. 2.11 (a), proposed in 2008 [26]. It is realized by winding a metal strip in a rectangular shape with a certain pitch angle.

For a typical rectangular helix, the pitch angles for the vertical and horizontal metal strips are equivalent, ie. $\Psi_1 = \Psi_2$. The field theory of the rectangular helix immersed in free space, and surrounded by a dielectric substrate and metal shield, has been presented in [26] and [27], respectively. To account for the effect of transverse confinement in the field analysis, the sheath-helix model involving four UC screens, as depicted in Fig. 2.11 (b), has been proposed [26]. In this case, the transverse variation is not ignored, ie. $\partial / \partial y \neq 0$, in the analysis. Since the structure has symmetry, only a quarter of the structure needs to be considered for analysis, ie. region I ($0 < x < a$ and $0 < y < b$), region II ($x > a$ and $0 < y < b$), region III ($0 < x <$

Figure 2.11: (a) Important dimensions of a rectangular helix. (b) Sheath-helix model for the rectangular helix.

a and $y > b$), and region IV ($x > a$ and $y > b$). Omitting $\exp(j\omega t - j\beta z)$, the longitudinal fields for the four regions for the transverse anti-symmetric mode are

$$\text{Region I: } E_{z1} = A_1 \cosh(ux) \cosh(vy) \quad (2.24a)$$

$$H_{z1} = B_1 \cosh(ux) \cosh(vy) \quad (2.24b)$$

$$\text{Region II: } E_{z2} = A_2 \cosh(vy) e^{-u(x-a)} \quad (2.24c)$$

$$H_{z2} = B_2 \cosh(vy) e^{-u(x-a)} \quad (2.24d)$$

$$\text{Region III: } E_{z3} = A_3 \cosh(ux) e^{-v(y-b)} \quad (2.24e)$$

$$H_{z3} = B_3 \cosh(ux) e^{-v(y-b)} \quad (2.24f)$$

$$\text{Region IV: } E_{z4} = A_4 e^{-u(x-a)} e^{-v(y-b)} \quad (2.24g)$$

$$H_{z4} = B_4 e^{-u(x-a)} e^{-v(y-b)} \quad (2.24h)$$

where u and v are transverse decay coefficients in the x - and y -direction, respectively, and are related to the propagation constant β by

$$u^2 + v^2 = \beta^2 - k^2 \quad (2.25)$$

The $A_1, B_1, \dots, A_4$, and B_4 are the amplitude coefficients. Other field components can be obtained using (2.21). For simplification, the weak fields in region IV are neglected. The boundary conditions follow the sheath-helix model, ie. tangential electric field components are continuous at $x = a$ and $y = b$; electric field components along the helix direction are zero; magnetic field components along the helix direction are continuous.

Figure 2.12 shows the normalized phase velocity of the rectangular helix published in Fu's paper [26]. All structures have identical period L . The rectangular

Figure: 2.12 Comparison of normalized phase velocity between the rectangular helix and circular helix [26].

helix is assumed to have an aspect ratio b/a of 4. The phase velocity of three circular helices A, B, and C, with diameter of $2a$ and $2b$, and with cross-section perimeter of $4a + 4b$, respectively, are included for comparison. The phase velocity for the circular helix A and B is much different from that of the rectangular helix since the pitch angles are quite different for these helices. One important observation is that, for a similar cross-section perimeter, the rectangular helix exhibits more dispersion than that for the circular helix. Fu has pointed out that the rectangular helix can have strong axial electric field by decreasing the helix's transverse dimension.

2.2.6 Other Types of SWSs

In general, the configuration of circular helix and helix-derived structures involves dielectric rods for supporting the structure and removing the generated heat. This kind of dielectric loading structures have low to moderate power handling capability due to low thermal dissipation. Other dielectric loading structures include meander-line [77] and biplanar interdigital [78] SWSs.

Other types of SWSs used in TWTs include folded waveguide [79]-[80], coupled cavity [81]-[83], and vane loaded waveguide [35], [84]. These structures are formed completely by metal and their RF operation does not involve dielectric loading. Such complete metal structures can handle higher power due to superior thermal dissipation capability and large transverse dimensions. Typically, the folded waveguide structure offers bandwidths of 20 - 30 %; the coupled-cavity structure offers bandwidths of 10 - 15 %; the staggered double vane-loaded waveguide structure [85] offers 30 % bandwidth.

Meander line SWS described in [37], [48], [77], [86]-[87] is a common planar SWS that can be fabricated using printed circuit techniques and microfabrication techniques. The bandwidth of the meander line structure can be quite limited, typically from 10 % to 20 %. Planar ridge waveguide is another planar structure that has been proposed for TWT applications at millimeter waves [33]. However, the attainable bandwidths, after including the input/output coupling sections, appear to be less than 10%.

2.3 Microfabrication of SWSs

With regard to high frequency VEDs, microfabrication of several SWSs has been reported in the literature. Table 2.1 provides a brief summary of the various microfabricated SWSs for THz VEDs in the last decade. Many micro-fabricated SWSs such as the folded waveguide [88]-[90], coupled cavity [91]-[93], vane loaded waveguide [35]-[36], double corrugated waveguide [94], raised meander-line [37], biplanar interdigital [38], and circular helix [38] and [95] have been successfully developed. Lithography, electroplating, and micro-molding (LIGA) process has gained much attention due to its capability of producing smooth surface and high aspect ratio structures. The advancement of materials research has brought on a wide range of photoresists which do not rely on large and expensive x-ray synchrotron facility. As an example, SU-8 can be patterned using low-cost UV light and one can produce high aspect ratio structures which are required for millimeter-wave frequency range. Since the crosslinked SU-8 is difficult to remove after

Table 2.1: Summary of various recent microfabricated slow-wave structures (SWSs) for THz vacuum electron devices (VEDs).

SWS	PRIMARY AUTHOR	YEAR	FREQUENCY	MAIN PROCESSES	PHOTORESIST
Folded waveguide	S. Bhattacharjee [88]	2004	400 GHz	X-ray LIGA; UV LIGA; DRIE	PMMA; SU-8
	Y. M. Shin [89]	2006	100 GHz	X-ray LIGA	PMMA
	R. Zheng [90]	2010	220 GHz	UV LIGA	SU-8
Coupled Cavity	Y. M. Shin [91]	2003	95 GHz	X-ray LIGA	PMMA
	C. W. Baik [92]	2008	100 GHz	DRIE	Details not given
	O. Kwon [93]	2010	100 GHz	X-ray LIGA	PMMA
Vane loaded waveguide	L. Earley [35]	2006	94 GHz	EDM	Not applicable
	Y. M. Shin [36]	2009	220 GHz	UV LIGA; DRIE	KMPR (UV LIGA)
Double corrugated waveguide	C. Paoloni [94]	2011	0.3 - 2 THz	UV LIGA	SU-8
Raised meander-line	S. Sengele [37]	2009	W-band	DRIE and UV LIGA	AZ1827, AZ 5214E
Biplanar interdigital	M. R. Lueck [38]	2011	650 GHz	CVD, DRIE and UV LIGA	Dry film
Circular helix (Octagonal; square)	J. A. Dayton [95]	2009	95 GHz;	CVD, RIE and UV LIGA	Details not given
	M. R. Lueck [38]	2011	650 GHz		

electroplating, other UV sensitive photoresists, such as KMPR [36], have been reported to replace SU-8.

Unlike the all metal structures, the structures that involve dielectric loading normally require more fabrication steps since the metallization is on top of a patterned dielectric substrate. For instance, the raised meander-line structure [37] involves selectively metalizing the high-aspect-ratio serpentine dielectric ridge. Many efforts have been made to reduce the dielectric surrounding the structure. The fabrication of circular helix [38] and [95] involves a thin sheet of diamond for supporting the structure; such a configuration enjoys excellent thermal dissipation, moderate dielectric loading, high dielectric strength and low loss. However, the circular shape is difficult to realize faithfully in microfabrication and may need to be approximated as octagonal or square [95]. Among the various SWSs, in general the helix structure holds a unique position since the fabrication has less stringent alignment issues due to their broadband nature [38].

Planar SWSs incorporating sheet beam have gained attention especially for THz frequency range due to possibility of fabrication using microfabrication techniques. However, there are two main obstacles for the development of sheet beam TWTs: 1. the formation and transport of a sufficiently high perveance and uniform beam, and, 2. the design and fabrication of suitable SWSs which can be integrated with a sheet beam with high degree of precision [50]. The development of high frequency and high power TWTs remains challenging and relies on the modeling and simulation tools, cathodes, and new fabrication techniques.

2.4 Simulation Tools for TWT Design

Simulation tools are vital for the design of TWTs since the theoretical analyses always require certain assumptions for reducing the complexity in the derivation. It is important for a simulation code to predict the performance with high accuracy especially for developing an actual working device. These performance parameters include the cold-test parameters (dispersion characteristics and coupling impedance) and the hot-test parameters (gain, power, and efficiency). In recent years, the

development of simulation codes has been driven heavily by the industry for accelerating the design process and reducing the overall development cost [96].

A variety of simulators, which involve a full solution of Maxwell's equations, have been developed to solve 3D structures with arbitrary geometries and certain boundary conditions. In general, these simulators can be categorized as based on time domain and frequency domain approaches. For the time domain approach, finite difference time domain (FDTD) [97], transmission-line-matrix (TLM) [98], and finite integration technique (FIT) [99] have been proposed. For the frequency domain approach, finite element method (FEM) has been proposed [100]. Among these two types of approaches, the time domain approach has been dominating in the commercial 3D electromagnetic softwares. There are several key advantages of the time domain approach compared to the frequency domain approach [101]:

1. Arbitrary time signals can be used as excitation
2. Broadband frequency-domain results can be obtained in a single simulation run
3. Transient field effects can be observed
4. Nonlinear effects can be incorporated.

However, the frequency domain approach is suitable for low-frequency problems, highly resonant structures and eigenmode computations. The FDTD method applies rectangular Cartesian mesh on the objects. The curved surfaces are represented by staircase approximation in the rectangular Cartesian mesh. The staircase approximation can lead to a poor estimation of the actual geometry. For TLM method, the number of unknowns per mesh cell is greater than the FDTD or FIT methods. Hence, TLM method requires more computer resources since the required memory space scales linearly with the number of unknowns. FIT uses Maxwell's equations in the integral form and the calculation domain is discretized on a dual grid-complex in space and time, which yields the so-called discrete grid equations. For the FIT method, the voltage along the mesh cell edges and the fluxes through the mesh cell surfaces represent the Maxwell's equations on the grid space which is applicable to both time and frequency domain. Such a representation is applicable to any type of mesh.

The design of a TWT requires at least an electromagnetic simulator and a Particle-In-Cell (PIC) simulator to obtain the cold-test and hot-test parameters,

respectively. The most common electromagnetic simulators in the market for simulation of SWSs include Computer Simulation Technology (CST) Microwave Studio (MWS) [102], and Ansoft High Frequency Structure Simulator (HFSS) [103]. Several PIC simulation codes have been developed for plasma physics including ARGUS [104], MAGIC [105], Christine [106], VORPAL [107], CST MAFIA, and CST PIC [102]. Among these simulators, the CST MWS and CST PIC solvers are chosen here for the simulation of the proposed SWS for TWT application.

CST MWS is a window-based simulator that employs the FIT method for solving Maxwell's equations. It enjoys the benefits of FIT method while providing additional geometry approximation feature, i.e., perfect boundary approximation (PBA) [108]. The PBA maps the simulated structure into a hexagonal mesh and allows the curved surfaces to be defined in a cuboid mesh cell which consists of two different materials. In addition, the thin sheet technique (TST) extends the capabilities of PBA by enabling independent treatment of two dielectric parts of a cell, separated by a metallic sheet. The PBA together with TST provides excellent approximation for curved surface, easy meshing, low memory requirement and high speed for large calculation domains. These provide the motivation for simulating the proposed planar helical SWS using CST since the inclined helix parts can be modeled accurately and effectively. The automatic meshing allows the mesh cells to be partially filled for more accurate representation of shapes [109]. The simulated structures can be drawn in a user-friendly window environment similar to the typical computer aided design (CAD) software. The structures drawn in the CST can also be exported to other standard 2D and 3D CAD tools and hence it reduces the effort in preparing for fabrication. CST MWS has been utilized in the characterization of the cold-test parameters of the helix SWSs extensively [110]-[113].

The charged particles can be included in the CST PIC solver for obtaining the hot-test parameters. The equation of motion and the Lorentz force are included in the calculation to take into account the mutual coupling between the charged particles and the electromagnetic fields. The moving charged particles are considered as current in the evaluation of Maxwell's equations [114]. The PIC simulation requires high computational resources due to a large number of charged particles present in the calculation domain [96]. In addition to the transient

electromagnetic fields, the relativistic motion of the charged particles may have to be simulated. The fields excited by the charged particles have to be considered at every time step [101] which consumes a large amount of time especially for TWTs since the interaction region is typically many wavelengths long. Recently, there have been an increasing number of TWT designs using the CST PIC solver [115]-[117] due to its user-friendly window environment.

2.5 Summary

The existing literature on the slow-wave structures related to the circular helix for traveling-wave tube applications has been reviewed in detail with the focus on the analysis, traveling-wave tube configuration, materials, input and output couplers etc. There exists a clear need of explorations on high frequency slow-wave structures that offer easy fabrication. From this point of view, the planar helix structure appears to be quite attractive. The chapter also includes a brief review of the literature on microfabricated slow-wave structures. The chapter ends with a brief review of the simulation tools that are currently available for the design of TWTs.

CHAPTER 3

COLD-TEST PARAMETERS OF PLANAR HELIX WITH STRAIGHT-EDGE CONNECTIONS (PH-SEC)

3.1 Introduction

While circular helix is the most common slow-wave structure (SWS) for traveling-wave tubes (TWTs), it is not amenable to fabrication using printed circuit or microfabrication techniques. Moreover, since device dimensions scale inversely with frequency, at high frequencies the fabrication of the electron gun, magnetic focusing structure, collector and SWS using conventional manufacturing technology becomes very difficult. The planar helix with straight-edge connections (PH-SEC) has been introduced in Chapter 1. It can be considered to be derived from the rectangular helix [26]. The PH-SEC structure is suitable for printed circuit fabrication as well as microfabrication due to its planar configuration. It offers several advantages. First, printed circuit TWTs can be fabricated at a very low cost compared to the conventional TWTs due to relatively fewer machined parts required [48]-[49]. Second, an advantage of a planar geometry is the possibility of use of a sheet electron beam. Finally, the PH-SEC does not entail an inherent limitation on its bandwidth since its guiding properties resemble those of a circular helix.

Several analysis methods for analyzing helix SWSs have been reviewed in Chapter 2. However, no existing method can be applied directly to the PH-SEC SWS. The field analysis method for the rectangular helix SWS, as shown in Fig. 3.1 (a) with pitch angle $\Psi_1 = \Psi_2$ in Fig. 3.1 (b), has been proposed by Fu [26]. The dispersion characteristics of the rectangular helix have been obtained, using the approximations of (i) sheath-helix, (ii) average boundary conditions on the four sides, and (iii) neglecting the fields in the corner regions outside the helix. Even the

Figure 3.1: (a) Perspective view of a generalized rectangular helix. (b) Unfolded view of a rectangular helix turn. (c) Perspective view of a planar helix with straight-edge connections (PH-SEC).

approximate approach in [26] becomes tedious when applied to multi-layer structures which are likely to arise in practice [27].

The dispersion characteristics and the coupling impedance of a SWS are known as cold-test parameters. The study of cold-test parameters is critical for the design of TWTs. In this chapter, the cold-test parameters of the PH-SEC SWS, as shown in Fig. 3.1 (c) with $\Psi_2 = 0^\circ$ in Fig. 3.1 (b), are obtained using the Effective Dielectric Constant (EDC) method [118] in which one decomposes the original 3D waveguide into two related 2D structures. The EDC method yields accurate results in the frequency range that is far-from-cutoff; moreover, it is simple, fast, and can reduce the complexity of analysis especially for multilayer structures. In Section 3.2, we consider the PH-SEC immersed in free space. The procedure for applying the EDC method to the PH-SEC immersed in free space is established. The performance of the PH-SEC is compared with that of the rectangular helix immersed in free space. The effect of the pitch angle of the PH-SEC on the accuracy of the EDC method is also examined. Section 3.3 presents the analysis for the PH-SEC in the presence of multilayer substrates, a vacuum tunnel within the helical structure, and

metal shield. Such modifications will be required when the planar helix structure is used in TWTs. The general configuration considered for the theoretical analysis can lead to a number of simpler and practical configurations. The dispersion characteristics of various configurations are calculated using Modified Effective Dielectric Constant (MEDC) method which combines the two alternative ways of applying the conventional EDC method and produces more accurate results. Yet, the MEDC method preserves the simplicity of the basic EDC concept. A method of calculation for coupling impedance using 2D approximation of the actual structures is also introduced. Effects of variations in the dimensional and material parameters are studied in detail with a view to obtaining a flat phase velocity vs. frequency curve. Such results are important in determining the performance characteristics of TWTs. In order to validate the theoretical results, the calculated phase velocity and coupling impedance values are compared with simulation results obtained using the Computer Simulation Technology (CST) eigenmode solver.

3.2 PH-SEC Immersed in Free Space

The geometry of the PH-SEC is shown in Fig. 3.1 (c). Using the sheath-helix approximation, the PH-SEC consists of a pair of unidirectionally conducting (UC) screens conducting at an angle $\pm\Psi_1$ with respect to the y -axis. The conjunction ends of the UC screens are joined using straight connections ($\Psi_2 = 0^0$). The UC screens and the straight-edge connections are infinitely thin and are assumed to be perfectly conducting. The structures in Figs. 3.1 (a) and 3.1 (c) are assumed to be immersed in free space with relative permeability (μ_r) and permittivity (ϵ_r) of 1. The dimensions along the x - and y -directions are $2a$ and $2b$, respectively. L is the period of the structure; the period is only related to the width of the structure $2b$ in the y direction since $\Psi_2 = 0^0$. Hence, L is given by:

$$L = 4b \tan(\Psi_1) \quad (3.1)$$

The pitch angle of the overall structure is defined as effective pitch angle, Ψ_{eff} , which is given by:

$$\Psi_{eff} = \tan^{-1} \left(\frac{L}{4a + 4b} \right) \quad (3.2)$$

The effective pitch angle of the PH-SEC is critical for the analysis of dispersion characteristics using the EDC method.

3.2.1 Effective Dielectric Constant (EDC) Method

The EDC method has been extensively applied to open waveguides of rectangular cross-section [118]. The results of the method are accurate in the frequency range which is far-from-cutoff. In this method, the original 3D waveguide is replaced by two related 2D structures, each of which is easy to analyze. However, since the EDC method is developed based on rectangular dielectric waveguides, the method cannot be directly applied to the present structure. The 2D planar helix supports hybrid modes only, unlike the 2D dielectric waveguides which can support TE and TM modes.

Figure 3.2 (a) shows the cross-section of the PH-SEC. Let us consider the case $b/a \geq 1$ first. In the sheath-helix approximation, one has UC screens at $x = \pm a$ and straight-edge connections at $y = \pm b$. Using the EDC method, the structure is analyzed in two steps. In the first step, we consider the x -dependent profile only. This corresponds to a pair of UC screens, separated by a distance $2a$, and infinite in the y -direction, as shown in Fig. 3.2 (b). Thus, the ‘slowing-down’ effect of the helical structure is taken into account in this step. In the second step, we consider the y -dependent profile only, accounting for the transverse confinement in the y -direction. As shown in Fig. 3.2 (c), this corresponds to a symmetric dielectric slab of thickness $2b$, having a dielectric constant ϵ_{eff}' which is obtained in the first step. Four shaded corner regions, as shown in Fig. 3.2 (a), are neglected in the analysis. The details of the analysis are given in the following subsections, assuming $b/a \geq 1$. For the case $b/a < 1$, the sequence of the two steps is reversed; the UC screens with y -dependent profile are considered first, taking into account the ‘slowing-down’

Figure 3.2: Model for applying the Effective Dielectric Constant (EDC) method. (a) Rectangular cross section of the PH-SEC. (b) x -dependent profile. (c) y -dependent profile.

effect; the dielectric slab with x -dependent profile is considered next, accounting for the confinement in the x -direction.

3.2.1.1 x -Dependent Profile

The geometry of the x -dependent profile is shown in Fig. 3.2 (b). This corresponds to a pair of UC screens immersed in free space; the screens conduct in directions $\pm\Psi_1$ with respect to the y -axis. The symmetry of the structure implies that the

solutions can be decomposed into transverse symmetric (even) and transverse anti-symmetric (odd) cases [16], meaning E_x , E_y , H_x and H_y are symmetric or anti-symmetric, respectively, with respect to the x -axis.

To capture the slowing-down effect of the helical conductor in the original 3D waveguide, we use effective pitch angle, Ψ_{eff} , defined in (3.2) as the pitch angle for the UC screens. By substituting (3.1) into (3.2), Ψ_{eff} is related to the cross-section parameters as following:

$$\tan(\Psi_{eff}) = \frac{b}{a+b} \tan(\Psi_1) \quad (3.3)$$

The hybrid mode field solution and characteristic equations for the structure in Fig. 3.2 (b) have been derived in [16] in a manner similar to that described in Section 2.2.4. Choosing the transverse anti-symmetric mode, the characteristic equation is

$$\frac{k^2}{u^2} \cot^2 \Psi_{eff} = \coth(ua) \quad (3.4)$$

where $k^2 = \omega^2 \mu_0 \epsilon_0 \epsilon_{r1}$, u is the decay coefficient in the x -direction and it is related to the phase constant β' for this step as

$$u^2 = \beta'^2 - k^2 \quad (3.5)$$

β' is obtained from (3.4) and (3.5) numerically. Then, the effective dielectric constant, ϵ_{eff}' , for the second step is calculated as:

$$\epsilon_{eff}' = \left(\frac{\beta'}{k} \right)^2 \quad (3.6)$$

One may consider the transverse symmetric mode in a similar manner.

3.2.1.2 y-Dependent Profile

The slowing-down effect of the helical conductor in the original 3D waveguide has been taken into consideration in the x -dependent profile. Therefore the y -dependent

profile (Fig. 3.2 (c)) is considered to be simply an infinite dielectric slab immersed in free space, with a dielectric constant ϵ_{eff}' which is calculated from the x -dependent profile. The dielectric slab supports TM and TE modes which can be even or odd with respect to y . To support the transverse anti-symmetric mode of the original waveguide, and keeping in view that E_x is odd with respect to x in the previous step, here we consider modes with an even variation of E_x with respect to y . Thus we choose TE even mode for this step; this is the lowest order mode for a dielectric slab. The characteristic equation for the TE even mode is [119]:

$$\frac{v_1}{v_2} = \tan(v_2 b) \quad (3.7)$$

where v_1 and v_2 are the y -direction decay coefficients for the free space and substrate regions, respectively, and are related to the overall β in the z -direction by

$$\beta^2 = v_1^2 + k_1^2 = k_{eff}'^2 - v_2^2 \quad (3.8)$$

In (3.8), $k_1^2 = \omega^2 \mu_0 \epsilon_0 \epsilon_{r1}$ and $k_{eff}'^2 = \omega^2 \mu_0 \epsilon_0 \epsilon_{eff}'$ are the wave numbers in the free space and substrate regions, respectively. The dispersion characteristics of the original structure can be obtained by solving the characteristic equations (3.4) and (3.7) numerically.

In order to validate the dispersion characteristics obtained using the EDC method, the PH-SEC has been simulated using the CST eigenmode solver. The model used in the eigenmode solver is described in the next section.

3.2.2 Simulation Model

Full 3D geometry of the PH-SEC is simulated in the CST eigenmode solver. As shown in Fig. 3.3, one period of the PH-SEC is required for the eigenmode solver. The top and bottom faces include finite width but infinitesimally thin perfect electric conductor (PEC) strips inclined at $\pm \Psi_l$ with respect to the y -axis. The straight-edge connections are modeled as straight and finite diameter PEC cylinders. Since the

Figure 3.3: 3D model of the PH-SEC in Computer Simulation Technology (CST) eigenmode solver.

simulation results are to be compared with the results obtained using the EDC method, the strip width and diameter of the cylinders should be similar to that of the sheath helix. Hence, the strip width and diameter of the cylinders are chosen to be small compared to the period. The boundaries along the x - and y -axis are set to PEC since an open boundary (perfectly matched layer) is not accepted in the CST eigenmode solver. Therefore, some distance between the structure and the PEC boundary is required to emulate the fact that the structure is in free space.

3.2.3 Results and Discussion

Figure 3.4 shows the normalized phase velocity for the rectangular helix and the PH-SEC, calculated using the EDC method (solid lines). Also included are simulation results obtained for the original 3D structures shown in Fig. 3.1 (dotted lines). For the rectangular helix with $\Psi_1 = \Psi_2 = 2.86^\circ$ with $a = 0.06$ mm and $b = 0.24$ mm ($b/a = 4$), it is seen that the normalized phase velocity approaches 0.052 at high frequencies. This closely matches the simulated results presented in [26] for the same dimensions. Figure 3.4 also shows the results for the PH-SEC shown in Fig.

Figure 3.4: Comparison of normalized phase velocity between the rectangular helix and the PH-SEC.

3.1 (c) ($\Psi_2 = 0^\circ$). In this case, for $a = 0.06$ mm and $b = 0.24$ mm ($b/a = 4$), the normalized phase velocity approaches a lower value of 0.04 at high frequencies. Further, for $a = 0.24$ mm and $b = 0.06$ mm ($b/a = 1/4$), the normalized phase velocity approaches an even lower value of 0.01. Similarly, for the same cross-section perimeter, choosing $a = 0.15$ mm and $b = 0.15$ mm ($b/a = 1$), the normalized phase velocity approaches 0.026, a value between 0.04 and 0.01. Thus the PH-SEC can offer a phase velocity which is always lower than that for a rectangular helix with the same cross-section perimeter and same pitch angle Ψ_1 . The results also show that the phase velocity decreases monotonically as one moves from the case of large b value to small b value for a fixed cross-section perimeter. This effect can be explained using (3.3), since larger b value causes the effective pitch angle to move closer to Ψ_1 . The PH-SEC has an additional degree of freedom to adjust the phase velocity unlike the conventional rectangular helix or circular helix which can only rely on the single pitch angle. For PH-SEC, one can choose $b \ll a$ to reduce the phase

Figure 3.5: Normalized phase velocity for the PH-SEC for $\Psi_1 = 2.5^\circ, 5^\circ$ and 7.5° with $a = b = 1.2$ mm.

velocity. Except at low frequencies, the results of the EDC method match the simulation results very well. The inaccuracy of the EDC method at low frequencies arises since the fields in the corner regions are neglected [118]; further, a large aspect ratio (b/a) also contributes to this inaccuracy as will be explained later in connection with Fig. 3.6.

Figure 3.5 shows the normalized phase velocity for PH-SEC for different values of Ψ_1 , with $a = b = 1.2$ mm. As expected, when the pitch angle decreases, the phase velocity also decreases. Again, except at low frequencies, the match between the calculated and simulation results is very good. The accuracy of EDC method drops for larger pitch angles since the sheath helix approximation used in the infinite planar helix does not quite hold. Figure 3.6 shows the effect of variations in b/a for the PH-SEC, with the period L and the cross section perimeter fixed at 0.48 mm and 12 mm, respectively. As b/a increases, the simulation results follow more closely the results of the x -dependent profile (infinite planar helix) and the accuracy of the EDC method deteriorates. This indicates that for large width-to-height ratios, it may be

Figure 3.6: Normalized phase velocity for the PH-SEC for $b/a = 1$ and 5 with fixed L and cross-sectional perimeter.

enough to consider just the x -dependent profile; in such cases, the y -dependent profile does not represent the original structure correctly.

3.3 PH-SEC in the Presence of Multilayer Dielectric Substrates

The dispersion characteristics of the PH-SEC immersed in free space have been obtained using the EDC method and the results have been presented in Section 3.2. For TWT application, the PH-SEC may be supported by dielectric substrates for convenience in fabrication and/or conduction of heat generated in the structure. In addition, the dispersion characteristics of the PH-SEC can be tailored by manipulating the extent of dielectric loading. In this section, we study the PH-SEC in the presence of multilayer dielectric substrates which are likely to arise in

Figure 3.7: General configuration of a shielded PH-SEC in the presence of multilayer dielectric substrates and a vacuum tunnel.

practice. Figure 3.7 shows the general configuration considered for the theoretical analysis. The general configuration can lead to a number of simpler and practical configurations as listed in Table 3.1, namely, unshielded normal helix, unshielded normal helix with vacuum tunnel, shielded normal helix with vacuum tunnel, shielded inverted helix with vacuum tunnel and shielded helix with vacuum tunnel.

Figure 3.8 (a) shows the cross-sectional view of the general configuration corresponding to Fig. 3.7. The straight-edge connections ($\Psi_2 = 0^0$) are embedded in the middle substrate layer with dielectric constant ϵ_{r2} and thickness $2a$ in the x -direction. Utilizing the sheath-helix approximation, it is assumed that a pair of UC screens is printed on top and bottom of the middle substrate layer with a pitch angle $\pm\Psi_1$ and width $2b$ in the y -direction. In order to accommodate a sheet electron beam, a vacuum tunnel ($\epsilon_{r1} = 1$) is centered within the helical structure with dimensions $2c$ and $2d$ along the x - and y -directions, respectively. L is the period of the structure. A substrate layer and a vacuum layer with dielectric constant ϵ_{r3} and ϵ_{r1} , and thickness t_3 and t_4 , respectively, are stacked between the UC screens and metal shields. The metal shields and the dielectric substrates are assumed to be infinite along the y -direction. The distance between the edges of the vacuum tunnel and the UC screens is t_2 ; the distance between the edges of the vacuum tunnel and the straight-edge connections is g .

Table 3.1 Configurations for planar helix with straight-edge connections (PH-SEC)

Configuration	Cross-sectional view
A. Unshielded normal helix	
B. Unshielded normal helix with vacuum tunnel	
C. Shielded normal helix with vacuum tunnel	
D. Shielded inverted helix with vacuum tunnel	
E. Shielded helix with vacuum tunnel

Figure 3.9: Normalized phase velocity calculated using the conventional EDC method for the unshielded normal helix (Table 3.1: A) for $b/a = 1$ and 3 with fixed L and cross-sectional perimeter ($\epsilon_{r1} = 1$, $\epsilon_{r2} = 3.02$). Simulation results for the 3D structure are also included.

3.3.1 Problems with the EDC Method

Application of the EDC method to the PH-SEC immersed in free space has been presented in Section 3.2. There are two alternative ways in which the 3D structure can be decomposed into 2D structures. We define these two alternative ways as EDC-Case I and EDC-Case II. In EDC-Case I, the planar helix and dielectric slab are infinite in y - and x -directions, as depicted in Fig. 3.8 (b) and 3.8 (c), respectively; in EDC-Case II, the planar helix and dielectric slab are infinite in x - and y -directions, as depicted in Fig. 3.8 (d) and 3.8 (e), respectively. Depending on the aspect ratio of the original 3D structure, one case may yield more accurate results than the other [120]. Figure 3.9 shows a comparison of the simulated phase velocity of the full 3D structure with the results calculated using EDC-Case I and EDC-Case

II for the unshielded normal helix (Table 3.1: A) for $b/a = 1$ and 3 with the period L fixed as 0.315 mm and the cross-section perimeter $4a + 4b$ fixed as 4.8 mm. The derivation of the characteristic equations for this configuration is given in the following subsections. Simulation results have been obtained by using the CST eigenmode solver. It can be noticed that the EDC-Case I overestimates and EDC-Case II underestimates the phase velocity, even in the far-from-cutoff region. Both cases produce similar magnitude of error for a square PH-SEC. Thus, unlike the case of the PH-SEC immersed in free space, in the presence of dielectric substrate, both cases do not capture accurately the slowness of the PH-SEC. Besides that, for larger aspect ratios, EDC-Case I gives a better estimate in the far-from-cutoff region which implies that the infinite planar helix in the step 1 of the EDC-Case I resembles more closely the PH-SEC with larger aspect ratio. Hence the aspect ratio of the PH-SEC has an impact on the accuracy of each EDC case.

3.3.2 Modified Effective Dielectric Constant (MEDC) Method

Combinations of EDC-Case I and EDC-Case II, so called “dual EDC method”, have been reported in [121] for the analysis of rectangular dielectric waveguides. In the dual EDC method, combination of the possible cases of applying the conventional EDC method is used to cancel the errors produced in each case, and hence a more accurate approximation is achieved. However, the dual EDC method developed for rectangular dielectric waveguides cannot be directly applied to the present structure. The infinite planar helix supports hybrid modes only, unlike the infinite (2D) dielectric waveguides which support TE and TM modes. In addition, in the case of planar helix, the first step – infinite planar helix – plays a dominant role in the solution. Based on the observation of errors for both EDC-Case I and EDC-Case II in the previous sub-section, the modified effective dielectric constant (MEDC) method proposed here combines the results for both cases of the conventional EDC method by using a weighted average of the results in each case. The weights for the 2D waveguides are based on the aspect ratio of the original 3D structure. This is

simpler compared to the dual EDC method and is directly related to the geometry of the problem. The propagation constant values obtained from 2D planar helices infinite in the y - and x -directions, depicted in Fig. 3.8 (b) and (d), respectively, are weight-averaged and form the ϵ_{eff}' for the subsequent 2D dielectric slabs infinite in the x - and y -directions, depicted in Fig. 3.8 (c) and 3.8 (e), respectively. The propagation constant values obtained from the 2D dielectric slabs are then weight-averaged again to represent the propagation constant of the original 3D structure. The details of the analysis based on the MEDC method are given below. It is assumed that $b/a \geq 1$. The description given below is for the general configuration of PH-SEC (i.e., $\Psi_2 = 0$ in Fig. 3.7).

3.3.2.1 Case I (Step 1) - Planar Helix Infinite in the y -Direction

The general configuration shown in Fig. 3.8 (a) is analyzed using the MEDC method. The 3D structure is decomposed into two 2D structures. Following EDC-Case I, the first 2D structure, planar helix which is infinite in y -direction, is shown in Fig. 3.8 (b). It includes a pair of UC screens separated by a distance $2a$ and conducting in directions $\pm\Psi_1$ with respect to the y -axis. The dielectric constant in the middle region, ϵ_{r1}' , should represent the vacuum tunnel as well as the substrate surrounding the vacuum tunnel in the y -direction in the original 3D structure. Hence, ϵ_{r1}' is obtained by taking the weighted average of the dielectric constant in each region along the y -direction

$$\epsilon_{r1}' = \frac{\epsilon_{r1} \times d + \epsilon_{r2} \times g}{d + g} \quad (3.9)$$

Other details of this 2D structure follow from the description of the general configuration in the beginning of Section 3.3. The multilayer 2D structure in Fig. 3.8 (b) is similar to the infinite planar helix structure reported in [23], with the difference that in the present case $\epsilon_{r1}' \neq \epsilon_{r1}$. Fields are assumed to have the form $e^{(j\omega t - j\beta z)}$. The longitudinal field components for the transverse antisymmetric mode for the configuration in Fig. 3.8 (b) are as follows:

$$E_z = \begin{cases} A \sinh[u_4'(f-x)] & e \leq |x| \leq f \\ B_1 e^{u_3'(x-a)} + B_2 e^{-u_3'(x-a)} & a \leq |x| \leq e \\ C_1 e^{u_2'(x-c)} + C_2 e^{-u_2'(x-c)} & c \leq |x| \leq a \\ D \cosh(u_1'x) & |x| \leq c \end{cases} \quad (3.10)$$

$$H_z = \begin{cases} M \cosh[u_4'(f-x)] & e \leq |x| \leq f \\ N_1 e^{u_3'(x-a)} + N_2 e^{-u_3'(x-a)} & a \leq |x| \leq e \\ R_1 e^{u_2'(x-c)} + R_2 e^{-u_2'(x-c)} & c \leq |x| \leq a \\ Q \cosh(u_1'x) & |x| \leq c \end{cases} \quad (3.11)$$

By using the sheath helix model, all fields must satisfy the boundary conditions of the structure shown in Fig. 3.8 (b):

- i. the tangential electric and magnetic field components are continuous at $x = \pm c$;
- ii. the tangential electric fields are continuous at $x = \pm a$;
- iii. the tangential electric field is zero and the tangential magnetic field is continuous along the direction of conduction at $x = \pm a$;
- iv. the tangential electric and magnetic field components are continuous at $x = \pm e$;
- v. the electric field is zero at the metal shield ($x = \pm f$).

Thus the characteristic equation for the transverse antisymmetric mode is derived as:

$$\frac{\frac{\epsilon_{r1}' \left(1 + \frac{u_1' \epsilon_{r2}'}{u_2' \epsilon_{r1}'} T_1 T_2 \right)}{u_1' T_1 + \frac{u_2' \epsilon_{r1}'}{u_1' \epsilon_{r2}'} T_2} + \frac{\epsilon_{r1}' \left(1 + \frac{u_4' \epsilon_{r3}'}{u_3' \epsilon_{r1}'} T_3 T_4 \right)}{u_4' T_4 + \frac{u_3' \epsilon_{r1}'}{u_4' \epsilon_{r3}'} T_3}}{\frac{T_1 + \frac{u_2'}{u_1'} T_2}{u_1' \frac{u_1'}{1 + \frac{u_1'}{u_2'} T_1 T_2}} + u_4' \frac{1 + \frac{u_3'}{u_4'} T_3 T_4}{u_4' \frac{u_4'}{T_4 + \frac{u_4'}{u_3'} T_3}}} = \frac{1}{k_0^2} \tan^2 \Psi_{eff} \quad (3.12)$$

where $T_1 = \coth(u_1'c)$, $T_2 = \tanh(u_2't_2)$, $T_3 = \tanh(u_3't_3)$ and $T_4 = \tanh(u_4't_4)$. u_1' , u_2' , u_3' and u_4' are decay coefficients in the x -direction. The phase constant for this structure

which has finite dimensions in the x -direction is called β'_{x_helix} . The decay coefficients are related to the phase constant β'_{x_helix} as

$$\begin{aligned} u_1'^2 &= \beta'_{x_helix}{}^2 - k_1'^2 & u_2'^2 &= \beta'_{x_helix}{}^2 - k_2'^2 \\ u_3'^2 &= \beta'_{x_helix}{}^2 - k_3'^2 & u_4'^2 &= \beta'_{x_helix}{}^2 - k_0'^2 \end{aligned} \quad (3.13)$$

where

$$\begin{aligned} k_1'^2 &= \omega^2 \mu_0 \varepsilon_0 \varepsilon_{r1} & k_2'^2 &= \omega^2 \mu_0 \varepsilon_0 \varepsilon_{r2} \\ k_3'^2 &= \omega^2 \mu_0 \varepsilon_0 \varepsilon_{r3} & k_0'^2 &= \omega^2 \mu_0 \varepsilon_0 \end{aligned} \quad (3.14)$$

Similar to the structure immersed in free space, the slowing-down effect of the original 3D structure is captured by changing the pitch angle of the UC screens Ψ_l to Ψ_{eff} , which simply follows (3.3). One may consider the transverse symmetric mode in a similar manner.

The characteristic equation for the 2D planar helix, infinite in y -direction, for each configuration listed in Table 3.1 can be deduced from (3.12) easily. For example, by setting $c = 0$, $t_2 = a$, $t_3 = 0$ and $t_4 = \infty$ in Fig. 3.8 (b), the characteristic equation for the unshielded normal helix can be obtained. Similarly, setting $t_3 = 0$ and $t_4 = \infty$ will lead to the characteristic equation for the unshielded normal helix with vacuum tunnel. Table 3.2 summarizes the longitudinal field components and characteristic equation for transverse antisymmetric mode for each configuration corresponding to the Table 3.1. The numerical solution of (3.12) which is the transcendental characteristic equation for the planar helix infinite in y -direction, yields the phase constant β'_{x_helix} .

3.3.2.2 Case II (Step 1) - Planar Helix Infinite in the x -Direction

Following EDC-Case II, the first 2D structure, planar helix which is infinite in the x -direction, is shown in Fig. 3.8 (d). It includes a pair of UC screens separated by a distance $2b$. In a manner similar to the previous sub-section, the dielectric constant in the middle region, ε_{rl} , is obtained by taking the weighted average of the

dielectric constant in the vacuum tunnel as well as the substrate surrounding the vacuum tunnel in the x -direction in the original 3D structure:

$$\epsilon_{r1}'' = \frac{\epsilon_{r1} \times c + \epsilon_{r2} \times t_2}{c + t_2} \quad (3.15)$$

Other details of this 2D structure follow from the description of the general configuration in the beginning of Section 3.3. The characteristic equation for this case can be obtained by setting $c = d$, $t_2 = g$, $t_3 = \infty$, $t_4 = 0$ and $\epsilon_{r3} = \epsilon_{r2}$ in Fig. 3.8 (b). Hence, the characteristic equation for the transverse antisymmetric mode is

$$\frac{\frac{\epsilon_{r1}''}{u_1''} \frac{1 + \frac{u_1'' \epsilon_{r2}}{u_2'' \epsilon_{r1}''} T_1 T_2}{T_1 + \frac{u_2'' \epsilon_{r1}''}{u_1'' \epsilon_{r2}} T_2} + \frac{\epsilon_{r2}}{u_2''}}{u_1'' \frac{T_1 + \frac{u_2''}{u_1''} T_2}{1 + \frac{u_1''}{u_2''} T_1 T_2} + u_2''} = \frac{1}{k_0^2} \tan^2 \Psi_{eff} \quad (3.16)$$

where $T_1 = \coth(u_1'' d)$ and $T_2 = \tanh(u_2'' g)$. u_1'' and u_2'' are decay coefficients in the y -direction. The phase constant for this structure which has finite dimensions in the y -direction is called β'_{y_helix} . The decay coefficients are related to the phase constant β'_{y_helix} as

$$u_1''^2 = \beta'_{y_helix}{}^2 - k_1^2 \quad u_2''^2 = \beta'_{y_helix}{}^2 - k_2^2 \quad (3.17)$$

where

$$k_1^2 = \omega^2 \mu_0 \epsilon_0 \epsilon_{r1}'' \quad (3.18)$$

For the case of unshielded normal helix (Table 3.1: A) the characteristic equation for the 2D planar helix infinite in x -direction is obtained by setting $d = 0$ and $g = b$ in (3.16) as follow:

$$\frac{\frac{\epsilon_{r2}}{u_2''} T_2 + \frac{\epsilon_{r2}}{u_2''}}{u_2'' \frac{1}{T_2} + u_2''} = \frac{1}{k_0^2} \tan^2 \Psi_{eff} \quad (3.19)$$

For the configurations *B* to *E* listed in Table 3.1, (3.16) can be used directly. The numerical solution of (3.16) yields the phase constant β'_{y_helix} .

The phase constant values obtained from (3.12) and (3.16) for the 2D planar helices are weight-averaged based on the aspect ratio of the original 3D structure

$$\beta'_{helix} = \frac{\beta'_{x_helix} \times b + \beta'_{y_helix} \times a}{a + b} \quad (3.20)$$

Through several comparisons of the results from the MEDC method applied to configurations with different aspect ratios, we found that for $b/a \geq 3$, β'_{helix} calculated in (3.20) generally well approximates the phase constant of the 3D structure. More detailed discussion on this aspect is given in Section 3.3.4. For smaller aspect ratios, e.g., $1 \leq b/a < 3$, the subsequent step involving infinite dielectric slabs is required for more accurate approximation. The dielectric constant for the subsequent infinite dielectric slabs is simply the following

$$\varepsilon'_{eff} = \left(\frac{\beta'_{helix}}{k_0} \right)^2 \quad (3.21)$$

3.3.2.3 Case I (Step 2) - Dielectric Slab Infinite in the *x*-Direction

The slowing-down effect of the helix in the original waveguide has been taken into account in the first step, 2D infinite planar helix structures, as described in the previous two subsections. The subsequent 2D structure for Case I, dielectric slab infinite in *x*-direction, is depicted in Fig. 3.8 (c). The dielectric slab of width $2b$, immersed in a medium with dielectric constant ε_{r2} , is considered to have a dielectric constant ε'_{eff} from (3.21). As explained in Section 3.2.1, we choose TE even mode for this step; this is the lowest order mode for a dielectric slab. The characteristic equation for the TE even mode is [119]:

$$\frac{v_1}{v_2} = \tan(v_2 b) \quad (3.22)$$

where v_1 and v_2 are the *y*-direction decay coefficients for the substrate regions with

Table 3.2 Longitudinal field components and characteristic equations for transverse anti-symmetric mode for the configurations corresponding to Table 3.1

Planar helix infinite in y-direction	
<p>A.</p>	<p>Longitudinal field components:</p> $E_z = \begin{cases} Ae^{-u_4'(x-a)} & x > a \\ D \cosh(u_2'x) & x < a \end{cases}$ $H_z = \begin{cases} Me^{-u_4'(x-a)} & x > a \\ Q \cosh(u_2'x) & x < a \end{cases}$ <p>Characteristic equation:</p> $\frac{\frac{\epsilon_{r2}}{u_2'} T_2 + \frac{\epsilon_{r1}}{u_4'}}{u_2' \frac{1}{T_2} + u_4'} = \frac{1}{k_0^2} \tan^2 \Psi_{eff}$
<p>B.</p>	<p>Longitudinal field components:</p> $E_z = \begin{cases} Ae^{-u_4'(x-a)} & x > a \\ C_1 e^{u_2'(x-c)} + C_2 e^{-u_2'(x-c)} & c < x < a \\ D \cosh(u_1'x) & x < c \end{cases}$ $H_z = \begin{cases} Me^{-u_4'(x-a)} & x > a \\ R_1 e^{u_2'(x-c)} + R_2 e^{-u_2'(x-c)} & c < x < a \\ Q \cosh(u_1'x) & x < c \end{cases}$ <p>Characteristic equation:</p> $\frac{\frac{\epsilon_{r1}}{u_1'} \frac{1 + \frac{u_1' \epsilon_{r2}}{u_2'} T_1 T_2}{T_1 + \frac{u_2' \epsilon_{r1}}{u_1'} T_2} + \frac{\epsilon_{r1}}{u_4'}}{\frac{T_1 + \frac{u_2'}{u_1'} T_2}{u_1' \frac{1 + \frac{u_1'}{u_2'} T_1 T_2}{T_2}} + u_4'} = \frac{1}{k_0^2} \tan^2 \Psi_{eff}$

Table 3.2 (continued)

Planar helix infinite in y-direction	
<p>C.</p>	<p>Longitudinal field components:</p> $E_z = \begin{cases} A \sinh[u_4'(f-x)] & a < x < f \\ C_1 e^{u_2'(x-c)} + C_2 e^{-u_2'(x-c)} & c < x < a \\ D \cosh(u_1'x) & x < c \end{cases}$ $H_z = \begin{cases} M \cosh[u_4'(f-x)] & a < x < f \\ R_1 e^{u_2'(x-c)} + R_2 e^{-u_2'(x-c)} & c < x < a \\ Q \cosh(u_1'x) & x < c \end{cases}$ <p>Characteristic equation:</p> $\frac{\frac{\epsilon_{r1}' \left(1 + \frac{u_1' \epsilon_{r2}'}{u_2' \epsilon_{r1}'} T_1 T_2\right)}{u_1' T_1 + \frac{u_2' \epsilon_{r1}'}{u_1' \epsilon_{r2}'} T_2} + \frac{\epsilon_{r1}'}{u_4'} \frac{1}{T_4}}{\frac{T_1 + \frac{u_2'}{u_1'} T_2}{u_1' \left(1 + \frac{u_1'}{u_2'} T_1 T_2\right)} + u_4' \frac{1}{T_4}} = \frac{1}{k_0^2} \tan^2 \Psi_{eff}$
<p>D.</p>	<p>Longitudinal field components:</p> $E_z = \begin{cases} A \sinh[u_3'(e-x)] & a < x < e \\ D \cosh(u_1'x) & x < a \end{cases}$ $H_z = \begin{cases} M \cosh[u_3'(e-x)] & a < x < e \\ Q \cosh(u_1'x) & x < a \end{cases}$ <p>Characteristic equation:</p> $\frac{\frac{\epsilon_{r1}' \frac{1}{T_1} + \frac{\epsilon_{r3}'}{u_3'} \frac{1}{T_3}}{u_1' T_1 + u_3' \frac{1}{T_3}}}{\frac{1}{k_0^2} \tan^2 \Psi_{eff}} = \frac{1}{k_0^2} \tan^2 \Psi_{eff}$

Table 3.2 (continued)

Planar helix infinite in y-direction	
<p><i>E.</i></p>	<p>Longitudinal field components:</p> $E_z = \begin{cases} A \sinh[u_3'(e-x)] & a < x < e \\ C_1 e^{u_2'(x-c)} + C_2 e^{-u_2'(x-c)} & c < x < a \\ D \cosh(u_1'x) & x < c \end{cases}$ $H_z = \begin{cases} M \cosh[u_3'(e-x)] & a < x < e \\ R_1 e^{u_2'(x-c)} + R_2 e^{-u_2'(x-c)} & c < x < a \\ Q \cosh(u_1'x) & x < c \end{cases}$ <p>Characteristic equation:</p> $\frac{\frac{\epsilon_{r1}'}{u_1'} \frac{1 + \frac{u_1' \epsilon_{r2}}{u_2' \epsilon_{r1}'} T_1 T_2}{T_1 + \frac{u_2' \epsilon_{r1}'}{u_1'} T_2} + \frac{\epsilon_{r3}}{u_3'} \frac{1}{T_3}}{\frac{u_1' \frac{u_2'}{1 + \frac{u_1'}{u_2'} T_1 T_2} + u_3' \frac{1}{T_3}}{\frac{u_1' \epsilon_{r2}}{u_2' \epsilon_{r1}'} T_1 T_2}} = \frac{1}{k_0^2} \tan^2 \Psi_{eff}$

dielectric constant ϵ_{r2} and ϵ_{eff}' , respectively, and are related to the phase constant, called β_{y_sub} in this case, as

$$\beta_{y_sub}^2 = v_1^2 + k_2^2 = k_{eff}'^2 - v_2^2 \quad (3.23)$$

In (3.23), $k_{eff}'^2 = \omega^2 \mu_0 \epsilon_0 \epsilon_{eff}'$ is the wave number in the slab region. The same characteristic equation applies to all the configurations listed in Table 3.1. The solution of the characteristic equation (3.22) yields the phase constant β_{y_sub} .

3.3.2.4 Case II (Step 2) - Dielectric Slab Infinite in the y-Direction

The 2D dielectric slab for Case II is depicted in Fig. 3.8 (e). The dielectric slab,

infinite in the y -direction, immersed in free space, is considered to have a dielectric constant ϵ_{eff} ; the slab width is taken as $2a + 2t_3$ in general, since the effect of the dielectric layer ϵ_{r3} has not been taken into account in this case so far. The slab width is changed to $2a$ for the configurations A to C in Table 3.1. The characteristic equation for the TE even mode is:

$$\frac{w_1}{w_2} = \tan[w_2(a + t_3)] \quad (3.24)$$

where w_1 and w_2 are the x -direction decay coefficients for the air and substrate regions, respectively, and are related to the phase constant, called β_{x_sub} in this case, by

$$\beta_{x_sub}^2 = w_1^2 + k_0^2 = k_{eff}^2 - w_2^2 \quad (3.25)$$

The solution of the characteristic equation (3.24) yields the phase constant β_{x_sub} . In the last step, the phase constant values for the 2D infinite dielectric slabs obtained from (3.22) and (3.24) are weight-averaged according to the aspect ratio as follows

$$\beta = \frac{\beta_{y_sub} \times b + \beta_{x_sub} \times a}{a + b} \quad (3.26)$$

to give β which is the final approximation of the phase constant of the original 3D structure.

3.3.3 Coupling Impedance

The value of the coupling impedance, K_c , is crucial for the application of a SWS in TWTs. Equation (2.15) can be rewritten for average coupling impedance as:

$$K_c = \frac{E_{z,av}^2}{2\beta^2 P} \quad (3.27)$$

where $E_{z,av}$ is the average value of the electric field of the transverse antisymmetric mode, calculated over the cross-section of the sheet beam, and P is the total power

propagating along the z -direction. In general, we assume a solid and uniform sheet electron beam with height and width half that of the vacuum tunnel and centered within the structure. For configuration A in Table 3.1, a hypothetical sheet electron beam with height and width half that of the helix is centered within the structure. For evaluating (3.27), the β value of the 3D structure, calculated from (3.20) or (3.26) depending on the aspect ratio, is used. The calculation of power P is simplified, similar to [23], by using the approximation of 2D planar helix which is infinite in y -direction, as shown in Fig. 3.8 (b). For the general configuration, the total power in the transverse antisymmetric mode can be obtained by

$$P = 2(P_i + P_{ii} + P_{iii} + P_{iv})(2b) \quad (3.28)$$

where P_i , P_{ii} , P_{iii} and P_{iv} are the power flows per unit width for the corresponding regions in the upper half of the structure in Fig. 3.8 (b). The width of the 2D structure in the y -direction is taken as $2b$. By applying Poynting theorem,

$$P = \frac{1}{2} \frac{\beta}{\omega \mu_0} \int_x \left(\left(\frac{\mu_0}{\epsilon_k} \right) |H_y|^2 + |E_y|^2 \right) dx \quad (3.29)$$

the power flow in each region is given by the following expressions:

$$P_i = \frac{1}{8u_1'} \frac{\beta}{\omega \mu_0} \left[\sinh(2u_1'c) - 2u_1'c \right] \left[\frac{k_1'^2}{u_1'^2} D^2 + \left(\frac{j\omega\mu_0}{u_1'} \right)^2 Q^2 \right] \quad (3.30a)$$

$$\begin{aligned} P_{ii} = \frac{1}{4u_2'} \frac{\beta}{\omega \mu_0} & \left\{ \left(e^{2u_2't_2} - 1 \right) \left[\frac{k_2'^2}{u_2'^2} C_1^2 + \left(\frac{j\omega\mu_0}{u_2'} \right)^2 R_1^2 \right] \right. \\ & - \left(e^{-2u_2't_2} - 1 \right) \left[\frac{k_2'^2}{u_2'^2} C_2^2 + \left(\frac{j\omega\mu_0}{u_2'} \right)^2 R_2^2 \right] \\ & \left. - 4u_2't_2 \left[\frac{k_2'^2}{u_2'^2} C_1 C_2 + \left(\frac{j\omega\mu_0}{u_2'} \right)^2 R_1 R_2 \right] \right\} \quad (3.30b) \end{aligned}$$

$$\begin{aligned}
P_{iii} = \frac{1}{4u_3'} \frac{\beta}{\omega\mu_0} & \left\{ \left(e^{2u_3't_3} - 1 \right) \left[\frac{k_3^2}{u_3'^2} B_1^2 + \left(\frac{j\omega\mu_0}{u_3'} \right)^2 N_1^2 \right] \right. \\
& - \left(e^{-2u_3't_3} - 1 \right) \left[\frac{k_3^2}{u_3'^2} B_2^2 + \left(\frac{j\omega\mu_0}{u_3'} \right)^2 N_2^2 \right] \\
& \left. - 4u_3't_3 \left[\frac{k_3^2}{u_3'^2} B_1 B_2 + \left(\frac{j\omega\mu_0}{u_3'} \right)^2 N_1 N_2 \right] \right\} \quad (3.30c)
\end{aligned}$$

$$\begin{aligned}
P_{iv} = \frac{1}{8u_4'} \frac{\beta}{\omega\mu_0} & \left\{ 4 \frac{k_0^2}{u_4'} t_4 A^2 + \left[\frac{k_0^2}{u_4'^2} A^2 + \left(\frac{j\omega\mu_0}{u_4'} \right)^2 M^2 \right] \right. \\
& \left. \times [\sinh(2u_4't_4) - 2u_4't_4] \right\} \quad (3.30d)
\end{aligned}$$

Similarly, the power flow for the 2D planar helix for configuration A shown in Table 3.2 is as follows:

$$P_i = \frac{1}{8u_2'} \frac{\beta}{\omega\mu_0} [\sinh(2u_2'a) - 2u_2'a] \left[\frac{k_2^2}{u_2'^2} D^2 + \left(\frac{j\omega\mu_0}{u_2'} \right)^2 Q^2 \right] \quad (3.31a)$$

$$P_{ii} = \frac{1}{4u_4'} \frac{\beta}{\omega\mu_0} \left[\frac{k_0^2}{u_4'^2} A^2 + \left(\frac{j\omega\mu_0}{u_4'} \right)^2 M^2 \right] \quad (3.31b)$$

For configuration B, the power flow in regions (i) and (ii) follows (3.30a) and (3.30b), respectively; the power flow in region (iii) follows (3.31b). For configuration C, the power flow in regions (i), (ii) and (iii) follows (3.30a), (3.30b) and (3.30d), respectively. For configuration D, the power flow in region (i) follows (3.30a) with c changed to a ; the power flow in region (ii) follows (3.30d) with u_4' , k_0 and t_4 changed to u_3' , k_3 and t_3 , respectively. Finally, for configuration E, the power flow in regions (i) and (ii) follows (3.30a) and (3.30b), respectively; the power flow in region (iii) follows (3.30d) with u_4' , k_0 and t_4 changed to u_3' , k_3 and t_3 , respectively. The relations between the constant coefficients A, B₁, B₂, C₁, C₂, D etc. for the general configuration are given in Appendix A. For other configurations, the

relations between the constant coefficients can be obtained in a similar manner by following the boundary conditions stated in sub-section 3.3.2.1.

3.3.4 Results and Discussion

The 3D PH-SEC in the presence of multilayer dielectric substrates is modeled for simulation using the CST eigenmode solver similar to the description in Section 3.2.2. The multilayer dielectric substrates are modeled using lossless dielectric slabs with specific thickness in the x -direction but particularly wide in the y -direction (refer to Fig. 3.3) in order to approximate the infinite width dielectric substrates in the analysis.

3.3.4.1 Comparison of Analytical and Simulation Results

Figures 3.10 (a) and (b) show the normalized phase velocity for the unshielded normal helix (Table 3.1: A) calculated using the MEDC method described in Section 3.3.2, as well as the coupling impedance calculated using the method described in Section 3.3.3. Simulation results for the actual 3D structure are also included for comparison. The period and the cross-sectional perimeter are fixed at 0.315 mm and 4.8 mm, respectively. The curve labeled “MEDC” represents the results calculated using both steps 1 and 2 in the MEDC method, while “MEDC: β'_{helix} ” represents the results calculated using only step 1, i.e., infinite planar helix. As seen in Fig. 3.10 (a), for $b/a = 1$, both steps of the MEDC method are required to get an accurate estimate of the phase velocity. However, for $b/a = 3$, step 1 itself gives accurate results; step 2 actually causes deterioration of results since for such aspect ratios this step does not represent the original structure correctly; this was observed for the EDC method applied to the structure immersed in free space too. Except at relatively low frequencies, the results of the MEDC method match the simulation results very well. The field strength at the corner regions is relatively significant for lower frequencies. The inaccuracy of the EDC method at relatively low frequencies arises

Figure 3.10: (a) Normalized phase velocity and (b) coupling impedance for the unshielded normal helix (Table 3.1: A) for $b/a = 1$ and 3 with fixed L and cross-sectional perimeter ($\epsilon_{r1} = 1$, $\epsilon_{r2} = 3.02$).

since the fields in the corner regions are neglected [120]. Results in Fig. 3.10 (a) indicate that a flatter phase velocity curve can be obtained by lowering the aspect ratio of the helix. The coupling impedance results in Fig 3.10 (b) also match the simulation results very well. This shows that the method of calculation of coupling impedance proposed in Section 3.3.3, using accurate phase constant together with 2D infinite planar helix approximation, works quite satisfactorily.

Figures 3.11 (a) and (b) show the normalized phase velocity and coupling impedance for the unshielded normal helix (Table 3.1: A) for different values of Ψ_l , with $a = b = 0.6$ mm. As expected, when the pitch angle decreases, the phase velocity also decreases. Again, except at low frequencies, the match between the calculated and simulation results is very good. Further, as the pitch angle increases, the actual structure deviates more from the sheath-helix model and the accuracy of the MEDC method reduces.

3.3.4.2 Effects of Different Configurations

The comparison of unshielded and shielded normal helix configurations, both with a vacuum tunnel (Table 3.1: B & C), is shown in Figs. 3.12 (a) and (b). The dimensions for both configurations are $a = 0.3$ mm, $b = 0.9$ mm, $c = 0.2$ mm, $t_2 = 0.1$ mm, $d = 0.6$ mm, $g = 0.3$ mm, $L = 0.315$ mm, $\epsilon_{r1} = 1$ and $\epsilon_{r2} = 3.02$. For the shielded configuration, two values of the metal shield distance from the planar helix, $t_4 = 0.3$ mm and 0.5 mm, are considered. As seen in Fig. 3.12 (a), the metal shield can reduce the phase velocity at lower frequency range and hence produce a flatter phase velocity curve compared to that for the unshielded configuration. The phase velocity curve can be further flattened by reducing the shielding distance t_4 . Figure 3.12 (b) shows that the metal shielding also reduces the variation of coupling impedance over the bandwidth.

Figures 3.13 (a) and (b) show the normalized phase velocity and coupling impedance for the shielded normal helix ($c = 0.2$ mm, $t_2 = 0.1$ mm and $t_4 = 0.5$ mm), shielded inverted helix ($t_3 = 0.5$ mm and $\epsilon_{r3} = 3.02$) and shielded helix ($c = 0.2$ mm, $t_2 = 0.1$ mm, $t_3 = 0.5$ mm and $\epsilon_{r3} = 3.02$) configurations (Table 3.1: C, D & E), all

Figure 3.11: (a) Normalized phase velocity and (b) coupling impedance for the unshielded normal helix (Table 3.1: A) for $\Psi_1 = 3^0, 5^0$, and 7^0 with $a = b = 0.6$ mm, $\epsilon_{r1} = 1$ and $\epsilon_{r2} = 3.02$.

(a)

(b)

Figure 3.12: Comparison of (a) normalized phase velocity and (b) coupling impedance between the unshielded and shielded normal helix with vacuum tunnel (Table 3.1: B & C).

Figure 3.13: Comparison of (a) normalized phase velocity and (b) coupling impedance between the shielded normal helix with vacuum tunnel, shielded inverted helix with vacuum tunnel and shielded helix with vacuum tunnel (Table 3.1: C, D & E).

with a vacuum tunnel, with $a = 0.3$ mm, $b = 0.9$ mm, $d = 0.6$ mm, $g = 0.3$ mm, $L = 0.315$ mm, $\epsilon_{r1} = 1$ and $\epsilon_{r2} = 3.02$. Among these configurations, the shielded normal helix with vacuum tunnel has the widest bandwidth but the largest variation in phase velocity as well as coupling impedance over the bandwidth. The shielded inverted helix and shielded helix with vacuum tunnel can produce much flatter phase velocity curves due to the dielectric loading outside the UC screens; however, for the latter configuration, more dielectric material surrounding the helical structure causes a slower phase velocity. The coupling impedance of the shielded helix with vacuum tunnel is very similar to that for the shielded inverted helix with vacuum tunnel.

Figures 3.14 (a) and (b) compare the effect of different amounts of dielectric-filling (c , t_2 , d and g) inside the helix for the shielded helix with vacuum tunnel (Table 3.1, E) when $a = 0.3$ mm, $b = 0.9$ mm, $t_3 = 0.5$ mm, $L = 0.315$ mm, $\epsilon_{r1} = 1$ and $\epsilon_{r2} = \epsilon_{r3} = 3.02$. Higher dielectric-filling inside the helix produces a slower phase velocity, lower coupling impedance and smaller bandwidth. A flatter phase velocity curve can be obtained by reducing the dielectric-filling inside the helix. Figure 3.15 shows the effect of varying dielectric constants ϵ_{r2} and ϵ_{r3} for the shielded helix with vacuum tunnel (Table 3.1: E) when $a = 0.3$ mm, $b = 0.9$ mm, $c = 0.2$ mm, $t_2 = 0.1$ mm, $d = 0.6$ mm, $g = 0.3$ mm, $t_3 = 0.5$ mm, $L = 0.315$ mm and $\epsilon_{r1} = 1$. As expected, a higher dielectric constant surrounding the helix, e.g. $\epsilon_{r2} = \epsilon_{r3} = 6.15$, reduces the phase velocity, coupling impedance as well as the bandwidth. As indicated by results in Fig. 3.15 (a), for this configuration, the bandwidth for the flattest phase velocity curve ($\epsilon_{r2} = 3.02$ and $\epsilon_{r3} = 6.15$), for $\pm 5\%$ variation, is 85.7% (k_0L from 0.0396 to 0.099). It should be possible to achieve an even flatter phase velocity curve by optimizing the parameters such as the aspect ratio, metal shield distance and dielectric constant of the material surrounding the helix.

3.4 Summary

A new slow-wave structure, planar helix with straight-edge connections, has been studied in detail. In addition to having the wideband properties of the circular helix, this structure has the advantages of compatibility with printed circuit or

Figure 3.14: (a) Normalized phase velocity and (b) coupling impedance for different amount of dielectric-filling inside the helical structure of the shielded helix with vacuum tunnel (Table 3.1: E).

(a)

(b)

Figure 3.15: (a) Normalized phase velocity and (b) coupling impedance for different dielectric constants at the inner and outer of the helical structure of the shielded helix with vacuum tunnel (Table 3.1: E).

microfabrication, and the possibility of using a sheet electron beam. The proposed planar helix slow-wave structure with straight-edge connections has been investigated with the help of analysis and simulations. The cold-test parameters, namely the dispersion characteristics and coupling impedance, of the proposed structure immersed in free space and immersed in multilayer dielectric substrates have been studied. It is shown that the dispersion characteristics of the proposed structure approximated using the effective dielectric constant method agree well with the simulation results in the far-from-cutoff region. A modified effective dielectric constant method has been applied to the proposed structure in the presence of multilayer dielectric substrates. The coupling impedance of the structure has also been estimated using the 2D planar helix approximation. The phase velocity and coupling impedance results calculated using the proposed methods agree well with the full 3D simulation results in the far-from-cutoff frequency range. Several possible configurations, which take into account practical modifications such as a vacuum tunnel, metal shield and multilayer substrates, have been examined and the effects of these modifications on phase velocity and coupling impedance have been presented.

The results presented in this chapter assume that all the metal parts are perfectly conducting with infinitesimally small thickness and radius so as to keep the analysis simple. The next chapter describes the physical realization of the planar helix with straight-edge connection using printed circuit and micro-fabrication techniques.

CHAPTER 4

FABRICATION AND CHARACTERIZATION OF PLANAR HELIX WITH STRAIGHT-EDGE CONNECTIONS (PH-SEC)

4.1 Introduction

The analytical and simulation results for the cold-test parameters for the planar helix with straight-edge connections (PH-SEC) immersed in air or multilayer dielectric substrates have been presented in Chapter 3. To keep the analysis simple, the PH-SEC is assumed to consist of infinitesimally thin and perfectly conducting wire. Thus, the effect of width and thickness of the inclined conductors and straight-edge connections is ignored.

In this chapter, we first study the possible realization of the straight-edge connections using cylindrical or tape vias, as shown in Fig. 4.1, and the effect of dimensional parameters, such as the thickness and width of the inclined metal strips, on the phase velocity and coupling impedance. The realization of the PH-SEC slow-wave structure (SWS) using printed circuit fabrication is described next. Design of wideband coplanar waveguide (CPW) feed for two of the configurations proposed in Chapter 3 together with measured results is presented to confirm the ease of fabrication, low loss, and the wideband potential of the PH-SEC. The measured S -parameters and phase velocity of the fabricated structures over the frequency range 0.5 - 10 GHz are compared with simulation results obtained using CST Microwave Studio (MWS). In addition, the PH-SEC with coplanar ground plane is proposed to reduce the variation of the phase velocity over the frequency range of operation.

As the frequency of operation increases, the microfabrication techniques become necessary for precise fabrication of the SWSs. Several PH-SEC SWSs incorporating the CPW feed have been designed and microfabricated for operation at W-band (75 GHz - 110 GHz). A microfabrication process has been proposed to

Figure 4.1: (a) Perspective view and (b) cross-sectional view of a planar helix with straight-edge connections (PH-SEC) using cylindrical vias. (c) The perspective view of the PH-SEC using tape vias.

produce high-aspect-ratio planar helical structures which are essential for sheet electron beam traveling-wave tube (TWT) applications. The photoresists used in this process can be easily stripped after electroplating of the three-dimensional structures. On-wafer cold-test measurements have been carried out on a number of fabricated structures. The parameters measured are return loss, attenuation, and phase velocity. Cold-test parameters of the SWS have also been obtained using simulations and the effects of fabrication, such as surface roughness, have been

accounted for by estimating effective conductivity of different parts of the microfabricated structures. Effects of silicon wafer resistivity and thermal issues have also been discussed. Planar helical SWSs fabricated in this manner have applications in TWTs operating at millimeter-wave and higher frequencies.

4.2 Effects of Dimensional Parameters of PH-SEC

The PH-SEC consists of three metal layers, as shown in Fig. 4.1 (a) and (b); the bottom horizontal layer, the top horizontal layer, and the straight-edge connections, in the form of cylindrical vias, that connect the two horizontal layers. The top and bottom layer conductors have thickness ST and width SW . Each cylindrical via has a diameter of VD and it connects top and bottom ring pads which have a diameter RD ($> VD$). Figure 4.1 (c) shows another possible realization of the straight-edge connections using tape vias; the tape thickness and width are identical throughout the helical turn. From the point of view of ease of fabrication, cylindrical vias, with ring pad, offer better connections between the horizontal layers and the straight-edge conductors. L is the period of the structure. The inner dimensions of the PH-SEC structure along the x - and y -directions are $2a$ and $2b$, respectively. The center of the PH-SEC structure coincides with the origin $(0, 0)$.

Figure 4.2 shows a comparison of the phase velocity and coupling impedance between the circular tape helix, rectangular tape helix and PH-SEC. These parameters are obtained using the CST eigenmode solver. All structures are immersed in free space. As explained in Section 3.2.2, the perfect electric boundary is put sufficiently far away from the PH-SEC in order to reduce its effect on the properties of the PH-SEC. For the PH-SEC, the realization of straight-edge connections using both cylindrical or tape vias is considered. Referring to Fig. 2.4 (a), the cross section circumference for the circular tape helix is defined using the inner radius a_i of the tape as

$$P_c = 2\pi a_i \quad (4.1)$$

Figure 4.2: Simulated (a) normalized phase velocity and (b) coupling impedance for circular helix, rectangular helix and PH-SEC for $b/a = 1, 23$ and $1/23$ with fixed P_c and L .

Similarly, the cross-section perimeter for the rectangular tape helix, referring to Fig. 2.11 (a), and PH-SEC is defined using the inner width $2b$ and height $2a$ as

$$P_c = 4a + 4b \quad (4.2)$$

The cross-section perimeter P_c , period L , tape width SW and tape thickness ST are assumed to be the same for all the structures. The frequency is normalized using an equivalent radius a_{eq} . For square and rectangular geometries, the equivalent radius is defined as

$$a_{eq} = \frac{P_c}{2\pi} \quad (4.3)$$

For the circular geometry, a_{eq} is the inner radius a_i . Unlike the circular helix, the rectangular helix and PH-SEC can have the freedom of different aspect ratios b/a . The detailed dimensions of each structure are listed in Appendix B. We first compare the circular helix with the case of a square cross section ($b/a = 1$). All the results in this category are close to each other with only minor differences since the period and cross section perimeter are the same for the three structures. The circular helix offers the flattest phase velocity, but the coupling impedance is also the lowest. The phase velocity of the rectangular helix is identical to that for the PH-SEC with tape vias, and is higher than that of the other structures. The square cross section tends to have a slightly higher phase velocity than that for the circular cross section; this is attributed to coupling of the fields near the right angle bends for the square geometry. For the PH-SEC with cylindrical vias, due to the vias and the ring pads (see Fig. 4.1(a)), the ‘effective perimeter’ is greater than that for the tape vias; this leads to a slightly lower phase velocity for the case of cylindrical vias.

In general, as the aspect ratio increases, the frequency range of operation increases and the coupling impedance improves significantly for the rectangular helix and the PH-SEC since the helical arms are brought closer to the axis of the structure. Figure 4.2 shows these results for $b/a = 23$. Further, unlike the rectangular helix, for the PH-SEC the phase velocity and coupling impedance for $b/a = 23$ and $1/23$ are different since the proportion of straight-edge parts is quite different in the two cases. For $b/a = 1/23$, the straight-edge connections at the two ends of a helical arm are very close to each other and carry currents in opposite directions. Hence,

Figure 4.3: Simulated axial component of the electric field E_z plotted in x - y plane at different phase for PH-SEC for $b/a = 1$.

there is a strong coupling between these straight-edge connections, resulting in a higher dispersion and lower coupling impedance.

Figure 4.3 shows the plots of the simulated axial component of the electric field (E_z) in the x - y plane at different phase values for PH-SEC for $b/a = 1$. The small black box at the bottom of each graph indicates that the fields are captured at the midpoint of the bottom inclined helix conductor along the z -direction. The strongest E_z field is rotating along the helical turn since the PH-SEC is a non-resonating structure. The field strength reduces gradually from near the metal part to the center of the structure ($x = y = 0$). Although the field is captured at a certain point in the z -direction, the average field along the z -direction suggests that the E_z field is symmetric along x - and y -axis which is corresponding to the transverse antisymmetric case in the dispersion analysis in Chapter 3.

The effect of variation in the period L has been reported in Chapter 3. The phase velocity and the coupling impedance increase for a larger period. Next, we study the effect of variations in ST , and VD (together with SW and RD), on the phase velocity and coupling impedance for the PH-SEC with cylindrical vias. The detailed dimensions of each structure are also listed in Appendix B. Figure 4.4 shows that the phase velocity and the bandwidth decrease for thicker metal strips due to an increase in the average turn length. Such an effect is quite obvious, especially for the higher

Figure 4.4: Simulated (a) normalized phase velocity and (b) coupling impedance for the PH-SEC with cylindrical vias, with different values of the metal strip thickness ST for $b/a = 1$ and 23.

aspect ratio. Generally, the coupling impedance increases for thin metal strips. The variation of the coupling impedance, however, does not show a clear pattern for $b/a = 23$.

4.3 Printed Circuit Fabrication of PH-SEC

Several possible configurations for PH-SEC in the presence of multilayer dielectric substrates have been introduced in Chapter 3. As proof of concept, for the unshielded normal helix (Table 3.1: A) and unshielded normal helix with air tunnel (Table 3.1: B), the design and fabrication using printed circuit techniques are described in this section. The design of wideband CPW feed and structure assembly are also described. Measured S -parameters, phase velocity, and attenuation are presented for the frequencies 0.5 - 10 GHz. In addition, the effect of coplanar ground planes, on the dielectric surfaces which also contain the top and bottom inclined helix strips, is also examined.

4.3.1 Unshielded Normal Helix

Unshielded normal helix configuration, as shown in Fig. 4.5 (a), is the simplest configuration for PH-SEC fabricated using printed circuit techniques. This configuration does not possess a tunnel within the PH-SEC for the electron beam; hence the structure can be easily fabricated using a single piece of printed circuit board (PCB). Roger RO3203 PCB with dielectric constant ϵ_{r2} of 3.02 and 17 μm copper cladding is chosen for this fabrication. The dimensions for the PH-SEC, following the notation in Fig. 4.1, are: $2a = 1.524$ mm, $2b = 7.5$ mm, $ST = 17$ μm , $SW = 1$ mm, and $L = 2.1$ mm. The environment surrounding the fabricated structure is air ($\epsilon_{r1} \approx 1$). The cylindrical vias are realized using plated-through-hole (PTH) technology with $VD = 0.5\text{mm}$ and $RD = 0.7$ mm.

Figure 4.5: (a) Cross-sectional view of the unshielded normal helix. (b) Schematic of the proposed coplanar waveguide (CPW) feed. (c) Photo of the fabricated structure with 25 periods. All dimensions are in millimeter.

4.3.1.1 Configuration for Coplanar Waveguide (CPW) Feed

Figure 4.5 (b) shows a part of the schematic of a CPW feed integrated with the unshielded normal helix structure. The CPW section starts from a via-end of the helix structure, followed by a tapered section to join a 50Ω CPW section which can be connected to a standard SMA connector. The length of the tapered CPW section and the impedance of the initial CPW section are important parameters to achieve wideband impedance matching. In addition, reduction in $2b$ and increase in L for the first turn can improve the matching by lowering the helix impedance. The CPW ground planes should extend for a few turns of the helix for better matching at the low frequency end. The detailed dimensions for the designed CPW feed are given in Fig. 4.5 (b). In order to provide same side feeding and to accommodate an electron gun and collector, the CPW feed at both ends of the helix structure is bent at right angle to the helix structure. Air bridges are added to ensure that the CPW ground planes are at the same potential. Figure 4.5 (c) shows a photograph of the fabricated structure with twenty five helix turns printed on the Roger RO3203 PCB.

4.3.1.2 Measurement Results

Figure 4.6 presents the measured and simulated S -parameters of the fabricated structure shown in Fig. 4.5 (c) over the frequency range from 0.5 - 10 GHz. The simulation results have been obtained using the CST MWS transient solver and include dielectric and conductor losses. The measured S -parameters match very well with the simulated ones. The frequency range over which S_{11} is below -10 dB extends from 1 GHz to 8 GHz, i.e., three octaves. The insertion loss (S_{21}) increases as frequency increases mainly due to the dielectric loss in the substrate and the conductor loss in the PTH.

Figure 4.7 shows the measured phase velocity values obtained using the phase difference of S_{21} , $\Delta \angle S_{21}$, between two fabricated structures with different number of periods (25 and 30 periods), with Δn being the difference in the number of periods, but with identical CPW feeds. The measured phase constant, β , is obtained as

Figure 4.6: Measured and simulated S -parameters of the fabricated structure shown in Fig. 4.5.

Figure 4.7: Measured, simulated and calculated normalized phase velocity of the fabricated structure shown in Fig. 4.5.

$$\beta = \frac{\Delta \angle S_{21}}{\Delta n \times L} \quad (4.4)$$

The phase velocity, v_p , can be obtained as ω / β . The simulated phase velocity results obtained from the eigenmode solver as well as the transient solver for the fabricated structure are also shown in Fig. 4.7. To obtain the phase velocity values in the transient solver, one plots the E_z field along the direction of propagation at a fixed instant of time. The measured phase velocity values are well matched with the simulated ones obtained from the transient solver and eigenmode solver, as well as the calculated ones obtained using modified effective dielectric constant (MEDC) method.

4.3.2 Unshielded Normal Helix with Air Tunnel

To accommodate an electron beam, a tunnel is added to the unshielded normal helix configuration. Figure 4.8 (a) shows the cross sectional view of the unshielded normal helix with air tunnel. The dimension for the PH-SEC, following the notation in Fig. 4.1, is: $2a = 2.439$ mm, $2b = 7.5$ mm, $ST = 17$ μ m, $SW = 1$ mm, $VD = 0.5$ mm, $RD = 0.7$ mm, and $L = 2.1$ mm. The air tunnel is embedded inside the PCB, Roger RO4003, with dielectric constant, ϵ_{r2} , of 3.55 and 17 μ m copper cladding. The

Figure 4.8: (a) Cross sectional view of the unshielded normal helix with air tunnel.

(b) Schematic of the PH-SEC with CPW feed.

air tunnel has a height $2c$ and a width $2d$ of 0.813 mm and 6 mm, respectively. Such an air tunnel within the dielectric substrate is difficult to realize using a single piece of PCB, hence multilayer dielectric substrates are used to build the structure. More detailed description of the structure assembly is given in the following subsection. A CPW feed has been designed for coupling the input and output microwave power. Similar to the description in Section 4.3.1.1, a $120\ \Omega$ CPW section (W-S-W: 1.5 mm- 1 mm- 1.5 mm), as shown in Fig. 4.8 (b), starts at the end of the helix and is tapered for impedance matching with a $50\ \Omega$ CPW section (W-S-W: 0.25 mm- 3.5 mm- 0.25 mm). Once again, the pitch angle for the first turn of the helical structure is increased to lower the helix impedance.

4.3.2.1 Fabrication and Measurement Results

The structure shown in Fig. 4.8 (b) has been fabricated using three layers of RO4003, each with a thickness of 0.813 mm, which are stacked together to produce an overall height of 2.439 mm of the helix structure. As shown in the photograph in Fig. 4.9 (a), the middle layer of the fabricated structure consists of two pieces of unmetallized RO4003, laterally separated by 6 mm to realize the rectangular air tunnel. The straight-edge connections are realized by soldering short pieces of silver plated copper wire with 0.5 mm diameter. Figure 4.9 (b) shows the entire structure held together by nylon screws and nuts.

Figure 4.10 presents the measured and simulated S -parameters over the frequency range from 0.5 - 7 GHz. The measured S -parameters match very well with the simulated ones. The -17 dB return loss (S_{11}) of the fabricated structure covers the frequency from 3.5 GHz to 4.6 GHz. Figure 4.11 shows the measured and simulated normalized phase velocity values. The method of phase velocity measurement follows the description in Section 4.3.1.2. The measured phase velocity values also match very well with the simulated values.

Table 4.1 shows the measured attenuation of the fabricated PH-SEC. The measured attenuation is obtained using the difference in the magnitude of S_{21} for two fabricated structures, which are identical except for having different number of helical turns (25 and 30). For comparison, a shielded circular helix with tungsten

(a)

(b)

Figure 4.9: Photograph of the (a) side-view and (b) top-view of the fabricated unshielded normal helix with air tunnel with 30 periods.

($1.9\text{E}7 \text{ S/m}$) for helix, copper ($5.8\text{E}7 \text{ S/m}$) for metal shield and beryllium oxide (loss tangent = 0.003, $\epsilon_r = 6.7$) for three supporting rods is also simulated in CST MWS. The dimensions of the circular helix are chosen so that the band edge for both the PH-SEC and the circular helix is similar. The simulated attenuation of the circular helix is also included in Table 4.1. The unshielded normal helix with air tunnel, constructed on RO4003 dielectric (loss tangent=0.0027, $\epsilon_r = 3.55$), and copper for helix and solder tin ($8.7\text{E}6 \text{ S/m}$) at the joints of vias and ring pads, exhibits an attenuation similar to that for the circular helix with three support rods. Although PH-SEC shows a higher attenuation at lower frequencies, the attenuation of the circular helix is obtained with ideal input and output RF couplers in simulation.

Figure 4.10: Measured and simulated S -parameters of the fabricated structure shown in Fig. 4.9.

Figure 4.11: Measured and simulated normalized phase velocity of the fabricated structure shown in Fig. 4.9.

Table 4.1 Measured attenuation of the fabricated structure shown in Fig. 4.9

Frequency (GHz)	Measured attenuation	Simulated attenuation
	PH-SEC	Circular helix
3	0.023 dB/mm	0.011 dB/mm
4	0.05 dB/mm	0.03 dB/mm
5	0.08 dB/mm	0.08 dB/mm

4.3.3 Unshielded Normal Helix with Air Tunnel in the Presence of Coplanar Ground Planes

The dispersion characteristics of the SWS are the main factor which determines the operating characteristics such as the bandwidth of a TWT [122]. Many dispersion shaping methods such as different shapes of metal vanes and dielectric support rods have been investigated for the TWTs based on circular helix SWS [123]-[127] to obtain multi-octave bandwidths while keeping a high value of the coupling impedance. In this sub-section, we modify the unshielded normal helix with air tunnel by introducing coplanar ground planes on the same surfaces as the inclined helix strips and examine the effect of this modification on the dispersion characteristics.

4.3.3.1 Effects of Coplanar Ground Planes

The cross-sectional view of the normal helix with air tunnel is shown in Fig. 4.12 (a); a perspective view of the realization of the coplanar ground planes is shown in Fig. 4.12 (b). The top and bottom dielectric faces consist of arrays of parallel and inclined helix strips. The coplanar ground planes can be added on the same faces of the helix strips. s indicates the gap between the edge of the ring pads and the edge of the coplanar ground planes.

(a)

(b)

Figure 4.12: (a) Cross-sectional view of the unshielded normal helix with air tunnel in the presence of coplanar ground planes. (b) Perspective view of the PH-SEC with coplanar ground planes printed on one face of the dielectric substrate (dielectric substrate is transparent).

The effects of coplanar ground planes for the unshielded normal helix with air tunnel are shown in Fig. 4.13. The dimensions for the PH-SEC, following the notation in Fig. 4.1, are: $2a = 1.524$ mm, $2b = 5.64$ mm, $ST = 17$ μm , $SW = 0.7$ mm, $VD = 0.36$ mm, $RD = 0.7$ mm, and $L = 2.1$ mm. The air tunnel with a height ($2c$) and a width ($2d$) of 0.508 mm and 4 mm, respectively, is embedded in a dielectric substrate with the dielectric constant ϵ_{r2} of 3.02. Figure 4.13 provides a comparison of the phase velocity and coupling impedance for the structures without coplanar ground, with coplanar ground on one face (top or bottom inclined helix strips face) of the substrate, and with coplanar ground on both faces (top and bottom inclined helix strips faces) of the substrate. In general, the structures with coplanar ground planes produce flatter phase velocity versus frequency curve since the presence of

(a)

(b)

Figure 4.13: (a) Normalized phase velocity and (b) coupling impedance of the unshielded normal helix with air tunnel, in the presence of coplanar ground planes on one face and two faces of the dielectric substrate, with $s = 2.5\text{ mm}$ and 0.5 mm .

coplanar ground planes confines the fields within the structure, especially at low frequencies. This effect is stronger for the PH-SEC with coplanar ground planes on both faces. The coupling impedance, in the presence of coplanar ground planes, remains significant although it is reduced compared to the PH-SEC without coplanar ground planes. Figure 4.13 also shows that the phase velocity versus frequency curve can be further flattened by reducing the gap between the edges of the ring pads to the coplanar ground planes.

4.3.3.2 Fabrication and Measurement Results

The PH-SEC with coplanar ground planes can be easily realized with CPW feed. A continuous ground plane for the input and output CPW feed has been designed and fabricated for the unshielded normal helix with air tunnel. As shown in Fig. 4.14, three layers of Roger RO3203, with the dielectric constant of 3.02, thickness of 0.508 mm and copper cladding of 17 μm , are used to construct the helix structure. The fabricated PH-SEC dimensions follow the dimensions stated in Section 4.3.3.1 except that the ring pads are replaced by square pads with 1 mm x 1 mm size. Besides that, the entire structure is held together by stainless steel screws and nuts.

Figure 4.15 shows the measured and simulated S -parameters of the fabricated structure, as shown in Fig. 4.14, with 22 helix periods. The conductor loss and dielectric loss are taken into account in the simulations using the CST MWS. In order to save the simulation time, the value of the maximum energy deviation for steady state is set high. Hence, the simulated S_{11} does not capture multiple reflections at higher frequencies. However, the simulated S_{21} is well matched with the measured values.

4.4 Microfabrication of PH-SEC at W-Band

The fabrication of some PH-SEC structures using printed circuit techniques at frequencies below 10 GHz has been presented in the previous section. To extend the

Figure 4.14: Photos of the fabricated PH-SEC with coplanar ground planes. (a) Top view. (b) Bottom view. (c) Side view.

Figure 4.15: Measured and simulated S -parameters of the fabricated structure shown in Fig. 4.14.

Figure 4.16: Cross-sectional view of the PH-SEC on a silicon substrate.

operating frequency range of the PH-SEC, e.g. K-Band or above, microfabrication is the only solution to produce the required small sizes and three-dimensional structures.

In this section, W-band PH-SEC structures are designed and fabricated using lithography, electroplating and micromolding (LIGA) steps to produce high-aspect-ratio helical structures. Figure 4.16 shows the cross-sectional view of the PH-SEC on a silicon substrate with the dielectric constant of 11.9 and a thickness of h_2 . A layer of silicon dioxide with the dielectric constant of 3.9 and a thickness of h_1 is assumed to be deposited on top of the silicon substrate. Other details of the PH-SEC structures follow the description in the beginning of Section 4.2. The dimensions of the metal shield along the x - and y -directions are $2c$ and $2d$, respectively. It should be noted that the metal shield is only used in the simulations that use eigenmode solver. Also, the metal shield is put sufficiently far away from the PH-SEC in order to reduce its effect on the properties of the PH-SEC. The center of the PH-SEC structure coincides with the origin $(0, 0)$. Table 4.2 summarizes the various dimensions for the fabricated PH-SEC structures on a thick silicon substrate.

A period of the PH-SEC configuration shown in Fig. 4.16 is simulated using the CST eigenmode solver. Figure 4.17 (a) shows the simulated dispersion diagram for the fundamental mode (mode 1) and the next higher order mode (mode 2). For

Table 4.2 Dimensions of the fabricated W-Band PH-SEC

Symbol	Dimension in micrometer (μm)
h_1	3
h_2	750
$2a$	44
$2b$	70, 150, 170, 190, 270
L	92.85, 112.85, 152.85
ST	3
SW	40
VD	40
RD	50

the selected set of dimensions, the fundamental mode shows a bandwidth of ~ 130 GHz. Mode 2 begins at 30 GHz and propagates as a slow-wave in the same direction as the fundamental mode for frequencies below 170 GHz. In fact, the start frequency of mode 2 is highly dependent on the size of the metal shield. In our simulations, rather large shield dimensions are used so as to accommodate the $750\text{ }\mu\text{m}$ thick silicon wafer that is used in the fabrication. By decreasing the dimensions of the shield, the start frequency of mode 2 can be pushed up and the frequency overlap with the fundamental mode can be avoided.

Figure 4.17 (b) shows the coupling impedance K_c for both modes 1 and 2, at four locations along the x -axis, with reference to Fig. 4.16, while keeping $y = 0$. Three of these locations are within the helical structure ($x = -12.5\text{ }\mu\text{m}$, 0 , and $12.5\text{ }\mu\text{m}$) and the fourth one is above the helical structure ($x = 37.5\text{ }\mu\text{m}$). The variation of coupling impedance along the y -axis is ignored since this variation is small for relatively large widths $2b$. The coupling impedance values for mode 2 are negligibly small over the frequency range that overlaps with the fundamental mode at these four locations. Hence mode 2 is insignificant for beam-wave interaction. The coupling impedance values for the fundamental mode are significantly higher than those for mode 2. However, due to the presence of the thick silicon substrate beneath

Figure 4.17: Simulated (a) propagation constant and (b) coupling impedance for the PH-SEC for the configuration shown in Fig. 4.16, with $2b = 190 \mu\text{m}$, $2c = 1556 \mu\text{m}$, $2d = 690 \mu\text{m}$ and $L = 152.85 \mu\text{m}$.

the PH-SEC structure, the variation in the coupling impedance along the x -axis is asymmetric, with generally higher values for negative values of x . This corresponds to higher field concentration closer to the silicon substrate. Thus, the electron beam (circular or sheet beam) may not be launched at the center of the structure. The transverse dimensions and the alignment of the electron beam must be more carefully designed for such an asymmetric field distribution. Alternatively, the amount of silicon beneath the structure can be reduced. This is expected to result in higher field strength and higher coupling impedance values at the centre of the structure.

Several PH-SEC structures with the helix width $2b$ varying from 110 μm to 310 μm , and helix period L varying from 92.85 μm to 152.85 μm , have been designed for 22 and 32 helix turns. The helix height $2a$, via diameter VD , ring diameter RD , strip width SW and strip thickness ST are fixed in these designs. To facilitate on-wafer measurement, CPW feed has also been designed for coupling the input and output microwave power to/from the helical structure. Similar to the printed circuit fabricated PH-SEC, the CPW starts with high characteristic impedance at the end of the helix and is tapered for impedance matching with a 50 Ω CPW section. The taper length varies from 600 μm to 800 μm for different designs. The silicon wafer is assumed to have a high resistivity (1,000~10,000 $\Omega\text{-cm}$). The present design produces simulated return loss better than -20dB and insertion loss better than -5dB over a wide frequency range, using bulk material conductivity values in the CST transient solver. Simulated S -parameters for some of these structures are presented in Section 4.4.2, together with the measured results.

4.4.1 Microfabrication Process

Microfabrication techniques have been briefly touched upon in Section 1.2. Several microfabricated SWSs have been summarized in Section 2.3. In the context of microfabrication techniques that may be suitable for the PH-SEC, microfabrication processes for solenoid inductors, an RF / microwave passive component that has a shape similar to that of the PH-SEC, have been reported extensively [128]-[130].

But the conventional multiexposure single development method adopted in these processes limits the achievable aspect-ratio of the solenoid structure. To fabricate high aspect-ratio PH-SEC, three ultraviolet (UV) LIGA steps, each step yielding one of the three metal layers of the structure shown in Fig. 4.16, are proposed.

Figure 4.18 shows the cross-sectional view at different stages of the proposed microfabrication process for the PH-SEC. A 3 μm silicon dioxide layer is deposited on the 8-inch high resistivity silicon wafer using Novellus PECVD. The bottom metal layer for helical conductors and CPW feeds, as shown in Fig. 4.18 (a), are formed using negative tone photoresist MaN-1440 lift-off process. The lithography parameters of single layer 7 μm thick MaN-1440 can be found in [131]. The metallization, composed of 300 Å titanium (Ti) / 2 μm copper (Cu) and 300 Å Ti / 2 KÅ gold (Au), is deposited using an electron beam evaporator. After wet stripping of MaN-1440, a seed layer, as shown in Fig. 4.18 (b), composed of 300 Å Ti / 2 KÅ Cu, is deposited using Balzer dc magnetron sputtering for electroplating of the straight-edge connections (cylindrical vias).

Figure 4.18 (c) shows the vertical posts formed by lithography of a thick positive tone photoresist layer AZ 40XT-11D and electroplating. The key properties of this resist are vertical profiles, standard wet stripping process, and good plating compatibility [132]. The optimized lithography parameters to produce 44 μm thick and 40 μm diameter holes in a single coating process are: main spin speed / spin time = 700 rpm / 60 s, soft bake / time = 110 °C / 6 mins, exposure energy = 700 mJ cm^{-2} and developing time = 6 mins. The accumulated photoresist at the edge of the substrate is removed by edge bead removal solution to improve the contact between the mask and the wafer during the exposure step. The patterned mold in the photoresist is filled with copper using laboratory electroplating tool. The electroplating time and current are controlled to form the solid vertical posts at the mold height.

A second seed layer, as shown in Fig. 4.18 (d), composed of 2 KÅ Ti / 2 KÅ Cu, is deposited on the thick photoresist using an electron beam evaporator for electroplating of the third metallization layer. A layer of 20 μm thick negative tone dry film photoresist, as shown in Fig. 4.18 (e), is laminated and patterned to form the electroplating mold for the suspended helical bridges. The temperature of the

Figure 4.18: Cross-sectional views of the proposed microfabrication process for PH-SEC. (i) Bottom copper (Cu) lift-off. (ii) First seed layer deposition. (iii) Patterning and electroplating of Cu vias. (iv) Second seed layer deposition. (v) Patterning and electroplating of Cu helical bridges. (vi) Photoresist and seed layers removal.

lamination is optimized at 85 °C to avoid the deformation of the bottom thick photoresist. The helical bridges are then produced by Cu electroplating.

The structure is first released through wet immersions in the dry film stripper. The dry film stripper will not change the condition of the bottom thick photoresist layer. Subsequently, the second seed layer is etched in the Cu and Ti etchants. Similar wet stripping and etching process are used for the AZ 40XT-11D layer and the first seed layer.

4.4.2 Results and Characterization of Additional Loss

Figure 4.19 shows the scanning electron microscope (SEM) micrograph of the microfabricated PH-SEC. The main challenges in this fabrication process include (i) producing high aspect ratio vertical pillars, or vias, using a thick positive photoresist, and (ii) building suspended helix bridges on top of the thick photoresist sacrificial layer. Before the fabrication of the 3D PH-SEC on a high resistivity wafer, the lithography process to produce 44 μm thick and 40 μm diameter holes in a single coating process is optimized using low cost low resistivity silicon wafers. These holes serve as the mold for the vertical pillars. The electroplating time and current are controlled to form the solid vertical pillars up to the mold height. The top surface of the vias has a convex profile, as shown in Fig. 4.19 (a), due to the non-uniform rate of electroplating. The convex profile causes poor connections between the vias and the helix bridges. Figure 4.19 (b) shows an unsuccessful via to bridge joint due to insufficient via height. To achieve good connections and maintain the thickness of the suspended bridges at 3 μm , the vias need to be over-plated so that most of the convex portion is above the mold. The convex portion can be removed by an additional step of polishing. However, we continue to build the helix bridges without removing the convex portion since the over-plated region is relatively small and it does not affect the subsequent process. A firm connection between the vias and the helix bridges is mainly determined by the via height. Successful via to bridge connections, as shown in Fig. 4.20, have been developed after several trials.

The dimensions of the successfully fabricated structures on the silicon wafer are measured using a profiler and a SEM. The thickness of the thick photoresist layer has a non-uniformity of around 5 to 6 μm across the 8-inch wafer. The vertical cylindrical vias have 6 to 9 μm variation in diameter along the height; the measured diameter is 49 μm at the bottom, 43 μm at a height of 20 μm from the bottom, and 52 μm at a height of 40 μm from the bottom. The height of the convex portion of the vias is 12 μm . Thickness of the bottom and top horizontal metal layers is 2.38 μm and 3.8 μm , respectively.

In the context of loss in high frequency microfabricated structures, surface roughness is considered to play an important role. The increase in power absorption

(a)

(b)

Figure 4.19: Scanning electron microscope (SEM) micrograph of the fabricated PH-SEC: (a) cross section view of the vertical post; (b) top view of an unsuccessful via to bridge joint.

Figure 4.20: Successful connections between the vertical posts and the suspended helical bridges.

in a transmission line due to surface roughness is defined in [133]-[134]. It is useful to characterize the additional power absorption due to the surface roughness in terms of increase in the resistivity of the material since the surface roughness cannot be considered in three-dimensional RF simulators directly [135].

As the first step towards characterizing the impact of surface roughness in our simulations, the surface roughness for the fabricated structures is measured using an atomic force microscope (AFM). Figure 4.21 shows the AFM images for the top surface of the helix bridges and the top surface of the bottom metal layer for an area of $5\text{ }\mu\text{m} \times 5\text{ }\mu\text{m}$. The root-mean-square (rms) roughness value for the bottom metal layer (based on the lift-off process) is 13.65 nm, while the rms roughness for the top bridges layer (based on the electroplating process) is 143.40 nm.

In the next step, the effective conductivity of the three metal layers is calculated, taking into account the surface roughness, following the models proposed in [134]-[135]. Let σ_1 , σ_2 and σ_3 represent the effective conductivity of the bottom horizontal metal layer, via layer, and the top horizontal metal layer (helix bridges layer), respectively. The direct current (dc) conductivity of bulk copper is

(a)

(b)

Figure 4.21: 5 μm x 5 μm atomic force microscope (AFM) images for (a) top surface of the helix bridges (vertical scale: 1500 nm/div) and (b) top surface of the bottom metal layer (vertical scale: 100 nm/div).

taken to be 5.8×10^7 S/m, corresponding to a skin depth δ_s of 0.21 μm at 100 GHz. σ_1 and σ_3 are determined based on the additional loss due to the surface roughness. According to [134], the absorption ratio of rough surface to smooth surface is given by:

$$\frac{P_{a,rough}}{P_{a,smooth}} = 1 + \frac{2}{\pi} \tan^{-1} \left[1.4 \left(\frac{h}{\delta_s} \right)^2 \right] \quad (4.5)$$

where h is the RMS height of the rough surface profile. In our case, the increase in the resistivity of the bottom metal layer due to the surface roughness of 13.65 nm is ≈ 1.004 based on (4.5); this corresponds to an effective conductivity of $\sigma_1 \approx 5.75 \times 10^7$ S/m for this layer. For the helix bridges layer, with a surface roughness of 143.40 nm, significantly more power absorption can occur due to the additional effect of oxidation of the rough electroplated copper surface. According to [135], the increase in resistivity for protrusions due to foreign objects is

$$R_{\max}(TE) = \left(\frac{4h}{\delta_s} \right) f_{bump} \quad (4.6)$$

where f_{bump} is the fraction of the surface area covered by the rough surface. In our case, the ‘foreign objects’ correspond to oxidized copper. The surface area covered by the rough surface is taken to be 100%, leading to an effective conductivity of $\sigma_3 \approx 4 \times 10^6$ S/m for this layer. The value of effective conductivity σ_2 for the via layer is obtained by fitting the simulated S_{21} to the measured S_{21} . We find that $\sigma_2 = 3 \times 10^4$ S/m best fits the measured results for the various fabricated structures. A comparison of the measured and simulated S_{21} is shown in Fig. 4.22 for structures with different widths $2b$ and a fixed period L of 152.85 μm . In general, a good match is seen between the measured and simulated values for most of the cases covered in this figure. However, a relatively poor match is seen around 100 GHz for $2b = 270$ μm . The reason for this discrepancy lies in the fact that the π -point for this case lies close to 100 GHz. The S_{21} values vary rapidly with frequency around the π -point. Due to the minor dimensional discrepancies between the fabricated and the simulated structures, the π -point in the two cases is slightly different, leading to a relatively large difference in the S_{21} values. For the smaller values of $2b$, the π -point

Figure 4.22: Comparison of measured and simulated S_{21} for the microfabricated PH-SEC with different helix widths $2b$ for a fixed period $L = 152.85 \mu\text{m}$ (22 periods). Different conductivity values for the different metal layers are considered in the simulations.

is well above 100 GHz; therefore such an effect is not observed for these values of $2b$ in this figure.

It is difficult to explain which factors could result in such a low value for σ_2 since our test sample-set is limited. There are important factors, such as contact resistance between the vias and the helix bridges, uniformity and alignment in the process, as well as oxidation of the metal layers, that are not fully captured in the models that have been used to estimate the increase in resistivity. However, the σ_2 value obtained here is of the same order of magnitude as the conductivity of electroplated copper, $0.07 \times 10^6 - 58 \times 10^6 \text{ S/m}$ reported in [136]. By improving the microfabrication process, one can expect to achieve effective conductivity values that are closer to the value for the bulk material and hence achieve lower insertion

loss. Figure 4.22 also includes results of simulated S_{21} , obtained using dc conductivity of bulk copper as 5.8×10^7 S/m, for a width $2b$ of $190 \mu\text{m}$. Unlike the case of conventional circular helix [137], the surface roughness of the three metallization layers of the PH-SEC is different from each other. Therefore it is difficult in this case to present in a simple way the change in attenuation as a function of surface roughness for the overall structure.

Figure 4.23 shows the comparison of measured and simulated S_{11} for a helix width $2b = 190 \mu\text{m}$ and helix period $L = 152.85 \mu\text{m}$. The designed PH-SEC structure with a CPW feed has a wide -20 dB S_{11} bandwidth, covering the frequency range from 60 GHz to 95 GHz. A fairly good match is seen between the simulated and measured results. The deviation from the simulated results is mainly attributed to the dimensional errors arising during the fabrication.

Figure 4.23: Comparison of measured and simulated S_{11} for the microfabricated PH-SEC with $2b = 190 \mu\text{m}$ and $L = 152.85 \mu\text{m}$ (22 periods).

The phase velocity, as shown in Fig. 4.24, is obtained using the difference in the phase of S_{21} for two structures which are identical except for having a different number of helical turns (22 and 32). The measured results are presented for three different values of the period L : $92.85 \mu\text{m}$, $112.85 \mu\text{m}$ and $152.85 \mu\text{m}$. The simulation results in Fig. 4.24 are obtained using CST transient solver. There is a very good match between the measured and the simulated results.

Figure 4.24: Comparison of measured and simulated phase velocity for the microfabricated PH-SEC for a helix width $2b = 190 \mu\text{m}$ with different periods L .

The measured axial circuit attenuation for a width $2b$ of $190 \mu\text{m}$ and period L of $152.85 \mu\text{m}$ are shown in Fig. 4.25. These attenuation values are obtained using difference in the magnitude of S_{21} for two fabricated structures with different number of helical turns. Due to the small random fluctuation in the measured S_{21} values, the accuracy of the measured axial attenuation values is limited to $\pm 0.5 \text{ dB}$. In general, the circuit attenuation increases as frequency increases. It may be noted that the circuit attenuation of the PH-SEC on high resistivity silicon is comparable to that for the raised meander-line SWS, which is reported to be 0.7 dB/mm at W-band [37].

4.4.3 Discussion

The various simulation results presented so far for the microfabricated PH-SEC structures are based on high resistivity silicon wafers with a dc resistivity of $10,000 \Omega\text{-cm}$ (corresponding to a loss tangent of 0.00017 at 90 GHz). Figure 4.26 shows the

Figure 4.25: Measured circuit attenuation for the microfabricated PH-SEC for $2b = 190 \mu\text{m}$ with $L = 152.85 \mu\text{m}$.

impact of variation in the loss tangent of the silicon wafer on the insertion loss for the microfabricated PH-SEC with different helix widths $2b$ at 90 GHz. The high resistivity silicon wafer, the one used in the final fabrication (1,000 - 10,000 $\Omega\text{-cm}$), exhibits much lower insertion loss compared to the low cost and low resistivity silicon wafers. As an example, compared to the high resistivity silicon wafer, the insertion loss is doubled if one uses a silicon wafer with 20 $\Omega\text{-cm}$ resistivity. In addition, based on the gradient of each curve shown in Fig. 4.26, the resistivity of the silicon wafer has a greater impact on helix structures with larger widths.

The helix temperature provides an important piece of information to determine the power handling capability in a corresponding TWT. The helix temperature depends upon the construction details of an actual device and an accurate determination of the temperature is quite complicated [138]-[139]. Here we present simplified simulation results using CST thermal co-simulation, ignoring the thermal contact resistance and variation in thermal conductivity with temperature. Figure 4.27 shows the simulated helix temperature for the microfabricated PH-SEC structure, in the absence of an electron beam, as a function of dissipated RMS power per unit length at 90 GHz. Thermal conductivity values used in the simulation are:

Figure 4.26: Simulated insertion loss (S_{21}) for the microfabricated PH-SEC structures as a function of the loss tangent of the silicon wafer at 90 GHz for different helix widths $2b$ with $L = 152.85 \mu\text{m}$ (22 periods).

Figure 4.27: Simulated highest helix temperature for the microfabricated PH-SEC structures as a function of dissipated root-mean-square (RMS) power at 90 GHz for different helix widths $2b$ with $L = 152.85 \mu\text{m}$ (22 periods).

148 W/°K/m for silicon, 1.4 W/°K/m for silicon dioxide, and 401 W/°K/m for copper. Based on the conductivity values used in Fig. 4.22, the *highest* temperature of the PH-SEC is captured. The highest temperature points occur at the center of the helix bridges since these locations are farthest from the silicon substrate. It is seen that the temperature increases significantly for the helix with larger widths since larger widths entail higher insertion loss. To get an idea of how these results may compare with those for the circular helix TWTs, one may consider the results for *average* temperature vs. power dissipation of a tape helix reported in [139]; those results consider a copper-plated molybdenum tape helix held in compression inside a copper shell using three anisotropic pyrolytic boron nitride (APBN) or diamond support rods. The results for helix temperature for the PH-SEC on thick silicon appear to be comparable to those for the circular helix using diamond support rods and significantly better than those for APBN support rods.

From the point of view of dielectric breakdown, the maximum value of E_x on the surface of silicon has also been examined. For 7 W_{rms} / mm dissipated power at 90 GHz, this value is 3.51×10^4 V/cm, 3.62×10^4 V/cm and 4.06×10^4 V/cm, for $2b = 150 \mu\text{m}$, $190 \mu\text{m}$ and $270 \mu\text{m}$, respectively. These values may be compared with the breakdown field for silicon and silicon dioxide which are 3×10^5 V/cm and 1×10^6 V/cm, respectively.

4.5 Summary

In this chapter, first of all, the effects of dimensional parameters of the planar helix slow-wave structures with straight-edge connections have been studied using simulation. Planar helix structures with straight-edge connections incorporating coplanar waveguide feed have been designed and fabricated. Unshielded normal helix and unshielded normal helix with air tunnel have been fabricated using printed circuit techniques. It is demonstrated that these planar helix configurations are easy to fabricate due to straight-edge connections. It has been shown that the dispersion characteristics of the planar helix with straight-edge connections can be controlled using coplanar ground planes. The planar helix slow-wave structures with straight-

edge connections have also been designed for W-band. The microfabrication process for these structures has been established. Many planar helix structures with varying lateral dimensions have been successfully fabricated based on the proposed process. The on-wafer cold-test measurement results have been presented in detail. Effective conductivity values, incorporating the effect of surface roughness, have been obtained for different metal layers in the fabricated structures. Effects of silicon wafer resistivity have also been discussed.

The fabrication of planar helix slow-wave structures with straight-edge connections has been demonstrated in this chapter. The measured results obtained from these structures validate the analytical and simulation results described in Chapter 3.

CHAPTER 5

RECTANGULAR RING-BAR WITH STRAIGHT-EDGE CONNECTIONS (RRB-SEC) FOR HIGH POWER APPLICATION

5.1 Introduction

Although the helix slow-wave structure (SWS) has the advantage of wide bandwidth, its application has been limited to relatively low power traveling-wave tubes (TWTs). For high power applications, the beam voltage is required to be high (> 5 kV). For velocity synchronism, this corresponds to a helix with a large pitch angle. Under such conditions, (i) the coupling impedance of the fundamental forward-wave mode is reduced due to non-interacting space harmonics, and (ii) the coupling impedance of the backward-wave mode is increased. These effects may result in oscillation [6]. Chodorow and Chu [6] proposed in 1955 the cross-wound twin helix to overcome the difficulties of a single tape helix. In 1956, Birdsall and Everhart [7] proposed the circular ring-bar structure as a practical form of the cross-wound twin helix. When it comes to high power applications, as shown later on in Section 5.2, the planar helix with straight-edge connections (PH-SEC) also suffers from similar problems as the circular helix. To address these problems, we propose here a new structure, a rectangular ring-bar structure with straight-edge connections (RRB-SEC); we consider the square shape as a special case of the rectangular shape. Analogous to the circular ring-bar, we show that the new structure has the potential to enable high power operation of planar TWTs operating at millimeter-wave and higher frequencies.

In this chapter, a detailed study of the properties of the RRB-SEC structure is presented. First, the simulation results for the dispersion characteristics and coupling impedance for the forward- and backward-wave of the PH-SEC are reviewed, for both square and rectangular cross sections. Next, the simulated dispersion

characteristics and coupling impedance for the RRB-SEC are compared with those for the circular ring-bar, circular helix and PH-SEC with a view to assess the suitability of the RRB-SEC for high power applications. The effects of structure dimensions on the properties of the RRB-SEC are studied. Two RRB-SEC configurations that can be microfabricated on a silicon wafer are described. All the simulation results for phase velocity and coupling impedance are obtained using the Computer Simulation Technology (CST) eigenmode solver. A W-band RRB-SEC for the simpler of the two configurations on silicon with coplanar waveguide (CPW) feed has been designed and microfabricated. The on-wafer cold-test measured S -parameters and phase velocity, covering the 80 - 110 GHz frequency range, are also presented. The effects of surface roughness and contact resistance are accounted for in the simulations by estimating effective conductivity of different parts of the microfabricated structures.

5.2 Limitations of PH-SEC

The effects of large pitch angle and large cross section perimeter on the phase velocity and coupling impedance for the PH-SEC are examined in this section. Particular attention is paid to the coupling impedance for the forward- and backward-wave modes. Figure 5.1 shows a comparison of these parameters for different values of the inner cross section perimeter, defined in (4.2), to period ratio P_c/L . The detailed dimensions of each structure are listed in Appendix C. To facilitate comparison, the phase velocity values of the backward wave mode are also plotted in the positive quadrant. The values for the backward-wave coupling impedance are plotted on a log scale so as to cover a large range.

As shown in Fig. 5.1, for the first case, with $b/a = 1$ and $P_c/L = 6.912$, the normalized phase velocity for the forward-wave is 0.139 at 85 GHz. In the second case, the b/a is maintained at 1 and P_c/L is lowered to 3.926 in order to achieve a higher normalized phase velocity of 0.283 at 85 GHz. In the second case, the coupling impedance for the fundamental forward wave is reduced and the coupling impedance for the backward wave is increased. These effects show that the PH-SEC

Figure 5.1: Simulated (a) normalized phase velocity and (b) coupling impedance for the planar helix with straight-edge connections (PH-SEC) with cylindrical vias for the fundamental forward and backward waves for different P_c/L .

structures with a large square cross-section and a large pitch angle may be prone to oscillations. In the third case, b/a is changed to 23 but P_c and L dimensions are identical to those for the first case. In this case, the normalized forward-wave phase velocity at 85 GHz is the highest at 0.38, the coupling impedance of the forward wave is increased significantly and, below ~ 110 GHz, the coupling impedance of the backward wave component is the lowest among the three cases. At 85 GHz, the ratio of the impedance for the backward-wave and forward-wave is around 5 % and 0.02 % for the first and the third case, respectively. However, it shoots up to 40 % for the second case.

5.3 RRB-SEC

The circular ring-bar structure was proposed to ease the difficulties encountered in using the circular helix for high power operation. The results for dispersion characteristics and coupling impedance of the circular ring-bar structure have been reported extensively [6], [140]-[143]. In the early days, the fabrication of the circular ring-bar involved saw-cutting of a hollow molybdenum tube [7]. To increase the operating frequency, lithography and etching on hollow preform have been reported [144]-[145]. Expensive laser cutting of molybdenum cylinder has also been reported [112]. However, the fabrication schemes reported in the literature can only produce the ring-bar structures for operation well below 100 GHz. In addition, the circular cross section of the ring-bar structure implies operation with a circular electron beam.

The limitations of the PH-SEC with square cross-section for high power application have also been brought out in Section 5.2. In this section, a novel RRB-SEC structure, shown in Fig. 5.2 (a), is proposed. The RRB-SEC is motivated by the circular ring-bar structure, which is shown in Fig. 2.7, and is a practical form of the cross-wound twin helices. In the symmetric waveguide mode of the cross-wound twin helices, the axial electric fields of the fundamental forward wave modes in the two helices combine and result in a high value of the axial coupling impedance; the

Figure 5.2: (a) Perspective view and (b) cross sectional view of the rectangular ring-bar structure with straight-edge connections (RRB-SEC) with cylindrical via.

space harmonics carry mainly the magnetic energy and result in low coupling impedance for the backward-wave [6].

The RRB-SEC is a planar version of the circular ring-bar structure and can be suitable for sheet beam applications. As shown in Fig. 5.2 (a), RRB-SEC consists of three metal layers - the bottom horizontal layer, the top horizontal layer, and the vertical straight-edge connections (or vias) that connect the two horizontal layers. The top and bottom horizontal layers consist of conducting strips, oriented along the y -direction, with thickness ST and width SW . Adjacent conducting strips are joined by a conducting bar oriented along the z -direction, with thickness ST and width BW . The conducting bars are analogous to the crossover (overlapping) of the cross-wound twin helices. Each cylindrical via has a diameter VD and it connects the top and bottom ring pads which have a diameter RD ($> VD$). A rectangular ring is formed by the top and bottom conducting strips and the two vias at the edges of the metal strips. L is the period of the structure. The center to center distance between two adjacent rectangular rings is $L / 2$. As shown in Fig. 5.2 (b), the cross section dimensions of the RRB-SEC along the x - and y -directions are $2a$ and $2b$, respectively, similar to that for the PH-SEC. The center of the rectangular rings coincides with $(0, 0)$. For the circular ring-bar structure, referring to Fig. 2.7, the inner radius and bar width (arc length) are denoted as a_i and BW .

5.3.1 Characteristics of RRB-SEC

Figure 5.3 shows the phase velocity and coupling impedance of the circular helix, circular ring-bar and RRB-SEC, with $b/a = 1$, 23 and $1/23$. The RRB-SEC is considered with both cylindrical vias and tape vias. We assume that the metal shield, used in simulations, is sufficiently far away from the structures so that, effectively, the structures can be considered to be immersed in free space. In addition, P_c , L , SW , ST and BW are the same for all the structures. The detailed dimensions of each structure are listed in Appendix C.

For the circular and square ($b/a = 1$) ring-bar structures the phase velocity and frequency range of operation are close to each other since the period and cross section perimeter are the same for both structures. Similar to the PH-SEC, the RRB-SEC with square cross section tends to have a slightly faster phase velocity and higher coupling impedance than that for the circular cross section due to strong field coupling at the four right angle corners. For the RRB-SEC with cylindrical vias, the dimensions of the vias and ring pads are greater than that of the tape vias; hence, there is a slight difference in the phase velocity and the coupling impedance for the two cases. In addition, the phase velocity and coupling impedance for the circular and square ring-bar structure are greater than that for the circular helix.

Unlike the circular ring-bar structure, the RRB-SEC allows an additional degree of freedom with respect to the aspect ratio b/a . For $b/a = 23$, the phase velocity, coupling impedance and bandwidth are greater than that for the square RRB-SEC. This effect is similar to that for the rectangular helix and PH-SEC (see Fig. 4.2), since the RRB-SEC can be considered to be a combination of two PH-SECs. For $b/a = 1/23$, the phase velocity does not change much compared to that for the square case; however, interestingly, the coupling impedance is significantly greater than that for $b/a = 23$. The reason for this can be explained using the current flow direction in the structure. The simulated current flow for the PH-SEC and RRB-SEC is shown in Fig. 5.4. The arrows indicate the direction of current flow. In the case of RRB-SEC, for $b/a = 1/23$, the straight-edge connections (along the x -direction) in a particular ring are quite close to each other with the current flow in

Figure 5.3: Simulated (a) normalized phase velocity and (b) coupling impedance for the circular helix, circular ring-bar and RRB-SEC for $b/a = 1, 23$ and $1/23$ with fixed P_c and L .

Figure 5.4: Simulated current flow for (a) PH-SEC and (b) RRB-SEC.

the same direction. This effect does not occur in the case of the PH-SEC with $b/a = 1/23$; for the PH-SEC the current flow in the straight-edge connections (along the x -direction) in a particular turn is in opposite directions and therefore, unlike the RRB-SEC, there is no build up of a strong axial electric field.

Figure 5.5 shows the simulated axial component of the electric field (E_z) plotted in the x - y plane at different phase values for RRB-SEC for $b/a = 1$. The small black box at the bottom of each graph indicates that the fields are captured at the midpoint of the bottom conducting bar along the z -direction. Unlike the PH-SEC, the strongest E_z field for the RRB-SEC moves upward and downward. The E_z

Figure 5.5: Simulated axial component of the electric field E_z plotted in x - y plane at different phase for RRB-SEC for $b/a = 1$.

field strength is strongest near the conductor and reduces towards the center of the structure ($x = y = 0$).

Figure 5.6 shows the phase velocity and coupling impedance for the fundamental forward and backward waves for the RRB-SEC with cylindrical vias, with $b/a = 1$ and 23. Two results for the square PH-SEC with cylindrical vias, with $P_c/L = 3.14$ and 2.511, are also plotted in this figure. The dimensions of the PH-SEC are selected so that the phase velocity values of the fundamental forward wave of the PH-SEC and the RRB-SEC coincide at $k_0 a_{eq} = 0.3$. The ratio of the backward- to-forward impedance for both cases of the PH-SEC is $\sim 19\%$ at $k_0 a_{eq} = 0.3$. For the RRB-SEC, this ratio is much lower, being 0.84% and 0.22% for $b/a = 1$ and 23, respectively.

5.3.2 Effects of Dimensional Parameters of RRB-SEC

As the frequency increases, the effects of even small changes in the dimensions may be significant. In this sub-section, we study the effect of variations in L , ST , SW , BW , and VD (together with RD) on the phase velocity and coupling impedance for the RRB-SEC with cylindrical vias, in free space. The detailed dimensions of each structure are listed in Appendix C. In general, the nature of these effects is similar to that reported for the circular ring-bar structure [7] and [142].

Figure 5.7 shows the phase velocity and coupling impedance for the RRB-SEC with different values of the period L , for $b/a = 1$ and 23. For a given perimeter, the period mainly controls the phase velocity. As the period increases, the phase velocity increases and the bandwidth reduces. The coupling impedance may increase or decrease as the period increases depending on whether the frequency range is below or above a certain ‘cross-over frequency’. A similar effect has also been observed in [142] for the circular ring-bar structure. The cross-over frequency is shifted to the band-edge for $b/a = 23$.

The effects of different values of the bar width BW on the phase velocity and coupling impedance for $b/a = 1$ and 23 are shown in Fig. 5.8. As the bar width increases, the current can have a shorter path to move from one ring to the next,

(a)

(b)

Figure 5.6: Simulated (a) normalized phase velocity and (b) coupling impedance for the RRB-SEC with cylindrical vias for the fundamental forward and backward waves for $b/a = 1$ and 23. Results for the PH-SEC with cylindrical vias, with $P_c/L = 3.14$ and 2.511 for $b/a = 1$, are also included.

having the same effect as reducing the perimeter for a given period. As a result, the phase velocity and bandwidth increase as the bar width increases. A similar effect for the circular ring-bar structure was reported in [7]. Similar to the effect of the period in the previous figure, the coupling impedance may increase or decrease as the bar width increases, depending on whether the frequency range is below or above a certain ‘cross-over frequency’.

The effects of varying the strip thickness ST , strip width SW , and via-diameter VD on the phase velocity and coupling impedance have also been examined. Figure 5.9 shows that the phase velocity for thicker metal strips is higher at low frequencies and becomes lower above a certain cross-over frequency. The coupling impedance of the fundamental forward wave decreases for thicker metal strips due to an increase in the stored energy [7]. The phase velocity and bandwidth decrease as the strip width decreases (figure not included). The effect of change of via diameter has also been studied (figure not included). For $b/a = 1$, the effect of variation of via diameter is similar to that of varying the strip thickness. For $b/a = 23$, the variation of coupling impedance does not show a clear pattern. An important observation is that the highest value of the coupling impedance occurs when the via-diameter matches the strip width.

5.4 Configurations for Microfabrication of the RRB-SEC on a Silicon Substrate

The non-uniform field distribution of the PH-SEC on a thick silicon substrate has been pointed out in Chapter 4. Such an asymmetric field distribution may cause gain loss if it is used for actual tube operation. To avoid such problems, it is desirable to reduce the asymmetry as much as possible. One possible way to achieve this is by reducing the amount of silicon beneath the structure. The cross-section views of the RRB-SEC with cylindrical vias, on a thick silicon substrate and on a thin silicon substrate with trench, are shown in Fig. 5.10 (a) and (b), respectively. The configuration in Fig. 5.10 (b) reduces the field asymmetry; of course this configuration requires additional silicon etching steps. Silicon has a dielectric

Figure 5.7: Simulated (a) normalized phase velocity and (b) coupling impedance for the RRB-SEC with cylindrical vias, with different values of the period L for $b/a = 1$ and 23.

Figure 5.8: Simulated (a) normalized phase velocity and (b) coupling impedance for the RRB-SEC with cylindrical vias, with different values of the bar width BW for $b/a = 1$ and 23.

Figure 5.9: Simulated (a) normalized phase velocity and (b) coupling impedance for the RRB-SEC with cylindrical vias, with different values of the metal strip thickness ST for $b/a = 1$ and 23.

Figure 5.10: Cross-sectional view of the RRB-SEC with cylindrical vias on (a) a thick silicon substrate and (b) a thin silicon substrate with a trench.

constant of 11.9. The metal shield, used in simulations, is not shown in this figure since the shield is assumed to be placed sufficiently far away from the structure. In addition, for simplicity, the effect of a thin layer of silicon dioxide (SiO_2), usually deposited on a silicon wafer for enhancing the electrical insulation, is ignored.

The phase velocity and coupling impedance for both RRB-SEC configurations mentioned above, with cylindrical vias and $b/a = 1$ and 23, are shown in Figs. 5.11 and 5.12, respectively. Both structures are assumed to have an identical P/L ratio; h_1 for the thick silicon case is $350 \mu\text{m}$; h_1 and h_2 for the case of trench in silicon are $50 \mu\text{m}$ and $220 \mu\text{m}$, respectively. Other dimensions of each structure can be found in Appendix C. As expected, due to dielectric loading, the bandwidth and phase velocity for the thick silicon configuration is lower than that for the trench in silicon. The coupling impedance values at $x = -a/2, 0$ and $a/2, y = 0$ (see Fig. 5.10), for $b/a = 1$ and 23 are plotted in Figs. 5.12 (a) and (b), respectively. For the

Figure 5.11: Simulated normalized phase velocity for the RRB-SEC with cylindrical vias for $b/a = 1$ and $b/a = 23$ for the silicon configurations.

thick silicon configuration, the variation in the coupling impedance along the x -axis is asymmetric, with generally higher values for the negative values of x for both aspect ratios. On the other hand, for the trench in silicon configuration, the coupling strengths at $\pm a / 2$ are relatively more uniform. The uniformity is much better for the high aspect ratio structure ($b/a = 23$) since the silicon is far away from the center of the structure; in addition, the coupling impedance at the centre of the structure also increases compared to the thick silicon case. Hence, the trench in silicon solves the problem of non-uniform field distribution inside the RRB-SEC while maintaining sufficient contact with the RRB-SEC for thermal conduction.

5.5 Microfabrication and Measurement Results of W-band RRB-SEC

In this section, we present the design, microfabrication, and measurement of the RRB-SEC with CPW feed at W-band. For this proof-of-concept fabrication, we

Figure 5.12: Simulated coupling impedance for the RRB-SEC with cylindrical vias for (a) $b/a = 1$ and (b) $b/a = 23$ for the silicon configurations.

choose the configuration with a thick silicon substrate, as shown in Fig. 5.10 (a), with the standard thickness of 750 μm , due to a simpler microfabrication process. The fabrication process steps, as shown in Fig. 5.13, are the same as those for the PH-SEC described in Chapter 4. The capability of building high density or high

Figure 5.13: Fabrication process steps for the RRB-SEC with cylindrical vias: (a) bottom copper (Cu) lift-off, (b) first seed layer deposition followed by patterning and electroplating of Cu vias, (c) second seed layer deposition followed by patterning and electroplating of Cu helix bridges.

aspect ratio vias is once again verified in this experiment. Several RRB-SEC structures are fabricated on an 8-inch high-resistivity silicon wafer. The dimensions of the fabricated structures are listed in Appendix C.

The RRB-SEC with CPW feed is designed using the CST Microwave Studio. A CPW section starts from the bottom of the first rectangular ring, as shown in Fig. 5.13 (a), followed by a tapered section, with a taper length of 595 μm , to join a 50 Ω section. To ensure the excitation of the symmetric mode, the signal line of the CPW with width 40 μm is centered at $y = 0$ (see Fig. 5.10). The edge of the CPW ground planes is 31 μm away from the edge of the vias along the z -direction. The length of the tapered CPW section and the impedance of the initial CPW section are important parameters to achieve good impedance matching.

Briefly, a 3 μm SiO_2 layer is deposited on top of the high resistivity silicon wafer. The bottom metal layer and CPW feeds, as shown in Fig. 5.13 (a), are formed using the negative tone photoresist MaN-1440 lift-off process. Figure 5.13 (b) shows the vias formed by lithography of a thick positive tone photoresist layer (AZ 40XT-11D) and subsequent electroplating. A layer of negative tone dry film photoresist, as shown in Fig. 5.13 (c), is laminated and patterned to form the electroplating mold for the suspended rectangular ring bridges. This process can be easily scaled to fabricate an RRB-SEC with a higher aspect ratio and higher frequency.

The dimensions of the successfully microfabricated structures are measured using a profiler and a scanning electron microscope (SEM). The SEM micrographs of the successfully fabricated RRB-SEC are shown in Fig. 5.14. Thickness of the bottom and top horizontal metal layers is 2.38 μm and 3.8 μm , respectively. The surface roughness for the fabricated structures is measured using an atomic force microscope (AFM). The root-mean-square (rms) roughness value for the bottom metal layer (based on the lift-off process) is 13.65 nm, while the rms roughness for the top bridges layer (based on the electroplating process) is 143.40 nm. Details of the other dimensions are the same as those for W-band PH-SEC as described in Chapter 4. Considering that the skin depth for bulk copper at 100 GHz is 210 nm, the additional loss due to the surface roughness for the bottom metal layer is very small. However, the surface roughness of the top bridges layer is significant in terms of skin depth and can cause a considerable increase in the conductor loss.

(a)

(b)

Figure 5.14: Scanning electron microscope (SEM) micrographs of the microfabricated RRB-SEC structure. (a) Bird's-eye view showing the RRB-SEC, with cylindrical via, with coplanar waveguide (CPW) feed. (b) Side view of the electroplated vias showing the convex profile at the top.

Figure 5.15: Measured and simulated S -parameters for the microfabricated RRB-SEC with cylindrical vias for $2b = 150\mu\text{m}$. Other dimensions of the structure can be found in Appendix C. The surface roughness and contact resistance have been accounted for as effective conductivity for each metallization layers during simulation. Measured S_{21} of PH-SEC with identical cross section perimeter is included for comparison.

Figure 5.15 shows a comparison of the measured and simulated S -parameters for the microfabricated RRB-SEC structures. In simulations, the effective conductivity of the three metal layers is modified to take into account the surface roughness, following the models proposed in Chapter 4. The effective conductivities of the bottom horizontal metal layer, via layer and the top horizontal metal layer, are $5.75 \times 10^7 \text{ S/m}$, $3 \times 10^4 \text{ S/m}$ and $4 \times 10^6 \text{ S/m}$, respectively. The results demonstrate that the RRB-SEC structure with a CPW feed has a wide -20 dB S_{11} bandwidth, covering the frequency range from 80 GHz to 110 GHz. The simulated S_{21} results agree well with the measured result since the effect of surface roughness and contact resistance are considered in the simulation by using modified conductivity as described in Section 4.4.2 for the PH-SEC. The deviation from the measured results is mainly attributed to the dimensional errors arising during the fabrication. The measured S_{21} for a 22-turn PH-SEC with an identical cross section perimeter, but a

Figure 5.16: Comparison of measured and simulated phase velocity for the microfabricated RRB-SEC with cylindrical vias for $2b = 270\mu\text{m}$. Other dimensions of the structure can be found in Appendix C.

different period L ($152.85\ \mu\text{m}$), is also included. It is seen that the RRB-SEC has an insertion loss quite similar to that for the PH-SEC.

The measured phase velocity, as shown in Fig. 5.16, is obtained using the difference in the phase of S_{21} for two structures which are identical except for having a different number of helical turns (22 and 32). The simulation results in Fig. 5.16 are obtained using the CST transient solver. The measured and the simulated results match within $\sim 8\%$.

5.6 Summary

This chapter begins by showing that the ratio of the coupling impedance of the backward- and forward-wave can be rather high for the planar helix slow-wave structure with straight-edge connections with a large pitch angle and large cross sectional dimensions. To solve this problem, a new planar structure, rectangular ring-bar with straight-edge connections, has been proposed and investigated in

detail. It has been demonstrated that the dispersion and coupling impedance properties of the rectangular ring-bar with straight-edge connections are very similar to those for the circular ring-bar structure. Thus, the ratio of the coupling impedance for the backward- and forward-wave for the rectangular ring-bar with straight-edge connections is significantly less than that for the corresponding circular tape helix and the planar helix with straight-edge connections. The effects of dimensional parameters for the rectangular ring-bar with straight-edge connections on the dispersion characteristics and coupling impedance have been studied. Two configurations for the rectangular ring-bar with straight-edge connections which can be microfabricated on a silicon substrate have also been proposed. The simpler of these two configurations has been designed for operation at W-band and fabricated on a high-resistivity silicon wafer. For the fabricated structures, measured S -parameters and phase velocity match well the simulation results.

CHAPTER 6

GAIN AND OUTPUT POWER FOR W-BAND PLANAR HELICAL SLOW-WAVE STRUCTURES BASED ON STRAIGHT-EDGE CONNECTIONS

6.1 Introduction

The microfabrication of a W-band planar helix slow-wave structure (SWS) with straight-edge connections (PH-SEC) on a silicon substrate using a UV LIGA process has been described in Chapter 4. The PH-SEC on a thick silicon substrate is a relatively simple structure to fabricate. However, in this case, due to the thick silicon substrate, the variation in the coupling impedance perpendicular to the substrate is asymmetric, with generally higher values closer to the substrate. Such an asymmetric field distribution may reduce the efficiency and gain for operation in a traveling-wave tube (TWT). One possible way to reduce the asymmetry is to reduce the amount of silicon beneath the SWS. Hence a rectangular ring-bar with straight-edge connections (RRB-SEC) on a thin silicon substrate with a trench in silicon has been proposed in Chapter 5.

In this chapter, hot-test parameters, namely, power, gain, and efficiency for a square PH-SEC and a square RRB-SEC, both on a thin silicon substrate, i.e., with a trench in silicon beneath the structure, as shown in Fig. 6.1, are studied for TWT amplifier application at W-band (75 - 110 GHz). For this configuration, 3D particle-in-cell (PIC) simulation results obtained using the Computer Simulation Technology (CST) PIC solver are presented. The cold-test parameters of the structure are first optimized using the CST eigenmode solver. The input-output couplers using simplified coplanar waveguide (CPW) are designed using the CST Microwave Studio (MWS). The linear and non-linear amplification of the TWT are examined

Figure 6.1: Dimensions of the planar helix with straight-edge connections (PH-SEC) and rectangular ring-bar with straight-edge connections (RRB-SEC) on a silicon substrate with a trench in silicon. All dimensions are in micrometers.

next; the spectral purity and possibility of oscillations in the hot structure are also considered. A rectangular waveguide coupler is also designed for higher power applications. The hot-test parameters are compared for the simplified CPW and the rectangular waveguide couplers. The results for RRB-SEC are also compared with the results of PH-SEC.

6.2 Simulation Model for Planar Helix with Straight-Edge Connections

Figure 6.1 shows the dimensions of the PH-SEC on a silicon substrate with a trench in silicon. The PH-SEC is designed for an aspect ratio of 1 in this simulation model. A layer of thin silicon oxide is assumed on top of silicon for better isolation. A metal shielding is considered in the simulations. The dimensions of the metal shield are chosen such that the cut-off frequency of the first higher order mode is beyond 110 GHz. All metal, including the metal shield, is copper with the bulk conductivity of 5.8×10^7 S/m. Figure 6.2 (a) shows the dispersion curves for the fundamental mode

(a)

(b)

Figure 6.2: (a) Dispersion curves of the fundamental mode and the two higher order modes. 5.8 kV beam line is also included. (b) Coupling impedance of the fundamental mode.

1 and two higher order modes 2 and 3, obtained from the eigenmode solver, together with the 5.8 kV beam line. The beam line intersects with modes 1, 2, and 3 at 85 GHz, 139 GHz and 235 GHz, respectively, showing possible beam-wave interaction at these frequencies. The coupling impedance of the forward fundamental mode varies between 15 Ω to 85 Ω at W-band, as shown in Fig. 6.2 (b).

The structure is then designed with input-output couplers using the CST Microwave Studio. The following subsections describe the designs of input-output couplers using the simplified CPW and the rectangular waveguide.

6.2.1 Simplified Coplanar Waveguide (CPW) Coupler

The PH-SEC structure incorporating a CPW coupler has been introduced in Chapter 4. The CPW starts with a high impedance section from the helix followed by a tapered section to join a standard 50 Ω section. To obtain a low return loss (< -20 dB) for a wide frequency range, the impedance of the CPW section near the helix and the length of the tapered section needs to be optimized. Hence, we propose here a simplified CPW coupler that can be tuned systematically to obtain a low return loss over a wide frequency range.

As shown in Fig. 6.3, a ground strip is extended from the metal shield, with a distance s from the edge of the ring pad. Discrete ports are used to define the impedance Z_0 between the ground plane and the PH-SEC. Figure 6.4 shows the plots of S_{11} for three sets of values for s and Z_0 . By varying s and the corresponding Z_0 values, a frequency range of low return loss (< -20 dB) can be tuned over the entire band. Although such a simplified coupler cannot be realized in practice, the hot-test parameters can be obtained readily for a wide frequency range using the CST PIC solver.

6.2.1.1 Particle-In-Cell (PIC) Simulation

The PH-SEC structure designed with simplified CPW couplers is then simulated in

Figure 6.3: Perspective view of the PH-SEC incorporating a simplified coplanar waveguide (CPW) coupler.

Figure 6.4: Return loss for different values of s (ground to ring pad distance in Fig. 6.3) and the corresponding discrete port impedance Z_0 .

the CST PIC solver. In order to obtain the non-linear behavior of the TWT and avoid oscillations due to high gain, an idealized sever is included in the PIC solver. The idealized sever consists of two simplified CPW couplers with a small separation between them, and is placed 58.5 periods away from the input port and 137.5 periods away from the output port. The simplified CPW couplers are used to absorb the forward and the reflected signal at the sever location. The design goal is to achieve a peak output power of 30 dBm (1 W) and a 15 % 3-dB saturated gain bandwidth at W-band. A cylindrical electron beam is used to interact with the square PH-SEC. As depicted in Fig. 6.3, the cylindrical electron beam is centered at the origin. A beam voltage of 5.8 kV and current of 8 mA are selected after several trials. The beam filling factor is 50 % and the current density is 70.7 A/cm^2 . The beam propagates in the $+z$ direction under a homogeneous axial magnetic field of 0.5 T. The dielectric and conductor losses, with bulk resistivity, are considered in the simulation. The hot-test parameters for a particular frequency are obtained by injecting an input signal at that frequency.

6.2.1.2 Results and Discussion

Figure 6.5 shows the particle trajectory plot near the end of the structure at 97.5 GHz. The colors indicate alternating high and low energy electrons due to velocity

Figure 6.5: Particles trajectory plot captured at 97.5 GHz.

Figure 6.6: Input and output time signals at 97.5 GHz.

modulation and bunching. Figure 6.6 shows the input and output time signals at 97.5 GHz. The input signal is a single-frequency sinusoidal function with an input power of 17 dBm (50 mW). Thus, the amplitude of the input signal is 0.2236 with units of square root of power. The output signal saturates at 1.2 ns with an amplitude of 1.39 corresponding to a gain of 16 dB. The input and output powers calculated in this manner refer to the peak values. The phase space plot of the bunched electron beam along the propagation direction for the saturated output at 7.5 ns at 97.5 GHz is shown in Fig. 6.7. The average kinetic energy of the electrons is reduced towards the end of the interaction region indicating that the energy of electrons is transferred to the electromagnetic wave.

Figure 6.8 shows the simulated linear gain, saturated gain, and saturated output peak power vs. frequency curves for the PH-SEC structure shown in Fig. 6.3. The peak linear gain is 24.9 dB at 97.5 GHz. The saturated output peak power varies from 28 dBm (0.63 W) to 34 dBm (2.51 W) over the frequency range of 85 – 110 GHz. The 3-dB bandwidth of the saturated gain is about 15 %.

Figure 6.9 shows the output power and the electronic efficiency vs. input power curves at 97.5 GHz. The output power for both lossy and lossless circuits is included. The saturated output power for the lossy circuit is 33.24 dBm (2.11 W)

Figure 6.7: Phase space plot of the bunched electron beam along the propagation direction at 97.5 GHz for PH-SEC shown in Fig. 6.3.

Figure 6.8: Linear gain, saturated gain, and saturated output peak power vs. frequency for the PH-SEC structure shown in Fig. 6.3.

Figure 6.9: Output power and electronic efficiency vs. input power at 97.5 GHz for the PH-SEC structure shown in Fig. 6.3.

with the saturated gain of 19.26 dB. Since a collector is not designed in this case, the electronic efficiency η_e can be calculated as follow [55]

$$\eta_e = \frac{P_{out} - P_{in}}{V_{beam} \times I_{beam}} \times \frac{1}{\eta_{cir}} \quad (6.1)$$

where P_{in} and P_{out} correspond to the radiofrequency (RF) input and output power in watts; V_{beam} and I_{beam} correspond to the voltage and current for the electron beam; η_{cir} corresponds to the circuit efficiency. The maximum electronic efficiency is about 7.3 %. This is significantly better than the electronic efficiency of 4.5 % for a folded waveguide SWS [115]. The efficiency can be improved further by a proper design of the collector and the SWS near the output end.

The frequency spectrum for the input, output, and reflected signals for $P_{in} = 17$ dBm (saturated input power) at 97.5 GHz are shown in Fig. 6.10. These spectra are obtained by performing Discrete Fourier Transform (DFT) on the time signals using the CST post processing tool. As shown in Fig. 6.10, the peaks at 195 GHz in the output and the reflected signals correspond to the 2nd harmonic; the power level of the 2nd harmonic is significantly lower than that for the fundamental. A similar level of the 2nd harmonic has also been observed in [146]. Since the PH-SEC with a

Figure 6.10: Frequency spectrum of the input, output and reflected signals at $P_{in} = 17$ dBm (saturated input power) at 97.5 GHz for the PH-SEC structure shown in Fig. 6.3.

metal shield is a two-conductor structure, and the simplified CPW coupler together with the discrete port also supports direct current (dc), there is a peak output signal near dc but the corresponding power level is quite small. The oscillations at the fundamental frequency are unlikely to occur since the reflected power is well below the input power level. There is non-significant backward wave growth corresponding to several frequencies where the beam line crosses the higher order backward modes.

6.2.2 Rectangular Waveguide Coupler

In the previous sub-section (6.2.1), the saturated output power, linear and non-linear gain, and electronic efficiency are obtained using the simplified CPW couplers. Due to the narrow spacing between the CPW ground planes and the signal line, the CPW coupler has limited power handling capability. Hence, a rectangular waveguide coupler, as shown in Fig. 6.11, is designed for high power handling. Figure 6.11 (a) shows the PH-SEC with metal shield incorporating rectangular waveguide couplers.

(a)

(b)

Figure 6.11: (a) Perspective view of shielded PH-SEC with rectangular waveguide coupler. (b) Dimensions of the electric field probe (dimensions in micrometers).

The top cover of the rectangular waveguide and metal shielding is omitted in the figure. The rectangular waveguide has dimensions identical to those of the standard WR-10 ($1270 \times 2540 \mu\text{m}^2$) waveguide. The metal shield includes grooves to hold the SWS. Figure 6.11 (b) shows the dimensions of the electric field probe inside the WR-10 waveguide. As shown in Fig. 6.11 (a), an idealized sever, containing two WR-10 waveguide ports with electric field probes, is placed 58.5 periods away from

the input port (port 1) and 137.5 periods away from the output port (port 4). There are four waveguide ports defined in this TWT. Ports 2 and 3 are used to absorb the forward signal and reflected signal, respectively.

6.2.2.1 Results and Discussion

The rectangular waveguide couplers are designed using the CST MWS. The simulated return loss (S_{11}) at port 1 and insertion loss (S_{21}) from port 1 to port 2, for the structure shown in Fig 6.11, in the absence of PH-SEC and in the presence of PH-SEC, are shown in Fig. 6.12. When the PH-SEC is absent, as shown in Fig. 6.12 (a), there is no transmission from port 1 to port 2 and the input power is completely reflected back to port 1. Figure 6.12 (b) shows a good transmission from port 1 to port 2 with -5 to -6 dB insertion loss for 58.5 periods of PH-SEC and -20 dB return loss over the 85 to 108 GHz frequency range.

With the designed WR-10 couplers, the PH-SEC is simulated in the CST PIC solver with the beam settings following the description in Section 6.2.1.1. Figure 6.13 shows the saturated output power obtained for the PH-SEC using the rectangular waveguide couplers. The saturated output power is obtained for the frequency range where $S_{11} < -20$ dB for the case of the rectangular waveguide couplers. For comparison, results for the maximum saturated output power are similar for both types of couplers. However, the output power response is shifted towards lower frequencies for the rectangular waveguide couplers.

The frequency spectrum for the input, output, and reflected signals at $P_{in} = 14$ dBm (saturated input power) at 97.5 GHz are shown in Fig. 6.14. Similar to the case of the PH-SEC with simplified CPW couplers, there are small peaks at 195 GHz in the output and reflected signals. However, as compared to the case of the simplified CPW couplers (Fig. 6.10), the power level at these peaks is reduced in this case. The oscillations are unlikely to occur since the reflected power at 195 GHz is well below the input power level. Unlike the structure with the simplified CPW couplers, the structure with rectangular waveguide couplers is a single-conductor structure. Therefore, the spectrum does not show peaks near dc for both the output and the

Figure 6.12: Simulated return loss at port 1 (S_{11}) and insertion loss between port 1 and port 2 (S_{21}), for the structure shown in Fig 6.11, (a) in the absence of PH-SEC and (b) in the presence of PH-SEC.

Figure 6.13: Simulated saturated output power vs. frequency for the PH-SEC with rectangular waveguide couplers shown in Fig. 6.11. Results for the structure with simplified CPW couplers are also included.

Figure 6.14: Frequency spectrum of the input, output and reflected signals at $P_{in} = 14$ dBm (saturated input power) at 97.5 GHz for the PH-SEC with rectangular waveguide couplers shown in Fig. 6.11.

reflected signals. The spectrum is also very clean compared to the case of the simplified CPW couplers. This indicates that the idealized sever using the rectangular waveguide couplers provides a better isolation between the two sections separated by the sever.

6.3 Simulation Model for Rectangular Ring-Bar with Straight-Edge Connections

The RRB-SEC on a silicon substrate with a trench in silicon, as shown in Fig. 6.1, has been designed for obtaining its hot-test parameters. The dimensions of the RRB-SEC are similar to those of the PH-SEC. Figure 6.15 (a) shows the dispersion curves for the fundamental mode 1 and two higher order modes 2 and 3. To operate the RRB-SEC at W-band frequency, a beam voltage of 19.7 kV is chosen for interacting with mode 1. The beam line intersects with modes 1, 2, and 3 at 90 GHz, 150 GHz, and 235 GHz, respectively, showing possible beam-wave interaction at these frequencies. The dispersion at the intersection point for mode 1 is large compared to that for the case of the PH-SEC (see Fig. 6.2 (a)), indicating that the synchronization of the velocity only occurs for a small frequency range. The comparison of coupling impedance for mode 1 for the RRB-SEC and the PH-SEC is shown in Fig. 6.15 (b). The coupling impedance for the RRB-SEC is significantly larger than that of the PH-SEC at W-band.

The RRB-SEC is designed with the input-output couplers and sever using WR-10 rectangular waveguides similar to those of the PH-SEC as described in Section 6.2.2. The total physical length of the RRB-SEC is kept identical to that of the PH-SEC (see Fig. 6.11 (a)). The sever, containing two WR-10 waveguide ports (Port 2 and 3), with electric field probe dimensions shown in Fig. 6.16, is placed 58.5 periods away from the input port (Port 1) and 137.5 periods away from the output port (Port 4). As depicted in Fig. 6.16, a cylindrical electron beam, centered at the origin, is used to interact with the square RRB-SEC. Since the RRB-SEC has more stable forward wave interaction, a higher peak output power of 43 dBm (20 W) is targeted. A beam current of 20 mA (2.5 times higher than the beam current for

(a)

(b)

Figure 6.15: (a) Dispersion curves of the fundamental mode and two higher order modes for RRB-SEC shown in Fig. 6.1. 19.7 kV beam line is also included. (b) Comparison of coupling impedance of the fundamental mode for the PH-SEC and RRB-SEC.

Figure 6.16: Dimensions of the RRB-SEC and electric field probe (dimensions in micrometers).

the PH-SEC), with current density of 177 A/cm^2 , is selected. Similar to the case of PH-SEC, the beam filling factor is 50 % and the beam is assumed to be under a homogeneous axial magnetic field of 0.5 T.

6.3.1 Results and Discussion

Figure 6.17 shows the return loss at port 1 and insertion loss between port 1 and port 2 for the RRB-SEC structure shown in Fig. 6.16. For a structure length of 58.5 periods of the RRB-SEC, the -20 dB return loss is achieved over 88 to 105 GHz frequency range, with the corresponding insertion loss varying from -4 to -7.5 dB. Figure 6.18 shows the saturated output peak power and gain vs. frequency curves for the RRB-SEC structure shown in Fig. 6.16. The saturated output peak power varies from 40 dBm (9.1 W) to 44.6 dBm (28.6 W) over the frequency range of 90 - 98 GHz. The saturated peak power for the RRB-SEC is greater than that for the PH-SEC (see Fig. 6.13) since the beam current for the RRB-SEC is higher than that for the PH-SEC. However, the 3-dB bandwidth of the saturated gain is about 6 % for the RRB-SEC which is smaller than that for the PH-SEC. The phase space plot for the bunched electron beam along the propagation direction for the saturated output

Figure 6.17: Simulated return loss at port 1 (S_{11}) and insertion loss between port 1 and port 2 (S_{21}), for the RRB-SEC structure shown in Fig. 6.16.

Figure 6.18: Saturated output peak power and gain vs. frequency for the RRB-SEC structure shown in Fig. 6.16.

Figure 6.19: Phase space plot of the bunched electron beam along the propagation direction at 93.75 GHz for RRB-SEC shown in Fig. 6.16.

at 9 ns for the RRB-SEC at 93.75 GHz is shown in Fig. 6.19. It can be seen that the average kinetic energy of the electrons is reduced towards the end of the interaction region.

The electronic efficiency vs. frequency for the RRB-SEC is shown in Fig. 6.20. The electronic efficiency is calculated based on equation (6.1). The maximum electronic efficiency for the RRB-SEC is 15.7 % at 95 GHz. This is significantly higher than the electronic efficiency of 7.3 % at 97.5 GHz for the PH-SEC. The frequency spectrum for the input, output, and reflected signals at $P_{in} = 20$ dBm (saturated input power) at 93.75 GHz is shown in Fig. 6.21. There exist small peaks at the 2nd harmonic 187.5 GHz in the output and reflected signals. However, the reflected power at 2nd harmonic is well below the input power level hence it is unlikely to cause oscillations. The spectrum is very clean since the rectangular waveguide couplers and a sever are used in this design.

Figure 6.20: Electronic efficiency vs. frequency of the RRB-SEC shown in Fig. 6.16.

Figure 6.21: Frequency spectrum of the input, output and reflected signals at $P_{in} = 20$ dBm (saturated input power) at 93.75 GHz for the RRB-SEC shown in Fig. 6.16.

6.4 Summary

In this chapter, first a square cross section planar helix with straight-edge connections has been simulated with a cylindrical electron beam at W-band using the CST, Particle Studio, Particle-In-Cell solver. The linear and non-linear amplification of the input signal has been examined. A simplified coplanar waveguide coupler is proposed to facilitate the simulation of hot-test parameters over a wide frequency range. The saturated output power varies from 0.65 W to 2.58 W over the frequency range 85 - 110 GHz. The maximum electronic efficiency is 7.3 % at 97.5GHz for a beam voltage of 5.8 kV and beam current of 8 mA. Design of a practical coupler using the W-band rectangular waveguide is also described. Compared to the case of the PH-SEC with simplified CPW couplers, the PH-SEC with rectangular waveguide couplers provides a cleaner output spectrum. Next, a square cross section rectangular ring-bar with straight-edge connections incorporating the rectangular waveguide couplers has been designed and simulated with a relatively higher current density cylindrical electron beam at W-band. With a beam voltage of 19.7 kV and beam current of 20 mA, a saturated output peak power varying from 9.1 W to 28.6 W over the frequency range 90 - 98 GHz has been obtained. The maximum electronic efficiency is 15.7 % at 95GHz. A design of a stable high output power traveling-wave tube has been achieved using the rectangular ring-bar with straight-edge connections.

CHAPTER 7

CONCLUSION AND RECOMMENDATIONS

7.1 Conclusion

Two new planar helical slow-wave structures (SWSs) based on straight-edge connections are proposed and investigated in this thesis. The proposed planar helical SWSs, planar helix with straight-edge connections (PH-SEC) and rectangular ring-bar with straight-edge connections (RRB-SEC), are the planar analogs of the circular helix and the circular ring-bar structures, respectively. These planar analogs retain the key advantages of their circular counterparts. For instance, the PH-SEC exhibits a broad bandwidth similar to that for the circular helix. Significantly, the straight-edge connections can be easily fabricated using printed circuit techniques or micro-fabrication techniques, leading to the possibility of low cost production of traveling-wave tubes (TWTs). To fabricate SWSs for high frequency applications, use of microfabrication techniques is mandatory. Moreover, by changing the aspect ratio, the proposed structures can be made suitable for sheet beam applications which offer many advantages for high frequency TWTs.

This thesis describes in detail the properties of the two new planar SWSs in the context of their application in TWT amplifier. The studies include the guiding characteristics, fabrication methods, as well as the performance of a TWT amplifier using one of these structures. Important conclusions of these studies are mentioned in the following.

The PH-SEC can offer a phase velocity which is always lower than that for a corresponding rectangular helix. To gain a better insight into the waveguiding properties of the structure, the PH-SEC has been investigated with the help of analysis and simulations. The dispersion characteristics of the proposed structure immersed in free space have been studied. The effective dielectric constant (EDC) method has been applied to obtain the dispersion characteristics. The EDC method simplifies the 3D field analysis problem into two related 2D field analysis problems.

It has been shown that the dispersion characteristics approximated using the EDC method agree well with the simulation results in the far-from-cutoff region.

As the next step, several possible configurations, which take into account practical modifications such as a vacuum tunnel, metal shield and multilayer dielectric substrates, have been investigated. A modified effective dielectric constant (MEDC) method has been proposed and applied to the PH-SEC in the presence of multilayer dielectric substrates. The MEDC method preserves the simplicity of the basic EDC concept while improving the accuracy of the approximated dispersion characteristics. The coupling impedance of the PH-SEC in the presence of multilayer dielectric substrates has also been estimated using the 2D planar helix approximation. The phase velocity and coupling impedance results calculated using the proposed methods agree well with the full 3D simulation results in the far-from-cutoff frequency range. It has been shown that the shielded inverted helix and shielded helix configurations, both with vacuum tunnel, can produce relatively flat phase velocity vs. frequency curve, together with significant values of coupling impedance.

In the above mentioned studies, the PH-SEC is assumed to be a winding of an infinitesimally thin perfectly conducting wire to keep the analysis simple. The effects of the thickness and width of the inclined helix conductors and the straight-edge connections have then been studied with the help of simulations. Two possible realizations of the straight-edge connections, using cylindrical vias and tape vias, have been examined and are compared with the circular and rectangular tape helix. It has been shown that the square PH-SEC is slightly more dispersive than the circular tape helix due to strong coupling at the four 90° corners. As the aspect ratio of the PH-SEC increases, the structure becomes more dispersive and the bandwidth increases. It has also been shown that the cold-test parameters (dispersion characteristics and coupling impedance) of the PH-SEC with high aspect ratio are sensitive to the change of the strip thickness, via diameter, strip width, and ring pad diameter.

The PH-SEC incorporating a coplanar waveguide (CPW) coupler has been designed and fabricated. As proof-of-concept, the unshielded normal helix and the unshielded normal helix with air tunnel configurations have been fabricated to demonstrate the potential for easy fabrication due to straight-edge connections

together with low attenuation and wide bandwidth. These configurations are fabricated using printed circuit techniques which work well for frequencies less than 10 GHz. The measured S -parameters and phase velocity values show a very good agreement with the simulated and analytical results. The unshielded normal helix with air tunnel in the presence of coplanar ground planes has also been examined and fabricated. It is shown that the distance of coplanar ground planes from the planar helix can be used to shape the dispersion curve of the PH-SEC.

The cold-test parameters of the PH-SEC in the presence of a thick silicon substrate beneath the structure have been studied at W-band using simulations. Significant coupling impedance values are found not only within the structure but also above the structure in this configuration. A microfabrication process using three steps - ultraviolet (UV) lithography, electroplating and micromolding (LIGA) - has been proposed and implemented for the designed W-band PH-SEC structure. The proposed fabrication process can produce high aspect ratio PH-SEC structures and is scalable to THz frequency. CPW couplers are designed for the PH-SEC for on-wafer measurements. Many planar helix structures with varying lateral dimensions have been successfully microfabricated. The on-wafer cold-test measurement results have been presented in detail. Effective conductivity values, incorporating the effect of surface roughness, have been obtained for different metal layers in the fabricated structures. The measured return loss and attenuation match well with the simulated results using the effective conductivity values. The issues of silicon wafer resistivity, highest helix temperature and maximum electric field strength have also been addressed. It has been shown that the PH-SEC with a high aspect ratio exhibits a higher insertion loss and higher helix temperature compared to a low aspect ratio structure.

It has been shown that the ratio of the coupling impedance for the backward- and forward-wave can be rather high for the PH-SEC. This may adversely affect the high power operation of a TWT amplifier incorporating the PH-SEC. To solve this problem, a new planar SWS, rectangular ring-bar with straight-edge connections (RRB-SEC), has been proposed and investigated in detail. It has been demonstrated that the dispersion and coupling impedance properties of the RRB-SEC are very similar to those for the circular ring-bar structure. Thus, the ratio of the coupling impedance for the backward- and forward-wave for the RRB-SEC is significantly

less than that for the corresponding circular helix and PH-SEC. Two configurations for the RRB-SEC which can be microfabricated on a silicon substrate have also been presented. The RRB-SEC on a thick silicon substrate is the simpler one to fabricate. However, due to the presence of a thick silicon substrate, the variation in the coupling impedance perpendicular to the substrate is asymmetric, with generally higher values closer to the substrate. Such an asymmetry may reduce the efficiency and gain of a TWT. Hence a configuration with a trench in silicon has also been proposed. The simpler of these two configurations has been designed for operation at W-band and microfabricated on a high-resistivity silicon wafer, using the microfabrication process similar to that for the PH-SEC. It has been shown that the measured *S*-parameters match well with the simulation results using the effective conductivity values.

The hot-test parameters (gain, power, efficiency) for a square PH-SEC and a square RRB-SEC, both on a thin silicon substrate with a trench in silicon beneath the structure have been studied for TWT amplifier operation at W-band. The input-output couplers using a simplified CPW have been proposed to obtain the hot-test parameters over a wide frequency band. It has been shown that such a TWT amplifier incorporating the PH-SEC can produce saturated output peak power varying from 0.65 W to 2.58 W over the frequency range 85 - 110 GHz for a beam voltage of 5.8 kV and beam current of 8 mA. The maximum electronic efficiency is 7.3 % at 97.5 GHz. Rectangular waveguide couplers have also been proposed that can handle these values of output power. It has been shown that the rectangular waveguide couplers reduce the near-dc frequency component significantly compared to that for the simplified CPW couplers. A design of a stable high output power TWT has been achieved using the RRB-SEC. With a beam voltage of 19.7 kV and beam current of 20 mA, a saturated output peak power varying from 9.1 W to 28.6 W over the frequency range 90 - 98 GHz has been obtained.

The studies reported in this thesis have potential applications in printed-circuit TWTs at low frequencies (<18GHz) and microfabricated TWTs at millimeter wave frequencies (30 - 300 GHz). The proposed microfabrication process can also be scaled to fabricate planar helical SWSs at terahertz frequencies (0.3 - 3 THz).

7.2 Recommendations for Future Research

The major conclusions of the thesis and their significance have been mentioned in the previous section. These studies mainly focus on developing the concept of straight-edge connections for the planar helical SWSs. Summarized below are several possibilities for future research.

1. The proposed method of analysis using effective dielectric constant method is based on the sheath helix approximation. Such an approximation does not give enough information on band-edge frequency, modes of propagation and forbidden region. An analysis considering the periodicity or tape helix approach can provide complete modal characteristics for the PH-SEC. Further, the analyses so far consider only the cold-test parameters. However, the hot-test parameters of the TWTs can also be predicted by including the electron beam parameters in the field analysis. In addition, the loss in the materials can also be accounted for in the analysis, providing more accurate predictions. Similar ideas can also be applied to the RRB-SEC.
2. More studies can be carried out to include the thermal contact resistance and the variation in thermal conductivity with temperature in the characterization of the thermal properties of the PH-SEC. These issues are critical for designing a suitable metal enclosure and a packaging method.
3. Attempts should be made to carry out hot-tests using the inverted-helix configuration of Chapter 3 and the rectangular waveguide couplers of Chapter 6. The hot-tests could target a relatively low microwave frequency range (~ 10 GHz) using printed-circuit techniques, or a relatively high frequency range (~ 100 GHz) using microfabrication techniques. In addition, more work can be carried out to obtain the hot-test parameters for the PH-SEC or RRB-SEC incorporating a sheet electron beam.
4. The studies presented in this thesis are in the context of TWT amplifier application based on the forward wave mode amplification. However, as shown in Chapter 6, there are significant coupling impedance values for the fundamental backward wave component of the PH-SEC. Hence, more work can be carried out for characterizing the beam interaction with the backward wave

mode for the PH-SEC and its applications, for example, in a backward wave oscillator (BWO). Also, a non-resonant perturbation measurement technique can be applied to obtain the coupling impedance for the forward and backward wave mode.

5. The helical structure is also popular for applications in antennas due to its wide bandwidth and possibility of circular polarization. Accordingly, the micro-fabricated PH-SEC has great potential to be developed as a THz antenna. Moreover, MEMS switches can be incorporated with the proposed structure for operation as an electronically tunable helix antenna. Such tunable antennas can offer many advantages, for example, polarization diversity.

AUTHOR'S PUBLICATIONS

Journal Papers

1. (Invited paper) **C. Chua**, S. Aditya, J. M. Tsai, M. Tang, and Z. Shen, "Microfabricated planar helical slow-wave structures based on straight-edge connections for THz vacuum electron devices," *International Journal of Terahertz Science and Technology*, vol. 4, no. 4, pp. 208-229, Dec 2011.
2. **C. Chua**, J. M. Tsai, S. Aditya, M. Tang, S. W. Ho, Z. Shen, and L. Wang, "Microfabrication and characterization of W-band planar helix slow-wave structure with straight-edge connections," *IEEE Trans. on Electron Devices*, vol.58, no.11, pp.4098-4105, 2011.
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Conference Papers

1. **C. Chua**, S. Aditya, Julius M. Tsai, and Z. Shen, "PIC simulation for W-band planar helix with straight-edge connections," in *IEEE International Vacuum Electronics Conference*, Monterey, 2012. (Accepted)
2. **C. Chua**, S. Aditya, Z. Shen, and J. M. Tsai, "Planar helix slow-wave structure with straight-edge connections in the presence of coplanar ground planes," *International Conference on Information, Communications and Signal Processing*, Singapore, 2011.
3. **C. Chua**, J. M. Tsai, M. Tang, S. Aditya, and Z. Shen, "Microfabrication of a planar helix with straight-edge connections slow-wave structure," *International Conference on Materials for Advanced Technologies*, Singapore, Advanced Materials Research, vol. 254, pp 17-20, 2011.
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6. **C. Chua**, and Z. Shen, “A planar broadband circularly polarized antenna combining printed dipole and microstrip-fed slot,” *IEEE Antennas and Propagation Society International Symposium*, San Diego, 2008.

Patent

1. **C. Chua**, S. Aditya, and Z. Shen, “Planar helix slow-wave structure with straight-edge connections,” Singapore Patent P-No. 170554 [WO2011/096890], granted 13 Apr. 2012.

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APPENDICES

A – RELATIONSHIP BETWEEN THE CONSTANT COEFFICIENTS IN CHAPTER 3

Relationship between the constant coefficients for general configuration in Fig. 3.8
(b):

The continuity of E_z , H_z , E_y and H_y at $x = \pm c$ gives

$$\frac{R_1}{R_2} = \frac{\frac{u_1'}{u_2'} + \tanh(u_1' c)}{\frac{u_1'}{u_2'} - \tanh(u_1' c)} \quad (\text{A-1})$$

$$\frac{C_1}{C_2} = \frac{\frac{u_1' \varepsilon_2}{u_2' \varepsilon_1'} + \tanh(u_1' c)}{\frac{u_1' \varepsilon_2}{u_2' \varepsilon_1'} - \tanh(u_1' c)} \quad (\text{A-2})$$

$$D \cosh(u_1' c) = C_2 \frac{2u_1' \varepsilon_2}{u_1' \varepsilon_2 - u_2' \varepsilon_1' \tanh(u_1' c)} \quad (\text{A-3})$$

$$Q \cosh(u_1' c) = R_2 \frac{2u_1'}{u_1' - u_2' \tanh(u_1' c)} \quad (\text{A-4})$$

The continuity of E_z and E_y at $x = \pm a$ gives

$$B_2 \left(\frac{B_1}{B_2} + 1 \right) = C_2 \left[\frac{C_1}{C_2} e^{u_2' t_2} + e^{-u_2' t_2} \right] \quad (\text{A-5})$$

$$N_2 \left(\frac{N_1}{N_2} - 1 \right) = R_2 \frac{u_3'}{u_2'} \left[\frac{R_1}{R_2} e^{u_2' t_2} - e^{-u_2' t_2} \right] \quad (\text{A-6})$$

The tangential electric field is zero and the tangential magnetic field is continuous at $x = \pm a$, along the direction of conduction, gives

$$\tan(\Psi_{eff}) = \frac{j\omega\mu_0}{u_3'} \frac{N_2 \left(\frac{N_1}{N_2} - 1 \right)}{B_2 \left(\frac{B_1}{B_2} + 1 \right)} \quad (\text{A-7})$$

$$\tan(\Psi_{eff}) = \frac{\frac{j\omega\epsilon_2}{u_2'} C_2 \left(\frac{C_1}{C_2} e^{u_2' t_2} - e^{-u_2' t_2} \right) - \frac{j\omega\epsilon_3}{u_3'} B_2 \left(\frac{B_1}{B_1} - 1 \right)}{N_2 \left(\frac{N_1}{N_2} + 1 \right) - R_2 \left(\frac{R_1}{R_2} e^{u_2' t_2} + e^{-u_2' t_2} \right)} \quad (\text{A-8})$$

The continuity of E_z , H_z , E_y and H_y at $x = \pm e$ gives

$$\frac{B_1}{B_2} = e^{-2u_3' t_3} \frac{\tanh(u_4' t_4) - \frac{u_3' \epsilon_4}{u_4' \epsilon_3}}{\tanh(u_4' t_4) + \frac{u_3' \epsilon_4}{u_4' \epsilon_3}} \quad (\text{A-9})$$

$$\frac{N_1}{N_2} = e^{-2u_3' t_3} \frac{\frac{u_4'}{u_3'} - \tanh(u_4' t_4)}{\frac{u_4'}{u_3'} + \tanh(u_4' t_4)} \quad (\text{A-10})$$

$$A \sinh(u_4' t_4) = B_2 \frac{2e^{-u_3' t_3}}{1 + \frac{u_3' \epsilon_4}{u_4' \epsilon_3} \coth(u_4' t_4)} \quad (\text{A-11})$$

$$M \cosh(u_4' t_4) = N_2 \frac{2 \frac{u_4'}{u_3'} e^{-u_3' t_3}}{\frac{u_4'}{u_3'} + \tanh(u_4' t_4)} \quad (\text{A-12})$$

where $\epsilon_1' = \epsilon_0 \epsilon_{r1}'$, $\epsilon_2 = \epsilon_0 \epsilon_{r2}$, $\epsilon_3 = \epsilon_0 \epsilon_{r3}$ and $\epsilon_4 = \epsilon_0 \epsilon_{r1}$.

B – SUMMARY OF STRUCTURE DIMENSIONS IN CHAPTER 4

DIMENSIONS IN MICROMETER (μm)

Fig.	Description	P_c	L	ST	SW	RD	VD
4.2	Circular helix	1382.4	200	20	50	-	-
	Rectangular helix	1382.4	200	20	50	-	-
	PH-SEC (tape via)	1382.4	200	20	50	-	-
	PH-SEC (cylindrical via)	1382.4	200	20	50	70	50
4.4	PH-SEC (cylindrical via)	1382.4	200	20, 40, 60	50	70	50

C – SUMMARY OF STRUCTURE DIMENSIONS IN CHAPTER 5

DIMENSIONS IN MICROMETER (μm)

Fig.	Description	P_c	L	ST	SW	RD	VD	BW
5.1	PH-SEC	691.2	100	20	50	70	50	-
	(cylindrical via)	2748	700	20	50	70	50	-
5.3	Circular ring-bar	1382.4	200	20	50	-	-	50
	RRB-SEC (tape via)	1382.4	200	20	50	-	-	50
	RRB-SEC (cylindrical via)	1382.4	200	20	50	70	50	50
	Circular helix	1382.4	200	20	50	-	-	-
5.6	RRB-SEC (cylindrical via)	1382.4	200	20	50	70	50	50
	PH-SEC (cylindrical via)	2260	900	20	50	70	50	-
	PH-SEC (cylindrical via)	2512	800	20	50	70	50	-
5.7	RRB-SEC (cylindrical via)	1382.4	200, 250, 300	20	50	70	50	50
5.8	RRB-SEC (cylindrical via)	1382.4	200	20	50	70	50	50, 100, 150
5.9	RRB-SEC (cylindrical via)	1382.4	200	20, 40, 60	50	70	50	50
5.11	RRB-SEC	944	150	3	40	56	40	40
5.12	(cylindrical via)							

Fig.	Description	P_c	L	ST	SW	RD	VD	BW
5.15	RRB-SEC (cylindrical via)	388	192.85	3	40	50	40	40
5.16	RRB-SEC (cylindrical via)	628	152.85	3	40	50	40	40