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Substrate-Integrated Waveguide Based Antenna in Remote Respiratory Sensing

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Substrate-Integrated Waveguide Based Antenna in

Remote Respiratory Sensing

by

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A THESIS

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Abstract

In a hospital environment, continuous respiratory monitoring is important. It could reduce mortality incidents such as in central sleep apnea. Remote monitoring can be achieved using electromagnetic (EM) waves. Experimental EM systems use horn and patch antennas which do not offer low-profile or wide bandwidth features. This thesis proposes a substrate-integrated waveguide based antenna in respiratory monitoring.

A leaky-wave array antenna is designed and fabricated. It operates between 8-9 GHz. The antenna size is 300 mm $\times$ 55 mm $\times$ 0.78 mm. It offers inexpensive beam-scanning ability to scan the bed. The antenna's beam width sufficiently covers a typical hospital bed.

The fabricated antenna has measured return loss higher than 10 dB and radiates more than 90% of power in frequency range of interest. Measured main beam scans in E-plane ($27^\circ \rightarrow 57^\circ$) with frequency. Measured peak gain is higher than 10 dBi.

Keywords: Remote Respiratory Sensing, Substrate-Integrated Waveguide, Leaky-Wave Antenna, Array Antenna

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List of Symbols, Abbreviations and Nomenclature

Symbol	Definition
CSA	Central Sleep Apnea
MRI	Magnetic Resonance Imaging
LDVi	Laser Doppler Vibrometry
EM	Electromagnetic
SIW	Substrate-Integrated Waveguide
LW	Leaky-wave
PCB	Printed Circuit Board
VNA	Vector Network Analyzer
ADC	Analog-to-Digital Converter
mm-waves	Millimeter-waves
RWG	Rectangular Waveguide
w	Width of SIW
d	Diameter of metal vias
s	Longitudinal via spacing
h	Substrate thickness
a	Width of RWG
b	Height of RWG
ϵ	Electric Permittivity
μ	Magnetic Permeability
TE	Transverse Electric
TM	Transverse Magnetic
β	Propagation/Phase Constant
k	Guided Wavenumber

k_c	Cutoff Wavenumber
ω_c	Cutoff Angular Frequency
f_c	Cutoff Frequency
α	Leakage Constant in LW Antenna
k_{LW}	Leaky Wavenumber
k_0	Free Space Wavenumber
θ_m	Main Beam Angle/Direction
ϵ_r	Relative Electric Permittivity
HMSIW	Half Mode SIW
ATSA	Antipodal Tapered Slot Antenna
HFSS	High Frequency Structural Simulator
FEM	Finite Element Method
dB	Decibels
c	Speed of light in free space
TEM	Transverse-Electric-Magnetic
ϵ_e	Effective dielectric constant in microstrip line
w_{feed}	Width of feeding microstrip line
w_e	Width of TEM waveguide
Z_e	Impedance of TEM waveguide
η_0	Free space wave impedance
Z_0	Characteristic Impedance
SMA	SubMiniature Version A (coaxial RF connector)
W	Slot Width
L	Slot Length
p	Slot Period
dB _i	Decibel Isotropic

N	Number of elements in an array
AUT	Antenna Under Test
CW	Continuous Wave
BPM	Breaths Per Minute
δ	Phase Difference
t	Plate Thickness

Chapter 1

Introduction

1.1 Motivation

In a hospital environment, continuous monitoring of respiratory activity has been deemed important due to the fact that respiratory failure is difficult to predict and it can become life-threatening in a few minutes, but it can also build up gradually [1]. There are numerous clinical situations where continuous monitoring of respiratory activity is considered necessary. For example, it may prevent central sleep apnea (CSA) incidence in patients moved from intensive care unit. Also, it is reported in [1] that unexpected cardiac failure during sleep can often be preceded by, or combined with, events of respiratory failure. Additionally, it is well known that after anaesthesia the respiration can be depressed due to anaesthesia drugs which affect both the respiratory center and respiratory muscles [1]. Another clinical situation that affects respiratory activity is the sedation of patients for various purposes during examinations such as magnetic resonance imaging (MRI) [1]. Moreover, prematurely born babies may need continuous monitoring of their respiratory activity, which may last for weeks, to decide whether or not they would need assisted mechanical ventilation.

As a result, developing appropriate devices to continuously monitor respiratory activity has the potential to reduce mortality rate due to respiratory failure. Respiratory activity can be described with parameters such as respiratory rate, respiratory depth (tidal volume: volume of air moved during normal breathing) and level of gas exchange taking place during respiration. Consequently, various methods and devices have been developed to provide information about these respiratory parameters. Some of the different respiratory monitoring methods that currently exist, which are used in clinical settings, are summarized in Table 1.1.

Table 1.1: Examples of Existing Respiratory Monitoring Devices [1]

Method/Device	Measuring Parameter	Sensor Placement
Electromyography	Electrical activity of skeletal muscles	Electrodes on skin
Strain-gauge transducers (resistive strain gauge)	Abdomen and thoracic circumference	Around abdomen and chest
Inductance plethysmography (uses embedded coils)	Abdomen and thoracic circumference	Around abdomen and chest
Fiber-optic plethysmography (uses fiber-optic strain gauge)	Abdomen and thoracic circumference	Around abdomen and chest
Spirometry	Air volume and/or flow	Inside oral region

Table 1.1 lists examples of respiratory monitoring methods/devices that are presently used. Note that all of them need to be in direct contact with the patient as seen in "Sensor Placement" column. Some of the listed monitoring devices are bulky and invasive such as the spirometer shown in Figure 1.1. The spirometer involves the patient breathing through a mouthpiece to measure respiratory activity which is inconvenient. Additionally, other monitoring devices involve the patient wearing them as a belt around the chest and abdomen as in Figure 1.2, which could be taken off by unconscious patients. Other scenarios include critically ill patients that may not tolerate a mouthpiece, or unconscious and sedated patients that cannot collaborate during measurement time, or burn-injured patients that cannot tolerate wearing a monitoring belt or electrodes as in Figure 1.3. Also, patients who are unconscious, heavily sedated or even sleeping tend to rip off any object from skin. Moreover, direct contact increases the risk of infections in a hospital environment, thus, it carries associated costs of equipment sanitation.

As a result of the aforementioned critical scenarios that a patient may undergo, there is a need to develop remote/non-contact devices that continuously monitor respiratory activity in a reliable and non-invasive way. Also, remote sensing offers the advantage of overcoming some of the artifacts and variations in sensitivity associated with electrodes/belts being strapped onto the subject [2].

Figure 1.1: Spirometry

Figure 1.2: Plethysmography

Figure 1.3: Electromyography

One approach of developing remote/non-contact respiratory monitoring devices is using optical methods such as the laser Doppler vibrometry (LDVi). In [3], a non-contact measurement method based on the use of LDVi to determine the respiration rate of a preterm patient is proposed. This optical method uses a single point laser beam, with a diameter of 1 mm and emitted power of less than 1 mW, where the laser beam impinges on the surface of the patient skin. Specifically, this optical method measures the instantaneous velocity of the laser beam point which corresponds to the skin velocity. The measurement setup of [3] has the optical head of the LDVi placed at about 1.5 m from the patient where the laser beam points normally at the patient's skin while lying at rest on a rigid surface. Moreover, the abdominal area at the center of the umbilicus is selected as the measurement area on the patient because the respiration acts of inspiration and expiration are the most detectable there. Consequently, the measured signal of the LDVi has an oscillatory characteristic due to the periodic movement of the abdominal area caused by the respiratory activity (inspiration and expiration). Additionally, the LDVi signal measures the velocity of displacement of the laser beam point on the skin of the abdominal area. After acquiring the raw data, it is first processed to eliminate unwanted components related to cardiac activity and motion artifacts

[3]. Then the signal peaks are identified in which the difference between two successive peaks is the instantaneous respiration period. Measurement results in [3] show that the LDVi is a reliable measurement method for the determination of the respiration rate. However, it is mentioned in [3] that there is a limit with the use of LDVi:

- The dependence of the back-reflected signal by the surface characteristics; where in [3] an ointment with zinc was applied on the skin in order to have sufficient amount of light reflected back to the LDVi optical head

Moreover, the chest's periodic motion due to respiration is damped once the patient is wearing clothes and covered with a blanket. Consequently, the LDVi method would fail in this case due to the fact that the laser beam cannot penetrate fabric. With the aforementioned limitations of the optical method, there is a need to develop remote respiratory sensing using a different approach. One way of doing that is using electromagnetic (EM) waves.

1.2 Electromagnetic Remote Respiratory Sensing

As early as 1975, a microwave technique based on the scattering of continuous wave radiation was developed for measuring respiratory movements with preliminary results for both human and animal subjects [4]. The procedure in [4] makes use of an X-band standard gain rectangular horn directed toward the chest of the subject. Then the reflection from the chest is detected and compared to the transmitted wave to give respiratory information. The measuring technique in [4] is described as being simple, non-invasive and non-contact with the possibility of performing measurements while the subject is fully clothed. Furthermore, the output frequency of the transmitted signal used in [4] was 10 GHz with incident power density in the $\mu\text{W per cm}^2$ range. The theory behind this microwave technique is that upon incidence on the subject, the microwave signal would be modulated by the moving chest wall. Then the scattered energy modulated by the respiratory activity would be detected by a separate receiving horn. After that, the received signal is compared with the transmitted

signal by computing the instantaneous ratio between them. The computed ratio is output as a voltage whose frequency corresponds to the rate of respiration.

Two experiments were conducted on both a rabbit and a human subject using the experimental setup in [4]. In both experiments, the distance between the subject and the horn radiator was 30 cm and a separate receiving horn was located next to the transmitting one. The preliminary results of respiratory measurement for both the rabbit and human subject showed that the output signal, which was the computed ratio between the scattered and the reference signals, was oscillating corresponding to the periodic chest wall movements. These preliminary results were promising and indicated the usefulness of the microwave technique over conventional techniques for respiration measurements. Problems such as skin irritation, restriction of breathing and loose electrode connections are eliminated with EM approach.

The early experiments of remote respiratory sensing using EM approach were not adopted due to the bulky, complex and expensive hardware. However, with technology advancement and specifically the continuous reduction of size and cost of microwave components and systems, the interest in using this sensing technique has greatly increased in recent years. EM sensing offers the following advantages [5]:

- It can be realized with small antennas illuminating from a distance
- Unlike laser-based methods, the antenna does not need to be accurately pointed toward a specific part of the subject; it can generally illuminate the region where the monitored subject is positioned
- EM sensing makes it possible to measure through clothes or fabric (bed sheets and blankets) without signal disturbances

A number of examples for EM remote respiratory sensing are found in the literature such as [5], [6], [7], [8] and [9]. Table 1.2 summarizes the antenna type, number of antennas and operating frequency used by these EM systems.

Table 1.2: Examples of EM Remote Respiratory Sensing

EM System	Antenna Type	Number of Antennas	Operating Frequency
Scalise [5]	Horn antenna	1 for Tx/Rx	6 GHz
Massagram [6]	Patch antenna	2 (1 for Tx, 1 for Rx)	2.4 GHz
Baltag [7]	Horn antenna	1 for Tx/Rx	8-10 GHz
Lubecke [8]	Patch antenna	2 pairs (1 for Tx, 1 for Rx)	10 GHz
Pieraccini [9]	4-element patch array	2 (1 for Tx, 1 for Rx)	2.42 GHz

Based on Table 1.2 and to the author’s knowledge, an antenna specifically designed and built for EM remote respiratory sensing is not found. For example, a horn antenna is bulky and non-compact; thus, it is not suitable to be used in a commercialized respiratory monitoring device in the future. As for patch antennas, it is difficult to achieve wide bandwidth with them which can be a restriction if it is desired to operate at multiple frequencies. As a result, there is a need for an antenna that is specifically designed for remote respiratory sensing. It is desired to design an antenna with relatively high gain in order to form a directive beam that precisely illuminates a patient’s bed and reduce EM exposure to the surrounding area. By designing a high gain antenna, it also ensures that the reflected signal is strong enough to be detected by the sensing system, thus, easing the requirement for complex and expensive RF components to detect weak signals and reducing cost. It is also desired for the sensing system to scan a patient’s bed and locate the moving chest for monitoring. This is important because patients are of different sizes and have various lying positions on a bed, thus, it is more efficient and easier to have an antenna that automatically scans the bed rather than manually tilting the antenna towards the patient’s chest. One way to achieve beam scanning in an antenna is the use of phase shifters that introduce phase change in excitation current (further detail is provided in Section 4.3); however, they are expensive. Consequently, it is desired to achieve beam scanning (an attribute that has not been explored before in remote respiratory sensing) at a low cost through the use of antenna properties such as frequency (further detail is provided in Section 4.2).

1.3 Design of Substrate-Integrated Waveguide Based Antenna in Remote Respiratory Sensing

The main goal of this research is to design and build an antenna to be applied in remote respiratory sensing. It is desired for this antenna to have the following attributes:

- Low profile and compact
- Simple and inexpensive fabrication process
- Beam-scanning ability to scan the patient's bed area and locate the moving chest (a feature that has not been implemented before in EM remote respiratory sensing to the best of the author's knowledge)
- Directive beam with a beam width that sufficiently covers the patient's bed area

This work presents a leaky-wave (LW) array antenna that incorporates substrate-integrated waveguide (SIW) technology. The operating frequency range is 8-9 GHz. To be more precise, the antenna developed in this thesis constitutes the following components:

- SIW technology
- LW antenna
- Array structure

These components collectively achieve the aforementioned desired attributes. Each of these components is briefly explained here and more detail is provided in the following chapters.

SIW is a relatively new, cost effective, low-loss signal transmission technology that is starting to replace conventional bulky waveguides, especially in high operating frequencies. It is a simple structure based on planar dielectric substrates with top and bottom metal layers, in which the substrate layer is perforated with metalized vias/holes. Therefore,

relatively inexpensive printed circuit board (PCB) fabrication processes can be utilized to fabricate the proposed antenna.

Additionally, a LW antenna with transverse slots [10] is employed in which the leakage/radiation is obtained by introducing periodic set of transverse slots on top of the SIW. This LW antenna is an excellent candidate for it has a beam-scanning ability with frequency which is necessary to scan the patient's bed area and locate the chest. Furthermore, an array structure is incorporated to produce a directive beam with a beamwidth that sufficiently covers the bed area, in which a typical hospital bed size is approximately $1.1 \text{ m} \times 2.3 \text{ m}$.

1.4 Thesis Outline

This thesis has the following outline:

- Chapter 2 provides relevant background about EM respiratory sensing, SIW technology and LW antennas.
- Chapter 3 proposes the sensing system setup and SIW based antenna to be used.
- Chapter 4 details the design methodology for the antenna and presents simulated results.
- Chapter 5 presents the measured results.
- Chapter 6 provides conclusion, contributions and suggestions for future work.

Chapter 2

Background

This section provides background material necessary to understand the different elements of this thesis. First of all, it provides and analyzes a number of examples in literature for EM remote respiratory sensing. These examples give an insight about the working principle of respiratory sensing using EM waves, the different components used in such sensing systems and the antenna types and operating frequencies applied. This section also introduces SIW technology, its operating principles and design techniques. Moreover, this section provides background about LW antennas and their theory. All these different topics collectively help in understanding the research work presented in this thesis and what it entails from design procedures and results.

2.1 Examples of EM Remote Respiratory Sensing

2.1.1 Example #1

A preliminary experiment of an apnea detector that used radar to monitor respiratory movement dates back to 1971 [11]. The detector is composed of a radar transmitter and receiver operating in the 10 GHz range; i.e. wavelength of around 3 cm, and at very low power level. If an object is moving (e.g. a moving chest wall) and has a component of motion normal to the wave, there is a phase shift in the received wave. Depending on the relative movement of the object with respect to the operating wavelength, the generated shift could be either of several cycles or less than one cycle.

The maximum power density at the antenna aperture is $50 \mu\text{W}/\text{cm}^2$, but at a working distance of 30 cm from the detector the power density is reduced to $1 \mu\text{W}/\text{cm}^2$. Moreover, the instrument has deliberately been made rather insensitive to movements at frequencies

higher than those likely to be encountered in breathing (i.e. greater than 80 breaths/minute), to lessen the likelihood of the instrument being activated by movements extraneous to the patient.

The detector in [11] was tested on newborn infants in incubators. The detector was about 2.5 cm above the incubator with the apparatus working through the walls of the incubator. The detector operation was successful in so far as triggering the alarm signal in two instances of premature babies who had frequent episodes of apnea. Additionally, a study was carried out on adult patients in a typical hospital ward. The detector was positioned vertically pointing down on to the patient's thorax and the respiration was readily detected. An alarm was triggered after the patient held his breath and remained motionless.

2.1.2 Example #2

Another example of an EM sensing system for respiratory activity is developed in [5]. The working principle of this system is based on the measurement of the EM signal back scattered by the monitored subject. In particular, the phase variation of the input reflection coefficient of the antenna illuminating the subject during normal respiratory activity is analyzed. Moreover, the effect of the distance between the subject and the sensing apparatus on the measured signal is evaluated. The study also mentions how the system's sensitivity can be further improved by reducing the antenna beam width.

The experimental setup of [5] uses a vector network analyzer (VNA) to generate an output signal at 6 GHz to feed the antenna. The VNA is also used to measure the reflection coefficient of the antenna (S_{11}) which depends on both the antenna itself and on the reflected signal back scattered from the body. Since the reflection coefficient is a complex quantity, both amplitude and phase time variations can be used to gain information about respiratory activity. The system in [5] used phase variations in order to have a sensitive system. This is because the thorax displacement (ΔD) due to respiratory activity is a small percentage of the total distance between the antenna and subject; hence, amplitude variation is small.

However, ΔD is a significant fraction of the operating wavelength ($\lambda = 5$ cm), therefore the variation of the phase of the reflected signal (ΔP) is easily detectable. Furthermore, a broadband horn antenna was used in the experimental setup of [5] to transmit/receive EM waves. Additionally, an LDVi was used as a reference instrument. Also, the phase of S_{11} signal was filtered to remove disturbances such as motion artifacts, high frequency noise and heartbeats.

An experiment was conducted using the proposed system in [5]. The experiment involved the subject being held at five measurement distances (D): 50, 100, 150, 200 and 250 cm. Additionally, the subject assumed three postures: standing, sitting and lying, for the different measurement distances. The mean and standard deviation of ΔP as a function of the distance (D) for sitting, standing and lying postures were computed. The results showed that the mean of ΔP decreased as the distance D increased. This signal reduction was due to the fact that the radiated beam for larger D was no longer focused exclusively on the subject's thorax, but some of the incident power was also transmitted to the background, thus, it did not contribute to the generation of the reflected signal. It was also noted that at each distance D, the measured phase variations for all three postures were close in values, which indicated the independence of ΔP mean from posture. Furthermore, the mean respiratory rate measured by the LDVi and the proposed EM system for different subject positions (standing, sitting and lying) were compared. The correlation coefficient calculated between the two systems was 0.97 showing the proposed EM system being a viable measurement system.

2.1.3 Example #3

Another microwave sensing system in [6] proposes a non-invasive way of measuring respiratory tidal volume in which it is the volume of air that is inhaled or exhaled in a single breath. The system uses Doppler radar to measure respiration rate and air volume. Specifically, a Doppler radar motion sensing transceiver transmits a radio wave signal and receives

motion-modulated signal reflected from a subject. Since the subject's chest is periodically moving during breathing, the reflected signal undergoes a phase shift proportional to the chest wall displacement. Moreover, there exists a linear relationship between the chest wall displacement and air volume changes, thus, the output of the Doppler radar system should be proportional to the tidal volume [6]. Therefore, the Doppler radar system offers a reliable unobtrusive non-contact chest wall displacement measurement.

The system in [6] uses two 2.4 GHz patch antennas; one for transmitter and one for receiver. Additionally, after the received reflected signal is amplified and filtered, it is then digitized using analog-to-digital converter (ADC). Reference signals are also recorded simultaneously using a spirometer for comparison with radar system measurements. The results in [6] show the relative volume displacement measured by the radar corresponds well with the spirometer measurement. Moreover, the tidal volume is calculated from the peak-to-peak amplitude of the volume displacement. The tidal volume of the radar correlates well with the tidal volume from the spirometer. Furthermore, the instantaneous respiration rate for both the radar system and spirometer is similar.

The aforementioned examples show that remote respiratory sensing is successful using EM waves. In all three examples, the phase shift/variation of the reflection coefficient of the antenna (S_{11}) is the measured parameter and through the use of filters, amplifiers and ADC's, respiratory activity is inferred. Different antenna types and operating frequencies are utilized as well; Example #1 uses a radar transceiver at 10 GHz (antenna type not mentioned), Example #2 uses horn antenna at 6 GHz and Example #3 uses two patch antennas at 2.4 GHz. This shows the flexibility in the design of EM sensing systems with desired antenna structure and operating frequency. However, there is room for improvement to these EM systems such as incorporating a scanning feature at which the antenna beam scans the patient's bed and locates the moving chest for monitoring. This scanning feature in a sensing system has not been explored before and makes it more efficient and easier than

manually tilting the antenna structure towards the patient's chest.

2.2 Substrate-Integrated Waveguide Technology

In recent years, interest in utilizing millimeter-waves (mm-waves) has significantly increased. Mm-waves occupy the frequency spectrum from 30-300 GHz. The reason behind the increased interest in using mm-waves is the inherent advantages these waves offer such as broadband and high-resolution techniques. A variety of applications using mm-waves have been recently proposed including wireless networks, automotive radars, imaging sensors and biomedical devices [12]. However, the success of these applications commercially depends on the availability of a cost-effective technology suitable for the mass-production of components and systems operating at mm-wave frequencies. For example, active components such as local oscillators and mixers are being developed in the form of chip-sets to operate at mm-wave frequencies. There are also other components needed in mm-wave systems; however, they cannot be conveniently integrated in chip-sets because either they are too large or the required performance cannot be achieved by integrated components. Such components include antennas, selective filters and power amplifiers. This results in systems being assembled from one or more chip-sets combined with other components which are fabricated with different technologies. At frequencies lower than mm-waves, system components are typically fabricated in planar technology (microstrip or coplanar waveguides). However, at frequencies higher than 30 GHz microstrip and coplanar waveguides suffer from transmission losses and radiation while cylindrical waveguides, though they have low loss, are expensive and complex to integrate with. Therefore, for successful development of mm-wave systems there is a need for a technology that enables the implementation of different system components with high performance and low cost.

One such technology that has been recently explored by researchers is SIW. SIW is a new form of transmission line. In particular, SIW is an integrated waveguide-like structure

fabricated by using two rows of conducting cylinders/vias embedded in a dielectric substrate that electrically connect two parallel metal plates. Figure 2.1 shows the geometry of an SIW structure. SIW technology enables the implementation of classical non-planar rectangular waveguide (RWG) in planar form. Figure 2.2 shows the geometry of an RWG. The vertical side walls of the RWG are replaced by the two rows of metal vias in an SIW. Moreover, SIW technology conforms with existing planar fabrication techniques such as PCB.

Not only are SIWs similar to RWGs in structure but also SIWs exhibit propagation characteristics similar to RWGs such as field pattern and dispersion properties. Furthermore, SIW structures preserve most of the advantages of conventional metallic waveguides including high quality-factor and high power-handling capability. However, the most significant advantage of SIW technology is the possibility to integrate different system components on the same substrate, including passive and active elements and even antennas. Consequently, there is no need for transitions between elements fabricated with different technologies; hence, reducing losses and parasitics. Moreover, the SIW technology has been applied to a wide variety of microwave components such as post and cavity filters, directional couplers, oscillators, power amplifiers, slot array and leaky antennas, six-port circuits and circulators [12]. Additionally, most of the classical waveguide components have been implemented in SIW technology. For example, an SIW based 3-dB coupler is proposed in [13]. Therefore, SIW technology has the potential to develop cost-effective, easy-to-fabricate and high performance systems operating at high frequencies.

2.2.1 Rectangular Waveguide Theory

RWGs are one of the earliest forms of transmission line and they are used to transport microwave signals. As the name suggests and as shown in Figure 2.2, RWGs have rectangular cross-section. Also, RWGs can be hollow (filled with air) or filled with a material of electric permittivity ϵ and magnetic permeability μ . There are various RWG components such as couplers, detectors, isolators, attenuators and slotted lines that are commercially available

Figure 2.1: Geometry of SIW (w : width of SIW, d : diameter of metal vias, s : longitudinal via spacing, h : substrate thickness) [12]

Figure 2.2: Geometry of RWG (a : width of RWG, b : height of RWG, x : length of RWG)

and operating from 1-220 GHz. One disadvantage of RWGs is that they are bulky, thus, they cannot be easily integrated with miniaturized circuits and systems. In this case, planar transmission lines are utilized such as microstrip and stripline.

There is also the concept of propagating modes in RWGs which will be described briefly here along with key theoretical equations. Based on a Cartesian coordinate system, there are two types of modes that can propagate in RWGs; namely, transverse electric (TE) and transverse magnetic (TM) modes. TE modes are those having no electric field components in the modes' direction of propagation. For example, if TE modes' direction of propagation is along the z-axis (assuming Cartesian coordinate system) then there is no E_z component; only transverse components exist (E_x and E_y). The same definition applies to TM modes but with magnetic field applied instead. Only TE modes will be discussed briefly here along with relevant theoretical equations since they are the only modes that can propagate in SIW structures; TM modes cannot propagate in SIWs which will be explained later.

Each TE mode is defined by a combination of two subscript letters m and n ; TE_{mn} , which only take integer values 0, 1, 2.... For example, TE_{10} mode is the fundamental/dominant mode because it has the lowest cutoff frequency. For each TE mode a propagation constant β is defined as follows [14, p. 106]:

$$\beta = \sqrt{k^2 - k_c^2} = \sqrt{k^2 - \left(\frac{m\pi}{a}\right)^2 - \left(\frac{n\pi}{b}\right)^2}, \quad (2.1)$$

where k is the wavenumber of the material filling the RWG ($k = \omega\sqrt{\mu\epsilon} = 2\pi f\sqrt{\mu\epsilon}$), k_c is the cutoff wavenumber, a and b are the width and height of RWG respectively as shown previously in Figure 2.2. β of a TE mode is an important quantity in which having a real value corresponds to a propagating mode. Based on 2.1, β can be real only when

$$k > k_c = \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2}, \quad (2.2)$$

where

$$k_c = \omega_c\sqrt{\mu\epsilon} = 2\pi f_c\sqrt{\mu\epsilon}. \quad (2.3)$$

As it can be seen from equation 2.3, k_c is a function of frequency (f_c) and the filling material. Therefore, β is a function of frequency as well as the geometry of the guide. Also, based on equations 2.3 and 2.2 each TE mode has a cutoff frequency given by

$$f_c = \frac{k_c}{2\pi\sqrt{\mu\epsilon}} = \frac{1}{2\pi\sqrt{\mu\epsilon}} \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2}. \quad (2.4)$$

The mode with the lowest cutoff frequency is called the fundamental/dominant mode. Assuming $a > b$, the lowest f_c occurs for TE₁₀ ($m = 1, n = 0$) mode which is reduced from equation 2.4 to be

$$f_{c_{TE_{10}}} = \frac{1}{2a\sqrt{\mu\epsilon}}. \quad (2.5)$$

For further detail about RWG theory, [14] is an excellent resource.

2.2.2 Operation Principles of SIW structures

As mentioned earlier, SIW structures exhibit propagation characteristics similar to RWGs, provided that the metallic vias are closely spaced and leakage between the via gaps can be neglected [12]. Specifically, SIW modes are similar to a subset of guided modes in a RWG, namely TE_{*m*0} modes with $m = 1, 2, \dots$. For example, the fundamental mode of an SIW is similar to the fundamental TE₁₀ mode of a RWG, with vertical electric current density on the side walls. Figure 2.3 shows the electric amplitude profile of the fundamental SIW mode which is very similar to the fundamental RWG mode. Since the vertical metal walls in a RWG are replaced by via fences for SIW structures, propagating modes of SIW are very close to, but not exactly the same as, those of RWGs. Specifically, the vias are shorted to both top and bottom metal planes of the substrate to provide vertical current paths as shown in Figure 2.4. Consequently, the via pitch is an important parameter in SIW structures and it must be small enough to mimic the vertical metal walls of RWGs, otherwise the propagation characteristics of SIW will be significantly degraded.

Modes characterized with only vertical electric current density on the side walls are the ones that propagate in SIWs. That is why TE₁₀ mode can be supported in an SIW

Figure 2.3: Electric Amplitude Profile of Fundamental SIW Mode [12]

Figure 2.4: Left: SIW structure, Right: Surface current distribution for TE_{10} mode in RWG [15]

because the current path will not be cut by the via fences as shown in Figure 2.4. This holds for all TE_{m0} modes since their current distributions on the side walls are similar. However, horizontal components of the surface current exist on the sidewalls for all TM modes and TE_{mn} modes with nonzero n 's [15]. These horizontal current paths will be cut in SIW structures, which results in radiation. That is why only TE_{m0} modes exist in SIW structures.

After discussing and proving the similarity between SIW and RWG, empirical relations have been proposed between the geometrical dimensions of the SIW and the effective width a of the RWG with the same propagation characteristics. These empirical relations assist in the design of SIW structures by initially designing RWG with desired f_c and propagation characteristics then translating the RWG to an SIW using those empirical equations without the need of full-wave analysis tools. One of the most popular relations is [12]

$$a = w - \frac{d^2}{0.95s}. \quad (2.6)$$

An important aspect to consider when designing SIW structures is reduction of losses which is critical when operating at high frequencies. Just like RWGs, SIW structures would suffer from conductor and dielectric losses. Conductor losses (due to finite conductivity of metal walls) and dielectric losses (due to lossy dielectric material) in SIW can be estimated using classical equations in the literature used for RWGs. It is known that increasing the substrate thickness can significantly reduce conductor losses. The reason behind this is that for a given transmitted power, increasing the dielectric thickness reduces the amplitude of the electric current density flowing in the top and bottom metal layers, thus reducing the power dissipated [16, p. 527]. In other words, the attenuation constant due to conductor loss is almost proportional to the inverse of substrate thickness. Therefore, increasing h is an effective way to reduce conductor loss in SIW structures. However, variation in the diameter of the metal vias and in their longitudinal spacing has minor effects in reducing conductor loss [16, p. 527]. As for dielectric losses, they depend on the dielectric material as well as the

geometry of the SIW structure; specifically the SIW width [14, p. 98]. Consequently, dielectric losses can be reduced by using better dielectric substrate or changing the SIW width. Additionally, SIW structures suffer from a third loss mechanism that is radiation/leakage losses due to via gaps which are not present in RWGs. Radiation losses can be kept small if $s/d < 2.5$, with $s/d = 2$ being the recommended value [12]. In other words, keeping the via spacing twice the via diameter minimizes the radiation leakage by approaching the condition of continuous metal wall in RWGs. In general, comparing SIW structures to traditional planar transmission lines such as microstrip lines and coplanar waveguides, it is seen that SIW structures can guarantee comparable or lower losses [12].

For further information about SIW structures, design considerations, full-wave modeling and loss analysis the listed resources are a good starting point: [17], [18], [19], [20], [21], [22] and [23].

2.3 Leaky-Wave Antenna Theory

A LW antenna is a radiating structure that is primarily composed of a guiding structure with a leaking mechanism along its length. One of the earliest examples of LW antennas is a RWG with a continuous slit cut along one of its side walls as shown in Figure 2.5 [24] and [25]. There are two parameters that are important in understanding LW antenna theory: phase constant β and leakage constant α . Together β and α form what is called the leaky wavenumber k_{LW} as shown in equation 2.7:

$$k_{LW} = \beta - j\alpha . \quad (2.7)$$

β affects the direction of the radiated beam which is also called the main beam. Since β is dispersive that is, it changes with frequency, changing the frequency causes the beam direction to change as well. Consequently, a LW antenna can have a scanning beam by varying the frequency. As for α , it affects the beamwidth of the radiated beam. For example, if α is large then the main beam would have a large beamwidth because the leakage rate is

large producing a short effective antenna aperture. However, if α is small then the main beam would have small beamwidth due to the small leakage rate thus resulting a long effective aperture.

An individual LW antenna can be modeled as a line source with a power source at one end and a matched load at the other end. Figure 2.6 shows an example of a line-source model where it is assumed power leakage in the region above it. As power is continuously leaking along the length of the line, weaker radiation is observed in the region closer to the load which is depicted with thinner arrows. Additionally, the main beam of such model usually has a narrow beam in the scanning plane; i.e. xz-plane based on Figure 2.6, but a fan beam in the cross-plane [26]. One technique that can be used to reduce the beamwidth in the cross-plane is by placing the LW antenna in array structure.

2.3.1 Types of LW Antennas: Uniform and Periodic

There are two types of LW antennas: uniform and periodic. This distinction is based on whether the guiding structure has a uniform geometry along its length or it is periodically modulated. A *uniform* LW antenna is one that is uniform along the length of the guiding structure. A characteristic of uniform LW antennas is that radiation occurs in the forward quadrant with the beam near end-fire direction at higher frequencies. For the example in Figure 2.6, the forward quadrant is the region where $0^\circ < \theta < 90^\circ$ and end-fire direction is at $\theta = 90^\circ$. Another characteristic of uniform LW antennas is that the guiding structure, before introducing the leaking mechanism (e.g. a slit cut), supports a fast wave where $\beta < k_0$ and k_0 is the free space wavenumber. An example of a guiding structure that inherently supports a fast wave is RWGs.

The second type of LW antennas is the periodic type where some periodic modulation of the guiding structure is introduced, and it is this periodicity that produces the leakage. Specifically, the guiding structure supports a slow wave where $\beta > k_0$ before the introduction of periodic modulation. An example of a slow-wave structure is the microstrip transmission

Figure 2.5: Earliest example of LW antenna: RWG with a continuous slit cut along one of its side walls [26]

Figure 2.6: Line-source model of LW antenna (θ : main beam direction measured from z-axis) [27]

Figure 2.7: Microstrip based periodic LW antenna: (a) the comb line array and (b) series microstrip patch array [27]

line which is non-radiating on its own. However, after introducing appropriately designed periodic perturbations/discontinuities along the length of the microstrip line it starts to leak power along its length; hence, transforming into a LW antenna. Figure 2.7 shows microstrip based periodic LW antennas. It should be noted that the periodic modulation itself is uniform along the structure's length. Another example of leaky-mode excitation on microstrip lines is presented in [28].

The difference between uniform and periodic LW antennas on the basis of fast wave versus slow wave is further explained here. For a uniform LW antenna, the dominant mode is a fast wave ($\beta < k_0$) in which it radiates whenever the structure is open like in Figure 2.5. On the other hand, the dominant mode of a periodic LW antenna is a slow wave ($\beta > k_0$) that does not radiate even though the structure is open, such as a microstrip line. With the introduction of periodic perturbations/discontinuities, it produces an infinity of space harmonics (also called Floquet modes) and the periodic perturbations become radiating elements. Now these space harmonics collectively form a guided wave propagating in the periodic structure. Each space harmonic is characterized by phase constant β_n , where n is

any integer, and related to each other by [26]

$$\beta_n = \beta_0 + \frac{n2\pi}{period}, \quad (2.8)$$

where β_0 is the phase constant of the dominant mode with $n = 0$ (i.e. $\beta_0 = \beta$) and *period* is the periodic distance between perturbations. From equation 2.8, it is evident that a slow wave structure, with $\beta_0 > k_0$, can be induced to radiate from space harmonics with negative values of n and appropriate periodicity. In order to have an antenna radiating only a single beam, periodic LW antennas are designed so that only the first space harmonic ($n = -1$) is fast $\beta_{-1} < k_0$, thus, only one space harmonic is radiating and producing a single beam. A detailed treatment of the subject of space harmonics can be seen in [29].

Another difference between uniform and periodic LW antennas is that the latter have a beam scanning range from backward end-fire (based on Figure 2.6 $\theta = -90^\circ$) through broadside direction (perpendicular to LW antenna axis where $\theta = 0^\circ$) into the forward quadrant, except for a narrow region around broadside where an "open stop band" might occur [26]. However, the open stop band issue has recently been addressed and radiation at broadside has been achieved in [30], [27] and [31]. Therefore, the beam-scanning range is different between uniform and periodic LW antennas. For more information about radiation from fast and slow waves, [32] sheds more light about the topic.

It is also worth mentioning an important relation that provides a good approximation of the beam direction in LW antennas. Equation 2.9 shows how the main beam direction can be approximated using β of the k_{LW} normalized to k_0 [26].

$$\sin \theta_m \approx \frac{\beta}{k_0}, \quad (2.9)$$

where θ_m is the angle/direction of the main beam. It should be noted that equation 2.9 is only valid when $\beta < k_0$; hence, fast-wave range for radiation in LW antennas. An example for each type of LW antennas, and based on SIW technology, is provided next.

2.3.2 SIW Based Uniform LW Antenna Example

An example of a uniform LW antenna is provided in [33]. It is fabricated using an SIW with a long slot etched on the top metal of the SIW, allowing power to leak along its length. Figure 2.8 shows the structure with a straight long slot being at the center-line of the waveguide. It is observed that the side walls of the SIW are meandered/incurved with the same shape with the distance between the two walls remains constant. The reason behind this meandering is to optimize the side-lobe level in order to have single maxima in the radiation pattern which is the main lobe/beam. In this design, the center frequency is chosen to be 35 GHz, the relative permittivity ϵ_r and height of substrate is 2.2 and 1.575 mm respectively. The SIW dimensions are $w = 3.83$ mm, $s = 0.8$ mm and $d = 0.5$ mm using design principles discussed in Section 2.2.

Different design parameters of the antenna are chosen based on simulation of the structure in order to obtain a desired performance. For example, the slot width is varied in simulation to obtain the optimum radiation pattern. As a result, it is found out that having a narrow long slot produces the desired radiation pattern (low side-lobe level). Therefore, the slot width is set to be 0.45 mm. The offset distance x_0 of the slot at different position l , as shown in Figure 2.8, is designed based on a desired side-lobe level of -35 dB which resulted in the meandering profile shown in Figure 2.8. The final dimensions of the LW antenna render a beam-scanning antenna with the main beam moving toward the propagation direction of the leaky wave (y-axis) as frequency increases.

This is a brief summary of the SIW based uniform LW antenna example. Further detail about the design aspects and features of the antenna can be found in [33].

2.3.3 SIW Based Periodic LW Antenna Example

An example of a periodic LW antenna using SIW technology is presented in [34]. Specifically, the antenna architecture uses half mode SIW (HMSIW) to feed periodically spaced antipodal

Figure 2.8: SIW based uniform LW antenna with long slot on top of SIW [33]

tapered slot antenna (ATSA) elements. Figure 2.9 shows the geometry of the periodic LW antenna along with a zoomed-in section of an ATSA element. As the name suggests, HMSIW is based on the SIW structure cut vertically in half at the waveguide's center where a virtual magnetic wall can be approximated. As for the ATSA elements, they periodically perturb the fundamental guided mode causing radiation along the antenna's length through the $n = -1$ space harmonic.

In this example, the periodic LW antenna uses 15 ATSA elements and microstrip feed-lines. In particular, the microstrip transmission line is used for excitation as well as termination of a matched load. Furthermore, analytical techniques along with simulation are used to set the various design parameters shown in Figure 2.9. Moreover, a substrate material with permittivity of 10.2 is used for this antenna. Even though the antenna architecture uses HMSIW, the SIW design equations are used to establish the HMSIW dimensions based on the chosen substrate material and desired cutoff frequency since an HMSIW is just half-section of an SIW.

The final dimensions of the antenna geometry are set based on desired operation frequency between 8-12 GHz. The resultant antenna produces radiation pattern with a narrow fan beam profile and maximum radiation in the antenna's plane. Additionally, the antenna achieves a 10 dB gain with a scan range of 86° (-43° to 43°) including broadside direction. Moreover, the antenna efficiency is approximately 60-75% within the antenna bandwidth.

This ends the summary of the example for HMSIW based periodic LW antenna. The

Figure 2.9: HMSIW based periodic LW antenna with periodic radiating ATSA elements [34]

antenna architecture along with the various design parameters have been shown. The key results of the antenna such as radiation pattern profile, scan range, gain and efficiency have been provided. Further detail about the antenna's design specifications and performance can be found in [34].

2.4 Chapter Summary

This chapter provided background about key topics necessary to design and build an antenna for a remote respiratory sensing system. First of all, the chapter provided a number of examples of remote respiratory sensing found in literature that used the EM approach. The working principle of such sensing systems using EM waves was explained. Also, the different components, antenna types and operation frequencies for each of the examples were discussed. After that, the SIW technology was introduced which is used to develop the antenna. Then the principles of operation and design equations of SIW structures were deliberated. Finally, the last component of this chapter was about LW antennas. The basic

structure of a LW antenna was defined along with key parameters that give insight to the propagation of leaky waves. Moreover, the two types of LW antennas were studied and an example of each type was given.

The author would like to list additional resources for the topics of LW antennas and SIWs. An excellent review and summary of LW antennas can be found in [35]. More Examples of SIW based LW antennas can be found in [36], [10] and [37]. Other types of SIW based antennas that are not LW can be found in [38] and [39]. Also, an example of a LW antenna that is not based on SIW technology can found in [40].

Chapter 3

Proposed Sensing System

As mentioned in Section 1.3, the main goal of this research is to design and build an antenna to be applied in remote respiratory sensing system. This work presents a LW array antenna that incorporates SIW technology. Also, the operating frequency range is 8-9 GHz at which commercially available electronic components can be easily found for future implementation of the entire sensing system. An important feature to have in the sensing system is to scan the patient's bed and locate the moving chest for monitoring. The LW antenna in [10] provides an inexpensive way to scan the patient's bed where the antenna's beam can be scanned with frequency. Moreover, the LW antenna in [10] incorporates a periodic set of transverse slots on top of an SIW. This approach of leakage is uncomplicated in terms of the etching process during fabrication. As for SIW technology, it enables implementing a low-profile and compact antenna through simple and inexpensive fabrication processes such as PCB. Additionally, SIW structures render lower losses compared to other traditional planar structures, such as microstrip or coplanar lines, where they suffer from radiation loss and excitation of surface waves [12].

Another desired feature to have in the sensing system is to precisely cover the patient's bed area in order for the received signal to be back scattered mainly from the patient's chest and less from the background surrounding the patient. Therefore, an antenna that produces a directive beam with a beam width narrow enough to cover the bed area is needed. This can be achieved by incorporating the LW antenna in an array structure. In other words, instead of using a single LW line in the sensing system, another leaky line is added resulting in a two-leaky-line array antenna. An array antenna allows the formation of a more directive beam, with a narrower beam width, compared to a single LW antenna.

The goal of this chapter is to present the proposed sensing system along with its geometry. Also, the fabricated LW array antenna is shown along with its size.

3.1 Proposed Sensing System Geometry

Figure 3.1 shows the geometry of the sensing system. It is proposed to have the antenna placed at a vertical elevation of 1 m above the patient. With this antenna position, the beam will scan the bed and locate the chest as shown to monitor the respiratory activity. Also, Figure 3.1 shows the antenna's coordinate system; specifically, the x and z axes. Figure 3.2 shows the top view of the patient's bed along with its size and coordinate system (x' and y' axes). It should be noted that both x and y axes of the antenna's coordinate system are parallel to x' and y' axes of the bed's coordinate system, respectively. As mentioned earlier, patients may have different lying positions on a bed such as lying on their back, stomach and side. The ideal case is to have the patient lying on his/her back where the antenna scans the bed and locates the chest area to monitor breathing. However, for those cases where the patient is lying on his/her stomach or side, the breathing motion is damped and not as prominent as in the back lying position where the chest area is facing the antenna directly. One potential solution for such a complex scenario is to have two antenna systems, instead of one, where each system scans one side of the bed to monitor breathing.

3.2 Proposed Antenna

Figure 3.3 shows the fabricated LW array antenna with the top and bottom pictures showing the top and bottom sides of the antenna respectively. Also, Figure 3.3 shows the size of the fabricated antenna being 5.5 cm $\times$ 30 cm. Additionally, Figure 3.4 shows a zoomed-in section at one of the antenna ports. Since the proposed antenna is an array of two leaky lines, the input of the antenna is split into two pathways; one for each leaky line. Then at the output of the antenna, the two pathways merge again where a matched load of 50 ohms

Figure 3.1: Proposed sensing system geometry - side view

Figure 3.2: Patient's bed - top view

is placed. Furthermore, Figure 3.4 gives a perspective of the thickness of the antenna which is measured to be 0.78 mm.

3.3 Chapter Summary

This chapter provided a high-level overview of the proposed sensing system including its geometry. Moreover, this chapter presented the fabricated LW array antenna along with its different components such as the SIW structure, the transverse slots etched on the antenna's top side and the array of two leaky lines.

Figure 3.3: Fabricated LW array antenna (top and bottom sides)

Figure 3.4: Fabricated LW array antenna - zoomed in (left: an antenna port, right: antenna's thickness 0.78 mm)

Chapter 4

Design Methodology and Simulation Results

This chapter describes the detailed design procedure of the proposed LW array antenna shown in Chapter 3. As mentioned earlier, the proposed antenna has three main components; namely the SIW structure, LW antenna and the array structure. Each of these components, its design methodology and final dimensions, is discussed thoroughly in this chapter. Full-wave simulations are performed for the LW array antenna using ANSYS HFSS (a commercial FEM solver for 3-D full-wave electromagnetic structure simulations). The simulated results are presented in this chapter and include the return loss, insertion loss, radiation patterns, gain and radiation efficiency. Finally, this chapter shows projections of antenna beam at half-power values onto the bed's plane ($x'y'$ -plane shown in Figure 3.2) for simulated LW array antenna versus single LW antenna. This comparison visually shows the beam width difference between the simulated array and single antennas. Specifically, the width of the projected half-power beam for the array antenna is narrower compared to the single LW antenna due to the array effect which will be discussed in more detail in this chapter.

4.1 Design of SIW Structure

The first design step involved in the proposed LW array antenna is designing the SIW structure. As explained in Chapter 3, SIW technology enables implementing low-profile and compact structures through simple and inexpensive fabrication processes such as PCB. Additionally, SIW structures render lower losses compared to other traditional planar structures, such as microstrip or coplanar lines, where the latter suffers from radiation loss and excitation of surface waves [12]. Consequently, using SIW technology fulfills two requirements for the proposed antenna to have: low profile and compact as well as simple and inexpensive

fabrication process as mentioned in Section 1.3.

Before delving into the design process of the SIW structure, a number of key points need to be predetermined first. These key points include the substrate material to be used and the desired frequency of operation. The substrate material used in this thesis is RT/duroid 5880 from Rogers Corporation with $\epsilon_r = 2.2$ and thickness $h = 0.787$ mm [41]. As for the frequency of operation, it is desired to be within the X band (8-12 GHz) due to the availability of commercial microwave components in this frequency band for future integration into the sensing system.

The design of the SIW structure is started with designing an equivalent RWG. As mentioned in Section 2.2.1, TE_{10} mode is the fundamental/dominant mode in a RWG. Based on equation 2.5 the f_c of the TE_{10} mode depends on the width a of the RWG as well as the substrate material used. Since the desired frequency of operation is in the X band, one can compute a required for a desired f_c of the TE_{10} mode. Therefore, f_c is set to be 7.8 GHz for the TE_{10} mode so that the proposed antenna starts operating at the lower frequencies of the X band (~ 8 GHz). Rearranging equation 2.5, a is computed as follows:

$$a = \frac{c}{2f_{c_{TE_{10}}} \sqrt{\epsilon_r}} = 12.97mm, \quad (4.1)$$

where c is the speed of light in free space. Now, in order to translate the equivalent RWG into an SIW structure, both the via diameter d and via spacing s of the SIW need to be determined first based on equation 2.6. As mentioned in Section 2.2.2, it is recommended to have the ratio $s/d = 2$ in order to reduce leakage between via gaps. Therefore, both s and d are set to 1.6 mm and 0.8 mm respectively. After that, the width of the SIW structure w is computed as follows [12]:

$$w = a + \frac{d^2}{0.95s} = 13.39mm. \quad (4.2)$$

Both the equivalent RWG and its corresponding SIW structure are simulated in HFSS. Figures 4.1 and 4.2 show the magnitude of the electric field $|\vec{E}|$ for the TE_{10} mode inside the RWG and SIW structures respectively at 8 GHz. It can be seen that the profiles of $|\vec{E}|$ in both

figures are similar which validates the equivalence of RWG and SIW structures. Although Figure 4.2 shows some energy leakage from the via gaps, this leaked energy is, however, three orders of magnitude lower than the energy within the two via rows. Therefore, most of the energy is contained within the designed SIW structure and leakage from the via gaps will have negligible effects on the performance of the structure. Figure 4.3 shows the $\vec{E}$ profile for the TE_{10} mode in a cross-sectional view of the SIW structure at 8 GHz which again validates the equivalence of RWG and SIW structures.

Furthermore, β of the TE_{10} mode for the equivalent RWG is computed between 7-12 GHz using equation 2.1 and is shown in Figure 4.4. It can be noticed from Figure 4.4 that normalized β (β/k_0) is non-zero around 7.8 GHz which is the desired f_c for the TE_{10} mode. It should be noted that $\beta/k_0 < 1$ up until around 10.5 GHz. In other words, the SIW structure would be operating in the fast-wave region approximately between 7.8-10.5 GHz which is necessary for radiation as mentioned earlier in Section 2.3.1. Therefore, the upper frequency limit of operation is 10.5 GHz. This upper frequency limit is used to design the LW antenna; specifically, the size and periodicity of the transverse slots on top of the SIW which will be discussed in detail later in this chapter. It should also be noted that the frequency range 7.8-10.5 GHz only has the TE_{10} mode excited in the equivalent RWG, since f_c of the next excited mode TE_{20} is twice that of the TE_{10} mode ($7.8 \text{ GHz} \times 2 = 15.6 \text{ GHz}$) which is above 10.5 GHz.

4.1.1 Microstrip Excitation of SIW Structure

The simplest way to feed an SIW structure is by using a microstrip transmission line. Therefore, a microstrip line is used to both excite and terminate with a matched load the designed SIW structure. The first step in designing the microstrip line is to determine the required width for feeding the SIW. Figures 4.5 and 4.6 show a side view and a top view respectively of the microstrip-fed SIW structure. In [42], a mathematical approach is provided to compute the microstrip width needed to feed an SIW structure. Specifically, the microstrip line

is modeled by an equivalent TEM dielectric-filled waveguide as shown in Figure 4.7. The permittivity of the dielectric in the TEM waveguide is set to be the effective permittivity ϵ_e of the microstrip line. The well-known equation that approximates ϵ_e is [42]:

$$\epsilon_e = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \frac{1}{\sqrt{1 + 12h/w_{feed}}}, \quad (4.3)$$

where w_{feed} is the width of the microstrip line feeding the SIW as shown in Figure 4.6. Then the width of the TEM waveguide w_e is calculated to obtain the same impedance as in the microstrip line. The impedance of the TEM waveguide Z_e is given by [42]:

$$Z_e = \sqrt{\frac{\mu}{\epsilon_0 \epsilon_e}} \frac{h}{w_e}, \quad (4.4)$$

where ϵ_0 is free space permittivity and $\approx 8.854 \times 10^{-12}$ F/m. After that, equation 4.4 can be combined with the analytical equation for the microstrip line impedance as follows [42]:

$$Z_e = \sqrt{\frac{\mu}{\epsilon_0 \epsilon_e}} \frac{h}{w_e} = \begin{cases} \left(\frac{60}{\sqrt{\epsilon_e}}\right) \ln\left(8 \frac{h}{w_{feed}} + 0.25 \frac{w_{feed}}{h}\right) & \text{for } w_{feed}/h < 1 \\ \frac{120\pi}{\sqrt{\epsilon_e} \left[\frac{w_{feed}}{h} + 1.393 + 0.667 \ln\left(\frac{w_{feed}}{h} + 1.444\right)\right]} & \text{for } w_{feed}/h > 1 \end{cases} \quad (4.5)$$

The equivalent model for the step between the microstrip and the SIW is shown in Figure 4.8. It is acknowledged in [42] that this step model is not perfect since the capacitive effects at the end of the SIW in the transition plane are not taken into account because magnetic walls are used to close the SIW. Also, the field distribution in the microstrip model (TEM waveguide) does not perfectly match the real one. However, this approach provides a good starting point to approximate the required w_{feed} in order to have low input reflection (S_{11}).

An empirically derived equation is suggested in [42] which relates the permittivity and width ratios of the microstrip-to-SIW step model in Figure 4.8:

$$\frac{a}{w_e} = 4.38e^{-0.627 \frac{\epsilon_r}{\epsilon_e}}. \quad (4.6)$$

Now combining equations 4.3 and 4.6 and rewriting equation 4.5, the following equations

are obtained:

$$\frac{1}{w_e} = \begin{cases} (\frac{60}{\eta_0 h}) \ln(8 \frac{h}{w_{feed}} + 0.25 \frac{w_{feed}}{h}) & \text{for } w_{feed}/h < 1 \\ \frac{120\pi}{\eta_0 h [\frac{w_{feed}}{h} + 1.393 + 0.667 \ln(\frac{w_{feed}}{h} + 1.444)]} & \text{for } w_{feed}/h > 1 \end{cases} \quad (4.7)$$

$$\frac{1}{w_e} = \frac{4.38}{a} e^{-0.627 \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r \epsilon_r - 1}{2\sqrt{1 + 12h/w_{feed}}}}, \quad (4.8)$$

where η_0 is the free space wave impedance and is equal to 377 ohms. h , ϵ_r and a are known parameters but w_e and w_{feed} are unknown. By equating equations 4.7 and 4.8, w_{feed} can be solved for. With $h = 0.787$ mm, $\epsilon_r = 2.2$ and $a = 12.97$ mm, w_{feed} is found to be 3.98 mm. Then the SIW structure in Figure 4.5 with $w_{feed} = 3.98$ mm is simulated in HFSS. Figure 4.9 shows both the return loss $|S_{11}|$ and insertion loss $|S_{21}|$ in dB of the simulated microstrip-fed SIW structure. It can be seen from Figure 4.9 that $|S_{11}|$ is higher than -10 dB at a number of frequency points between 8-9 GHz. It is desired to have the return loss higher than 10 dB within most of the frequency range 8-10.5 GHz. Therefore, w_{feed} has been varied in simulation using parametric sweep option in HFSS. The parameter w_{feed} is swept between 2-6 mm with step size of 0.1 mm to observe any improvement in the return loss compared to the one when $w_{feed} = 3.98$ mm. With different w_{feed} values, it is observed that $|S_{11}|$ is still higher than -10 dB at different frequency points between 8-10.5 GHz. Therefore, another approach for mitigating the input reflection is needed.

It is mentioned in [10] that tapering the SIW width w enhances the bandwidth of operation. Therefore, w is tapered at both ends of the SIW structure as shown in Figure 4.10. Specifically, w is incrementally increased while keeping other parameters of the SIW constant. Now there is no defined method into determining the incremental width step size (w_{step}), length of the tapered SIW section ($l_{TaperedSIW}$) and w_{feed} after tapering except for simulation iterations of the three parameters: w_{step} , $l_{TaperedSIW}$ and w_{feed} . The final dimensions of these three parameters are chosen based on having $|S_{11}| < -10$ dB within 8-10.5 GHz.

Through simulation in HFSS, a number of incremental width steps have been tried to obtain acceptable performance of the return loss within the frequency range 8-10.5 GHz. Three different incremental width step sizes are used: 0.2 mm, 0.1 mm and 0.09 mm. After iterations of simulation, the results show an improvement in the return loss when w is incremented by steps of 0.1 mm, length of the tapered SIW section is 24 mm and $w_{feed} = 3.4$ mm as shown in Figure 4.11. It can be seen from Figure 4.11 that $|S_{11}| < -10$ dB within most of the frequency range except around 8 GHz where $|S_{11}|$ is slightly higher than -10 dB. Also, $|S_{21}|$ is improved after tapering the SIW width with being higher than -5 dB throughout the frequency range 8-10.5 GHz compared to the one in Figure 4.9.

After that the characteristic impedance Z_0 of the feeding microstrip line, with $w_{feed} = 3.4$ mm, is computed using the second equation of 4.5 since $w_{feed}/h > 1$. The computed Z_0 is 39.65 ohms which needs to be matched to a 50-ohm microstrip line since SMA connectors of 50-ohm impedance would eventually be used to enable measurement of antenna performance. The simplest way to match two microstrip lines of different widths is by introducing a linear taper between the two lines as shown in Figure 4.12. The first step into designing the linear microstrip taper is to determine the width of the 50-ohm microstrip line. A "LineCalc" tool, which is available in Advanced Design System (ADS) software by Keysight (Agilent), is used to compute the required 50-ohm microstrip width (w_{50ohm}) for $\epsilon_r = 2.2$ and $h = 0.787$ mm. The result from the "LineCalc" tool is $w_{50ohm} = 2.39$ mm. Then a microstrip line with a width of 2.39 mm is simulated in HFSS to obtain the S-parameters within the frequency range 8-10.5 GHz. Also, the value of w_{50ohm} is varied slightly from the calculated value 2.39 mm to see whether the S-parameters are improved. Specifically, w_{50ohm} is tested with 2.3 mm, 2.4 mm and 2.5 mm and it is observed that $w_{50ohm} = 2.4$ mm rendered better S-parameters in the frequency range 8-10.5 GHz compared to the other tested width values. Figure 4.13 shows the S-parameters of a simulated 50-ohm microstrip line with $w_{50ohm} = 2.4$ mm. It can be observed from Figure 4.13 that $|S_{11}| < -30$ dB and $|S_{21}| > -5$ dB between

8-10.5 GHz which means that the microstrip line is well matched to $Z_0 = 50$ ohms.

With w_{50ohm} and w_{feed} being set, what is left to determine in the microstrip taper architecture is its length $l_{ustripTaper}$ as shown in Figure 4.12. The parameter $l_{ustripTaper}$ is obtained by simulating the complete structure (SIW + microstrip feed line) in HFSS where $l_{ustripTaper}$ is varied in length to obtain acceptable S-parameters of the whole structure. After iterations of simulation, $l_{ustripTaper} = 6$ mm rendered excellent S-parameters with $|S_{11}| < -10$ dB and $|S_{21}| > -5$ dB between 8-10.5 GHz as shown in Figure 4.14. The final architecture of the tapered microstrip-fed SIW structure is shown in Figure 4.15 along with the dimensions of the different tapered sections.

This concludes the section related to the design and excitation of the SIW structure. The following section deals with the design of the LW antenna.

Figure 4.1: Simulated RWG $|\vec{E}|$ at 8 GHz ($a = 12.97$ mm, $b = 0.787$ mm and $\epsilon_r = 2.2$)

Figure 4.2: Simulated SIW $|\vec{E}|$ at 8 GHz ($w = 13.39$ mm, $h = 0.787$ mm, $\epsilon_r = 2.2$, $d = 0.8$ mm and $s = 1.6$ mm)

Figure 4.3: Simulated SIW $\vec{E}$ at 8 GHz (cross-sectional view) ($w = 13.39$ mm, $h = 0.787$ mm, $\epsilon_r = 2.2$, $d = 0.8$ mm and $s = 1.6$ mm)

Figure 4.4: Normalized β of TE_{10} for equivalent RWG ($a = 12.97$ mm, $b = 0.787$ mm and $\epsilon_r = 2.2$)

Figure 4.5: Microstrip-fed SIW - Side View

Figure 4.6: Microstrip-fed SIW - Top View

Figure 4.7: Microstrip line modeled by equivalent TEM waveguide [42]

Figure 4.8: Microstrip-to-SIW step model [42]

Figure 4.9: Simulated Microstrip-fed SIW S-parameters($w_{feed} = 3.98$ mm)

Figure 4.10: Tapered SIW Width

Figure 4.11: Simulated Tapered SIW S-parameters ($w_{feed} = 3.4$ mm, $w_{step} = 0.1$ mm, $l_{TaperedSIW} = 24$ mm)

Figure 4.12: Linear Microstrip Taper

Figure 4.13: Simulated 50-ohm Microstrip Line S-parameters ($w_{50ohm} = 2.4$ mm, $\epsilon_r = 2.2$, $h = 0.787$ mm)

Figure 4.14: Simulated Tapered Microstrip-fed SIW S-parameters ($w_{50ohm} = 2.4$ mm, $w_{feed} = 3.4$ mm, and $l_{ustripTaper} = 6$ mm)

Figure 4.15: Final Structure of Tapered Microstrip-fed SIW

4.2 Design of LW Antenna

After designing the SIW structure, the LW antenna with transverse slots in [10] is incorporated into the structure. As mentioned earlier, the transverse slots provides a simple and inexpensive way to leak the guiding SIW structure through an etching process during fabrication. Figure 4.16 shows the geometry of the slotted SIW LW antenna with three new design parameters being introduced: slot width (W), slot length (L) and slot period (p). Before explaining the design procedure of the transverse slots on top of the SIW in regards to choosing values for W , L and p , a summary of the theory and operation of this LW antenna will be discussed first.

The slotted SIW LW antenna shown in Figure 4.16 is a simple structure and is easy to feed with a microstrip line. This antenna works in the TE_{10} mode of the SIW. Leakage is obtained by introducing a periodic set of centered transverse slots on the top of the SIW (between via centers). Consequently, these slots interrupt the current flow on the top wall. Moreover, this antenna scans in the forward quadrant (xz -plane in Figure 4.16) as frequency changes. Also, three different modes can propagate in the transverse slotted SIW antenna: leaky mode, surface-wave-like mode and proper mode [10]. The leaky mode is when the radiation is mainly due to the LW, which is a fast wave, and the antenna radiates from near broadside (z -axis) to not too close to endfire (x -axis) as frequency increases. Therefore, the antenna works as a LW antenna during leaky mode where $\beta < k_0$. The surface-wave-like mode is when the antenna radiates near endfire (x -axis) and it becomes a surface-wave antenna since radiation is mostly produced by a surface wave ($\beta \approx k_0$), which is a slow wave. The proper mode is when the slotted SIW is not radiating and the energy is confined inside the structure ($\beta > k_0$ slow-wave region).

As for the slot parameters (W , L and p), it is mentioned in [10] that the slotted SIW structure has a small period p compared to the wavelength so that the antenna radiates from the dominant $n = 0$ space harmonic, thus, categorized as quasi-uniform LW antenna and not

periodic. Since this LW antenna is radiating from the dominant mode, it is not considered a periodic LW antenna. Therefore, the transverse slotted SIW is considered a quasi-uniform subtype of uniform LW antenna. Furthermore, the slot width W is made small compared to the slot length L to ensure pure polarization in the slots. In other words, the slot size is designed with the goal to have the tangential electric field in the slots oriented in a single direction (in this case it is the x direction based on Figure 4.16).

Since SIW can be taken as a conventional dielectric-filled RWG, the transverse slotted SIW can be analyzed as a dielectric-filled RWG with transverse slots which is done in [43]. The transverse slotted RWG with dielectric filling is an extension of the investigation performed on transverse slotted RWG with air filling in [44].

In this thesis, the design of p is based on the highest frequency in the range of interest 8-10.5 GHz; recalling that β of equivalent RWG is lower than k_0 up to 10.5 GHz as shown in Figure 4.4. Since the parameter p must be small compared to the wavelength, its value is set initially as approximately ten times smaller than the wavelength at 10.5 GHz; $p = \lambda_{0,10.5GHz}/10 \approx 2.8mm$. As for L , it is set initially to approximately half the width of the equivalent RWG; $L = a/2 \approx 6.5mm$. Since the parameter W must be small compared to L , it is initially set as ten times smaller than L ; $W = L/10 = 0.65mm$. Table 4.1 shows the initial dimensions of p , L and W for the transverse slotted SIW. Since the transverse slots on top of the SIW represent perturbations to the current flow, thus producing unwanted reflections, it is suggested in [10] to linearly taper the slots (tapering L while keeping other slot parameters constant) at both ends of the structure as shown in Figure 4.17. The length of the tapered slot section L_2 in Figure 4.17 is determined through simulation with the goal of minimizing end reflections to produce a more ideal field in the uniform slot region.

The transverse slotted SIW structure is simulated in HFSS using the initial slot dimensions in Table 4.1 and varying L_2 to obtain acceptable S-parameters. Figures 4.18, 4.19, 4.20, 4.21 and 4.22 show the simulated transverse slotted SIW along with its simulated S-

Table 4.1: Initial Slot Dimensions in Transverse Slotted SIW

Design Parameter	Value (mm)
p	2.8
L	6.5
W	0.65

parameters, radiation pattern, gain and radiation efficiency respectively with $L_1 = 117.6$ mm and $L_2 = 54.8$ mm.

Figure 4.19 shows that $|S_{11}| < -10$ dB within the range of interest which indicates that the tapered slots at both ends of the simulated structure did minimize unwanted end reflections. As for $|S_{21}|$, it has low values between 8-9 GHz where it starts slightly higher than -5 dB then decreases below -5 dB indicating that energy is leaked/radiated, and thus that the antenna is working in the leaky mode. Above 9 GHz, $|S_{21}|$ increases to values higher than -5 dB indicating that more energy is transmitted to the matched load than being radiated due to β approaching k_0 as frequency increases and thus leaving the fast-wave region.

Figure 4.20 shows the normalized radiation pattern in the E-plane (xz-plane) at three sample frequency points: 8.2, 8.4 and 8.6 GHz. The beam-scanning ability of the simulated LW antenna is evident where θ_m changes as frequency changes. Also, the side-lobe level is lower than -10 dB as shown in Figure 4.20.

Figure 4.21 shows the peak gain, which is the gain in the direction of maximum radiation (θ_m), of the simulated LW antenna between 8-10.5 GHz. It can be noticed that the gain increases from 5 dBi to 14 dBi between 8-9 GHz corresponding to low values of $|S_{21}|$ in Figure 4.19 where energy is radiated. Then the gain starts to decrease after 9 GHz indicating less energy is radiated as also evident from increasing values of $|S_{21}|$. The same trend can be seen in Figure 4.22 where the radiation efficiency also increases between 8-8.8 GHz then starts decreasing afterward with maximum radiation efficiency of 0.65 at ≈ 8.8 GHz.

The aforementioned simulated results show promising performance of the initial design for the LW antenna between 8-9 GHz in terms of scanning-beam behavior, low end reflections

Figure 4.16: Geometry of transverse slotted SIW LW antenna (W : slot width, L : slot length, p : slot period) [10]

Figure 4.17: Transverse Slotted SIW with Tapered Slots (L_1 : length of uniform slot section, L_2 : length of tapered slot section) [10]

Figure 4.18: Simulated transverse slotted SIW

Figure 4.19: Simulated S-parameters of transverse slotted SIW (initial design: $p = 2.8$ mm, $L = 6.5$ mm, $W = 0.65$ mm, $L_1 = 117.6$ mm and $L_2 = 54.8$ mm)

Figure 4.20: Simulated normalized radiation pattern in E-plane (xz-plane) of transverse slotted SIW (initial design: $p = 2.8$ mm, $L = 6.5$ mm, $W = 0.65$ mm, $L_1 = 117.6$ mm and $L_2 = 54.8$ mm)

and radiation efficiency. However, further iterations of simulation are performed by varying the design parameters p , L , W , L_1 and L_2 in order to achieve better antenna performance, specifically higher radiation efficiency and gain. Table 4.2 shows the final dimensions of the simulated LW antenna in which it achieves higher radiation efficiency and gain compared to the initial design. Also, Figures 4.23 to 4.26 show the simulated results of the final LW antenna design.

Figure 4.23 shows the simulated S-parameters of the final LW antenna design. It can be noticed that $|S_{11}|$ is slightly degraded with values higher than -10 dB between 8-8.15 GHz but it is lower than -10 dB the rest of the frequency range (8.2-10.5 GHz). This slight degradation in $|S_{11}|$ is deemed acceptable as a compromise to achieve higher radiation efficiency and gain as will be shown later. Figure 4.23 also shows $|S_{21}|$ where it reaches values lower than -10 dB indicating more leakage/radiation is achieved compared to the initial design.

Figure 4.24 shows the simulated normalized radiation pattern in E-plane (xz-plane) of the final LW antenna design. Again the beam-scanning ability is evident where θ_m changes as frequency changes (8.2, 8.4 and 8.6 GHz). Also, the side-lobe level is lower than -10 dB as desired. Later in Section 4.4, it is shown that the range of angles achieved by the final LW antenna design meets the requirement of scanning the patient's bed area.

Figure 4.25 shows a comparison of the simulated peak gain between the initial and final designs of the LW antenna. It can be noticed that the gain curve of the final design is slightly shifted to the left compared to the initial design's gain curve. This is due to the difference in slot size and period between the two designs where each design perturbs the fundamental mode differently in the SIW. However, the main goal of achieving higher gain is reached and the final design's gain is higher than the initial design's gain between 8.2-8.9 GHz.

Figure 4.26 shows a comparison of the simulated radiation efficiency between the initial and final designs of the LW antenna. The figure shows also a shift to the left of the final design's radiation efficiency compared to the initial design's curve. Again the main goal of

Table 4.2: Final Dimensions of Transverse Slotted SIW LW Antenna

Design Parameter	Value (mm)
p	2
L	7
W	0.7
L_1	138
L_2	44.8

achieving higher radiation efficiency is attained where the final design's radiation efficiency is higher than the initial design's one between 8.2-8.85 GHz. Also, the maximum simulated radiation efficiency of the final design is 0.74 at 8.7 GHz which is higher than the initial design's one (0.65 at ≈ 8.8 GHz).

Based on the simulated results of the final LW antenna design, it can be deduced that the antenna is operating in the leaky mode up to ≈ 8.8 GHz. This can be verified by obtaining β of k_{LW} in the transverse slotted SIW structure between 8-10.5 GHz. The frequency at which $\beta = k_0$ is the upper frequency limit for operating in the leaky mode. It is suggested in [10] that k_{LW} can be extracted from simulation by sampling the near-field electric field inside the structure along the uniform slotted SIW at a periodic distance p which is the period of the slots. Accordingly, the sampled near-field electric field can be modeled as [10]:

$$E_z = Ae^{-k_{LW}x}, \quad (4.9)$$

where the direction of propagation is in the +ve x-axis direction. Then Matlab can be used to numerically fit the right side of equation 4.9 to the sampled near-field data obtained from HFSS simulation, using a least-square criterion for determining the "best fit" as suggested in [10]. Figure 4.27 shows β of the simulated LW antenna (final design) along with $\beta_{TE_{10}}$ of the equivalent closed (no slots) RWG for comparison purposes. It is evident from Figure 4.27 that $\beta = k_0$ at ≈ 8.8 GHz for the simulated LW antenna indicating that the antenna is operating in the leaky mode (fast-wave region) between 8.2-8.8 GHz. Figure 4.28 shows a comparison of β between the initial (Table 4.1) and final (Table 4.2) designs of the simulated LW antenna. The initial design of LW antenna incorporates smaller slot size (W and L)

and larger slot period (p) compared to the final design. This modification in slot geometry has influenced the fast-wave range ($\beta < k_0$) as seen in Figure 4.28. The initial design of LW antenna has $\beta = k_0$ at ≈ 8.95 GHz, whereas the final design has $\beta = k_0$ at ≈ 8.8 GHz. The initial design of LW antenna has a slightly larger fast-wave range compared to the final design, but lower simulated peak gain and radiation efficiency than the final design.

This concludes the section about LW antenna design and its corresponding simulated results. The following section will deal with incorporating the LW antenna in a two-leaky-line array structure and the design procedure to achieve that.

Figure 4.21: Simulated gain of transverse slotted SIW (initial design: $p = 2.8$ mm, $L = 6.5$ mm, $W = 0.65$ mm, $L_1 = 117.6$ mm and $L_2 = 54.8$ mm)

Figure 4.22: Simulated radiation efficiency of transverse slotted SIW (initial design: $p = 2.8$ mm, $L = 6.5$ mm, $W = 0.65$ mm, $L_1 = 117.6$ mm and $L_2 = 54.8$ mm)

Figure 4.23: Simulated S-parameters of transverse slotted SIW (final design: $p = 2$ mm, $L = 7$ mm, $W = 0.7$ mm, $L_1 = 138$ mm and $L_2 = 44.8$ mm)

Figure 4.24: Simulated normalized radiation pattern in E-plane (xz-plane) of transverse slotted SIW (final design: $p = 2$ mm, $L = 7$ mm, $W = 0.7$ mm, $L_1 = 138$ mm and $L_2 = 44.8$ mm)

Figure 4.25: Simulated gain of transverse slotted SIW (final design: $p = 2$ mm, $L = 7$ mm, $W = 0.7$ mm, $L_1 = 138$ mm and $L_2 = 44.8$ mm)

Figure 4.26: Simulated radiation efficiency of transverse slotted SIW (final design: $p = 2$ mm, $L = 7$ mm, $W = 0.7$ mm, $L_1 = 138$ mm and $L_2 = 44.8$ mm)

Figure 4.27: Normalized β of Simulated LW Antenna (Final Design)

Figure 4.28: Normalized β of Simulated LW Antenna (Initial versus Final Design)

4.3 Design of Array Structure

After finalizing the design of the single LW antenna, which is the transverse slotted SIW structure in Figure 4.18, the next step is to incorporate the leaky line in an array structure. The reason for implementing an array structure is that the single LW antenna has a wide beam width that covers the outside area of the patient's bed as will be explained in more detail later in Section 4.4. Briefly, Section 4.4 presents the beam projection on to the patient's bed at power values that are half the beam's maximum of single LW antenna. It shows a wide beam with a beam projection on to the bed covering approximately 2 m in width. This is larger than the width of a typical hospital bed (≈ 1.1 m). As a result, part of the signal will be transmitted to the background and will not contribute to the generation of the reflected signal. In order to reduce the beam width of the single LW antenna, an array structure is required. A two-leaky-line array structure is designed in which two of the designed LW antennas in Section 4.2 are placed adjacent to each other as shown in Figure 4.29. Before explaining the design procedure of the array structure, a brief summary of array antenna theory is provided first.

An array antenna is a configuration of multiple radiating elements in space in which the elements' individual radiation patterns interact with each other to produce an overall directional radiation pattern. For example, many small antennas can be used in an array to obtain a level of performance similar to that of a single large antenna. However, the mechanical problems associated with a single large antenna are traded for the electrical problems of feeding several small antennas. One unique feature of arrays is that they offer the capability of electronic scanning of the main beam. Particularly, the radiation pattern can be scanned through space by changing the phase of the excitation currents in each element of the array. The array is then called a phased array. Furthermore, arrays can have different geometrical configurations. The most basic one is the linear array where the array element centers lie along a straight line. There is also the planar array such as the

rectangular or circular array where the element centers are contained within a rectangular or circular area.

Arrays offer advantages over single antennas. For example, the main beam of a parabolic reflector antenna is scanned by moving the entire structure, whereas arrays can be phase-scanned at the speed of the control electronics without moving the antenna. However, there are challenges associated with arrays such as complexity of the feeding network and mutual coupling between array elements. The radiation pattern of an array is determined by the type of individual elements used, their orientations, their position in space, and the amplitude and phase of the excitation currents. The analysis of arrays is simplified by first assuming each element of the array to be an isotropic point source. Then the resultant radiation pattern of an array of isotropic radiators is called the *array factor*. An isotropic radiator is a hypothetical, lossless antenna occupying a point in space in which it radiates uniformly in all directions. The array factor can then be used to compute the radiation pattern of actual arrays (arrays of element antennas which are not isotropic) through the principle of pattern multiplication. The principle of pattern multiplication is stated as follows: The electric field pattern of an array consisting of similar elements is the product of the pattern of one of the elements (the element pattern) and the pattern of an array of isotropic point sources with the same locations, relative amplitudes and phases as the original array (the array factor) [45]. Therefore, the complete (normalized) pattern of an array antenna is

$$F(\theta, \phi) = g_a(\theta, \phi)AF_n(\theta, \phi), \quad (4.10)$$

where $g_a(\theta, \phi)$ is the normalized pattern of a single element antenna of the array (the element pattern) and $AF_n(\theta, \phi)$ is the normalized array factor.

The number of elements N present in an array plays a role in changing beam directivity. As N increases, the main beam of the array antenna narrows, thus, producing a more directive beam. An increase in directivity represents an increase in the radiation intensity in the direction of maximum radiation (θ_m). Therefore, choosing N depends on the desired

directivity to be achieved. Also, another important design parameter in array antennas is the spacing between elements. The element spacing affects the presence of grating lobes which are additional major lobes that rise to an intensity equal to that of the main lobe/beam. In most situations, it is undesirable to have grating lobes and the way to eliminate them is by setting the element spacing smaller than a half-wavelength [45].

This concludes the brief summary about array antenna theory in which only key points are mentioned that are necessary to understand the design of the array structure in this thesis. For further detail about array antennas, [45] and [46] are excellent resources.

For the array structure designed in this thesis, N is chosen to be 2. The array of two leaky lines produced a directive beam with a beamwidth sufficiently covering the bed's width (1.1 m) as will be shown in Section 4.4 with half-power beam projections onto bed's plane. Also, the designed array structure is not a phased array in which the interelement excitation phase is zero. In other words, a parallel feed is used in the array structure as shown in Figure 4.29 in which the path length to each element (leaky line) from the feed point is equal; thus, the phase of the excitations will also be equal.

After choosing $N = 2$, the only design parameter left undetermined is the leaky-line separation. Figure 4.30 shows the geometry of the two-leaky-line array structure along with the leaky-line separation parameter. From Figure 4.30 the leaky-line separation is computed as follows:

$$\text{Leaky-line Separation} = w + d + r = 13.39\text{mm} + 0.8\text{mm} + r = 14.19\text{mm} + r. \quad (4.11)$$

Therefore, the unknown parameter r needs to be computed. As mentioned earlier, to avoid having grating lobes in array antennas the element separation should be less than $\lambda/2$. Consequently, the following relation is formed:

$$\text{Leaky-line Separation} < \lambda/2 \quad (4.12)$$

Now the highest operating frequency for the leaky mode is 8.8 GHz based on Figure 4.27. Therefore, λ at 8.8 GHz could be used in equation 4.12; however, λ at 9 GHz is used instead to

produce physically a more compact array structure. It should be noted that the wavelength difference between 8.8 and 9 GHz is very small (≈ 0.8 mm). As a result, equation 4.12 can be rewritten as follows:

$$\text{Leaky-line Separation} < \lambda_{0,9GHz}/2 \quad (4.13)$$

$$14.19\text{mm} + r < 16.65\text{mm}$$

$$r < 2.46\text{mm}.$$

r is chosen to be 2 mm which results in leaky-line separation of 16.19 mm.

Now that N and leaky-line separation parameters are set, the feeding network is explained next. Figure 4.29 shows the configuration of the parallel feed where the input power is split into half to equally feed the two leaky lines. At the other end of the array antenna, the two divided paths are merged again where the matched load is. For a 50-ohm input line, the 3 dB (equal split) power divider can be made by using two 100-ohm output lines [14, p. 316].

The design procedure for a 100-ohm microstrip line is the same as the one explained in Section 4.1.1 for the 50-ohm microstrip line. Therefore, the design process for the 100-ohm microstrip line is discussed briefly here. The "LineCalc" tool from ADS software is used to obtain an initial width value for the 100-ohm microstrip line (w_{100ohm}). Then w_{100ohm} is optimized through HFSS simulation to obtain high return loss and low insertion loss. Figure 4.31 shows the simulated S-parameters of the 100-ohm microstrip line with $w_{100ohm} = 0.7$ mm. It is observed from Figure 4.31 that $|S_{11}| < -20$ dB and $|S_{21}| > -5$ dB between 8-10.5 GHz which indicate that the width value of 0.7 mm is well matched to the 100-ohm microstrip line.

Figure 4.32 shows the 3 dB power divider with a 50-ohm input microstrip line and two 100-ohm output microstrip lines. It is also noticed that both the T-junction discontinuity and right-angled bends are compensated for to reduce unwanted reflections. The compensation methods used here are suggested in [16, p. 182]. Figure 4.33 shows the configuration of T-junction discontinuity compensation. Also, Figure 4.34 shows the configuration for com-

Figure 4.29: Architecture of LW Array Antenna

Figure 4.30: Geometry of LW array antenna showing "leaky-line separation" design parameter

pensated right-angled bend. The power divider in Figure 4.32 is then simulated to verify the performance of S-parameters. Figure 4.35 shows the simulated S-parameters of the designed power divider. It can be noticed from Figure 4.35 that $|S_{11}| < -15$ dB and both $|S_{21}|$ and $|S_{31}|$ are higher than -5 dB indicating low input reflection and high power transmission to the output ports. It is also observed that both $|S_{21}|$ and $|S_{31}|$ curves overlay each other (equal to each other) validating the equal split of power between the two output lines.

After the design of the power divider, the following step is to connect the power divider with the LW array antenna. Now that a 100-ohm microstrip line with $w_{100ohm} = 0.7$ mm is feeding a single leaky line with $w_{feed} = 3.4$ mm, a new microstrip taper is needed to linearly taper w_{100ohm} to w_{feed} (0.7 mm to 3.4 mm). The design process of the new microstrip taper is the same as the one discussed in Section 4.1.1 where $w_{50ohm} = 2.4$ mm is linearly tapered to $w_{feed} = 3.4$ mm. Through HFSS simulation, various taper lengths are trialed to obtain acceptable performance of S-parameters. With iterations of simulation, the new taper length is 15 mm. It is noticed that the new taper length (15 mm) is larger than the previous one (6 mm) due to the larger width difference that needs to be tapered (0.7 mm to 3.4 mm versus 2.5 mm to 3.4 mm). The design of the LW array antenna structure is now complete. Figures 4.36 to 4.39 show the simulated results of the LW array antenna shown in Figure 4.29.

Figure 4.36 shows the simulated S-parameters of the LW array antenna. It is observed that $|S_{11}| < -10$ dB between 8.3-8.8 GHz (leaky mode operation) and at 8.2 GHz $|S_{11}| = -9.3$ dB (close to -10 dB) which indicates the design of both the array structure and power divider is matched within leaky mode frequency range. As for $|S_{21}|$, it is less than -5 dB between 8.2-8.8 GHz where it reaches values below -10 dB between 8.5-8.8 GHz indicating high leakage/radiation of the array structure.

Figure 4.37 shows the normalized radiation pattern in E-plane at different frequencies for the simulated LW array antenna. Again the beam-scanning behavior is observed where θ_m is changing as frequency changes. It is also noticed that the side-lobe level is slightly

Figure 4.31: Simulated S-parameters of 100-ohm microstrip line ($w_{100\text{ohm}} = 0.7 \text{ mm}$)

Figure 4.32: Configuration of 3 dB Power Divider

Figure 4.33: Configuration of T-junction Discontinuity Compensation

Figure 4.34: Configuration of Compensated Right-angled Bend

Figure 4.35: Simulated S-parameters of Power Divider (S_{21} and S_{31} overlap)

higher than -10 dB at 8.2 and 8.4 GHz; however, this should not have adverse effects on the performance of the array antenna. As for the normalized pattern at 8.6 GHz, an odd-shaped beam is observed with two peaks. It is unknown the reason for this peculiar behavior but this as well should not have an adverse effect on the antenna's performance in the sensing system.

Figure 4.38 shows the simulated peak gain of the LW array antenna. Overall, the peak gain performance is excellent in which it is higher than 10 dBi between 8.2-8.8 GHz (leaky mode range). Lastly, Figure 4.39 shows the simulated radiation efficiency of the LW array antenna. It is observed that the radiation efficiency of the simulated antenna achieves values higher than 50% between 8.3-8.8 GHz, reaching above 70% in most of the operation band , and a value of 43% at 8.2 GHz (close to 50%). Therefore, the simulated radiation efficiency of the array antenna is suitable within the leaky mode range.

This concludes the section about array structure design along with its simulated results. The following section presents the half-power beam projection onto the bed's plane at different frequencies using simulated radiation patterns.

Figure 4.36: Simulated S-parameters of LW array antenna

Figure 4.37: Simulated normalized radiation pattern in E-plane (xz-plane) of LW Array Antenna

Figure 4.38: Simulated gain of LW array antenna

Figure 4.39: Simulated radiation efficiency of LW array antenna

4.4 Half-power Beam Projection Using Simulated Results

This section presents the half-power (3 dB) beam projection of both simulated LW array antenna (Figure 4.29) and single LW antenna (Figure 4.18) onto the bed's plane ($x'y'$ -plane shown in Figure 3.2). Specifically, the 3-dimensional (3D) simulated radiation patterns of both the array and single antennas are obtained at different frequencies. The obtained 3D radiation patterns from HFSS provide directional (angular) field strength values. Figure 4.40 shows the antenna's coordinate system as well as the bed's coordinate system. The two coordinate systems are separated by the 1-meter vertical distance as proposed in Section 3.1 with the xy -plane of the antenna's coordinate system parallel to the $x'y'$ -plane of the bed's coordinate system. Also, Figure 4.40 shows the angles θ and ϕ of the antenna's coordinate system where θ is measured from the z -axis and ϕ is measured from the x -axis and it is in the xy -plane. Now for computing half-power beam projection onto the bed's plane, first the 3D radiation patterns are normalized. Then (θ, ϕ) coordinates corresponding to normalized radiation pattern ≈ -3 dB (half power) are obtained. After acquiring (θ, ϕ) coordinates, the radial distances from the origin of the antenna's coordinate system to the bed's plane are computed as follows:

$$\text{Radial distance} = \frac{\text{Antenna-bed vertical distance}}{\cos \theta}, \quad (4.14)$$

where Antenna-bed vertical distance is 1 m. These radial distances are then used to compute the corresponding (x', y') positions on the bed's plane as follows:

$$x' \text{ position} = \text{Radial distance} \times \sin \theta \cos \phi \quad (4.15a)$$

$$y' \text{ position} = \text{Radial distance} \times \sin \theta \sin \phi \quad (4.15b)$$

The results of half-power beam projections onto the bed's plane at three different frequencies (8.2, 8.4 and 8.6 GHz) are shown in Figures 4.41 to 4.43.

Figure 4.41 shows a comparison of the half-power beam projection between array and single antennas at 8.2 GHz. Each dot in Figure 4.41 corresponds to an (x', y') position

computed in equation 4.15. The area enclosed by the green dots corresponds to the half-power beam projection of the single LW antenna. The area enclosed by the black dots corresponds to the half-power beam projection of the LW array antenna. The array factor effect is evident in Figure 4.41 where the beam width of the array antenna is narrower in the y direction than the one of the single antenna. The enclosed area by the array antenna (shown in red lines) spans approximately 1 m in y direction (close to the bed's width of 1.1 m) compared to the area enclosed by the single antenna which spans about 2 m in y direction. In regards to the x direction, both beams are positioned between 0.5 m and 1 m (upper section of bed).

Figure 4.42 shows the half-power beam projection at 8.4 GHz where the area enclosed by each beam is shifted to right compared to the previous position at 8.2 GHz. This is due to the beam-scanning ability in the xz -plane of both array and single antennas. It is evident again here that the array antenna's beam is narrower in y direction compared to the single antenna due to the array effect. The area enclosed by the beam is now positioned at the middle section of the bed corresponding to the right shift in x direction.

Figure 4.43 shows the half-power beam projection at 8.6 GHz with the area enclosed by the beam is shifted further to the right in x direction. It is observed that both beams are now pointing at the lower section of the bed with larger scanning areas compared to the previous ones due to larger computed radial distances. As frequency increases, θ increases as well where the beam is pointing farther away from the z -axis resulting larger radial distance as seen in equation 4.14.

This concludes the section about half-power beam projections and the comparison between LW array antenna and single LW antenna.

Figure 4.40: Illustration of Beam Projection Computation

Figure 4.41: Half-power Beam Projection at 8.2 GHz

Figure 4.42: Half-power Beam Projection at 8.4 GHz

Figure 4.43: Half-power Beam Projection at 8.6 GHz

4.5 Chapter Summary

This chapter explained in detail the design process of the LW array antenna. The design process was divided into three main parts corresponding to the three main components constituting the LW array antenna. The three main components were: SIW structure, single LW antenna and two-leaky-line array antenna. A combination of theoretical equations, empirical equations and HFSS simulations were used to determine the different design parameters of the three components. The SIW structure was designed first by considering an equivalent RWG and choosing its width based on the desired cutoff frequency. Then the equivalent RWG was translated into an SIW structure. After that, a microstrip feed was designed for the SIW and the final structure was simulated in HFSS rendering high return loss and low insertion loss. Next, the single LW antenna was designed by introducing periodic set of transverse slots centered at the top metal of the SIW. The slot size, slot period and length of tapered slot section were finalized through iterations of simulation to yield both acceptable S-parameters and high leakage/radiation. Subsequently, the array structure was designed based on the single LW antenna by incorporating two leaky lines in an array structure. The spacing between the leaky lines was set to less than half a wavelength to avoid having grating lobes based on array antenna theory. Also, a 3 dB (equal split) power divider was designed to provide a parallel feed to the array structure. The final LW array antenna was simulated in HFSS with simulated results showing that $|S_{11}| < -10$ dB between 8.3-8.8 GHz, $|S_{21}| < -10$ dB between 8.5-8.8 GHz, peak gain > 10 dBi between 8.2-8.8 GHz and radiation efficiency $> 50\%$ between 8.3-8.8 GHz and over 70% in most of the band of operation. Finally, a comparison of the half-power beam projection onto the bed's plane was shown between the LW array antenna and single LW antenna. The array effect was apparent in the beam projection comparison where the array antenna's beam was narrower in the y direction than the single antenna's beam. The beam scanning feature was observed as well in which the beam's position was shifting in the +ve x direction as frequency increased. Overall the LW

Table 4.3: Final Simulated Design Parameters of LW Array Antenna

Design Parameter	Value (mm)
w	13.39
d	0.8
s	1.6
SIW Width Increment	0.1
Length of Tapered SIW Section	24
p	2
L	7
W	0.7
Length of Uniform Slot Section	138
Length of Tapered Slot Section	44.8
Width of 50-ohm Microstrip	2.4
Width of 100-ohm Microstrip	0.7
Width of SIW-feeding Microstrip	3.4
Length of Microstrip Taper (from $w_{100ohm} = 0.7$ mm to $w_{feed} = 3.4$ mm)	15
Leaky-line Separation	16.19

array antenna meets the specified requirements in Section 1.3 and scans the entire patient's bed. Table 4.3 shows a summary of the final simulated design parameters for the LW array antenna.

Chapter 5

Fabricated Prototypes and Measured Results

In this chapter, the fabricated prototypes along with their measured results are shown. Specifically, three different prototypes are fabricated: a SIW structure, a single LW antenna and a LW array antenna. The measurements of these three prototypes include the S-parameters, radiation patterns and gain (the latter two measurements are for single and array antennas only and not SIW). The measured results are then compared with the simulated results presented in Chapter 4. Furthermore, all three prototypes are fabricated by Candor Industries using Rogers RT/duroid 5880 substrate material with $\epsilon_r = 2.2$ and thickness of 0.787 mm as mentioned earlier in Section 4.1 [41]. As for the two metal layers (top and bottom), copper is used as the conductor with thickness of 35 μm . Additionally, a pair of SMA end launch connectors, which are 50 ohms and rated up to 27 GHz, from Southwest Microwave are used to perform VNA measurements [47]. Figure 5.1 shows a picture of the SMA connectors used in measurements. Also, the S-parameter measurements of all fabricated prototypes are performed using Agilent 8719ES VNA rated from 50 MHz to 13.5 GHz. As for the radiation pattern and gain measurements of both single and array antennas, the Hewlett Packard 8722D VNA rated from 50 MHz to 40 GHz is used. Moreover, Agilent 85052DN 3.5 mm Economy calibration kit is used in all three prototype measurements to calibrate the VNAs and remove cable effects.

5.1 Fabricated SIW Prototype and Measurement

Figure 5.2 shows the fabricated prototype of the SIW structure. Both the top and bottom sides of the fabricated SIW are shown as well as its size (4 cm $\times$ 18 cm). Figure 5.3 shows the measured return loss compared to the simulated one. The discrepancy between measured

and simulated return loss is observed; however, measured $|S_{11}| < -10$ dB between 8-10.5 GHz except at a few frequency points where it is slightly higher than -10 dB. Figure 5.4 shows the measured insertion loss compared to the simulated one. Due to lower measured return loss compared to simulated result as seen in Figure 5.3, the measured insertion loss is higher than the simulated result as seen in Figure 5.4. However, the measured $|S_{21}| > -3$ dB which indicates acceptable power transmission to port 2.

One reason thought to be the major contributor to the discrepancy between measured and simulated results for the SIW structure was the interface of the connectors with the microstrip lines. After fabricating the prototypes, it was realized that the microstrip line of $w_{50ohm} = 2.4$ mm was too wide and was touching the outer metal case of the connector (essentially touching the ground plane) as shown in Figure 5.5. Consequently, the 50-ohm microstrip line was linearly tapered by cutting off the edges with an exacto knife as shown in Figure 5.6. Since the process of tapering the microstrip edges was done by hand with an exacto knife, it produced slightly asymmetric cuts at each side of the microstrip, resulting in unwanted reflections and discrepancy between measured and simulated results.

Figure 5.1: SMA Connectors

Figure 5.2: Fabricated SIW Prototype

Figure 5.3: Measured Return Loss of SIW Prototype

Figure 5.4: Measured Insertion Loss of SIW Prototype

Figure 5.5: Connector Interface with Microstrip Line

Figure 5.6: Microstrip Tapered Edges Using Exacto Knife

5.2 Fabricated Single LW Antenna Prototype and Measurement

Figure 5.7 shows the fabricated prototype of the single LW antenna. The transverse slots on top of the SIW, which form a single leaky line, are shown. The size of the fabricated LW antenna is $4\text{ cm} \times 28\text{ cm}$, as shown. Again, the microstrip edges at each end of the fabricated prototype are trimmed to enable valid connector-microstrip interface as explained in Section 5.1. Figure 5.8 shows the measured return loss compared to the simulated one. The discrepancy between measured and simulated return loss is observed which again is suspected to be due to the uneven tapered edges of the 50-ohm microstrip line. However, measured $|S_{11}| < -10\text{ dB}$ between 8.3-10.5 GHz and at 8.2 GHz $|S_{11}|$ is measured to be -8.67 dB. Further, the plots have similar shapes.

Figure 5.9 shows the measured insertion loss versus the simulated one. It is noticed that the measured curve is matching that of the simulated one. The measured insertion loss is slightly higher than that of the simulated one due to lower measured return loss than the simulated one as seen in Figure 5.8. One can also observe the drop of measured $|S_{21}|$ to values lower than -10 dB while measured $|S_{11}| < -10\text{ dB}$ indicating high energy leakage and radiation between 8.65-8.8 GHz.

Figure 5.7: Fabricated Single LW Antenna Prototype

Figure 5.8: Measured Return Loss of Single LW Antenna Prototype

Figure 5.9: Measured Insertion Loss of Single LW Antenna Prototype

5.2.1 Radiation Pattern and Gain Measurements

Radiation pattern and gain measurements were conducted using antenna measurement facilities at the University of Calgary. Specifically, the measurement process involved using a Raymond-EMC anechoic chamber equipped with an externally controllable turntable for antenna under test (AUT) rotation and an external Hewlett Packard 8722D VNA rated from 50 MHz to 40 GHz. Figure 5.10 shows the measurement setup used in radiation pattern measurements of the fabricated antennas. Port 1 of the VNA is connected to a standard gain horn which acts as a sensor by being a transmitting antenna. In this thesis, a standard gain ETS Lindgren pyramidal horn antenna [48] (3160 Series, Horn 3160-07, rated between 8.2-12.4 GHz) is used as the sensor and is shown in Figure 5.11. Port 2 of the VNA is connected to AUT (in this case the fabricated antenna prototypes) which acts as the receiving antenna. Both the transmitting (horn) and receiving (AUT) antennas are placed inside the anechoic chamber, whereas the VNA is placed outside the chamber as mentioned earlier. Also, the transmitting and receiving antennas are separated by a distance R as shown in Figure 5.10. In this thesis, the distance $R = 3.68$ m at which the horn antenna is placed at the back end of the chamber for far-field measurement as shown in Figure 5.12. There is also path loss in the traveling space between the horn and AUT and it is equal to (in dB):

$$20 \log\left(\frac{\lambda_0}{4\pi R}\right). \quad (5.1)$$

Figure 5.13 shows the turntable inside the anechoic chamber with AUT placed on top. During chamber measurements, AUT is secured using a simple antenna holder that is made from polystyrene foam to ensure accurate radiation pattern and gain measurements. The turntable enables rotation of AUT in one plane. For measurements in this thesis, rotation is in the xz -plane as shown in Figure 5.14 to measure radiation pattern at different θ values while $\phi = 0^\circ$. As mentioned earlier, θ is measured from the z -axis whereas ϕ is measured from the x -axis and it is in the xy -plane as shown in Figure 4.40.

Figures 5.15, 5.16 and 5.17 show the measured normalized radiation patterns of single

Figure 5.10: Radiation pattern and gain measurement setup [30]

Figure 5.11: ETS Lindgren standard gain horn mounted for co-polarized E-plane measurement

Figure 5.12: Horn antenna placed at the back end inside the anechoic chamber ($R = 3.68$ m)

Table 5.1: Main Beam Angle Comparison (Measured vs. Simulated) of Single LW Antenna with $R = 3.68$ m

Frequency (GHz)	θ_m (Measured)	θ_m (HFSS)
8.2	27°	28°
8.4	41°	42°
8.6	60°	64°

LW antenna in E-plane at 8.2, 8.4 and 8.6 GHz respectively. Also, the simulated results are shown for comparison purposes. At each frequency point, it is observed that the main beams (both measured and simulated results) are well matched and the side-lobe levels are lower than -10 dB. Also, the beam-scanning phenomenon is evident as frequency changes. Table 5.1 shows a comparison between the measured and simulated main beam angles at different frequencies. As frequency increases, θ_m also increases indicating the main beam moving away from the z-axis (since θ is measured from z-axis). Both measured and simulated main beam angles are close in value with 1° difference except at 8.6 GHz where the difference between measured and simulated main beam angles is 4°.

As for gain measurement, a method called "Gain-Transfer (Gain-Comparison) Measurement Method" is used [46, p. 869]. It is the most commonly used method to measure the gain of an antenna. This technique uses a gain standard antenna (with a known gain) to determine absolute gains. Initially, relative gain measurements are performed, which when compared with the known gain of the standard antenna, yield absolute values. The procedure requires two sets of measurements. In one set, using the AUT as the receiving antenna, the received power (P_T) into a matched load is recorded. In the other set, the AUT is replaced by the standard gain antenna (in this thesis it is the standard gain ETS Lindgren pyramidal horn 3160-07) and the received power (P_S) into a matched load is recorded. In both sets, a non-standard horn, operating between 8-12.4 GHz, is used as the transmitting antenna. Also, in both sets, the geometrical arrangement is maintained intact (other than replacing the receiving antennas), and the input power is maintained the same. Then two equations

can be written of the form of [46, p. 867]:

$$G_t + G_r = 20 \log\left(\frac{4\pi R}{\lambda_0}\right) + 10 \log\left(\frac{P_r}{P_t}\right), \quad (5.2)$$

where G_t is gain of transmitting antenna (dB), G_r is gain of receiving antenna (dB), P_r is received power (W) and P_t is transmitted power (W). Equation 5.2 is a logarithmic decibel form of the *Friis Transmission Equation* [46, p. 88]:

$$\frac{P_r}{P_t} = \left(\frac{\lambda_0}{4\pi R}\right)^2 G_t G_r. \quad (5.3)$$

Now writing two equations of the form of equation 5.2, it can be shown that they reduce to [46, p. 870]:

$$G_T = G_S + 10 \log\left(\frac{P_T}{P_S}\right), \quad (5.4)$$

where G_T and G_S are the gains (in dB) of the AUT and standard gain horn 3160-07 respectively.

Figure 5.18 shows the measured peak gain of the single LW antenna versus simulated one between 8.2-8.75 GHz. It can be observed that the measured peak gain curve has a similar trend to the simulated one; however, the measured peak gain is slightly lower than the simulated one. This might be due to path loss effects and uneven taper of microstrip-connector interface. Also, the measured peak gain is higher than 10 dBi except between 8.2-8.35 GHz.

Overall, the fabricated single LW antenna prototype shows similar performance to the simulated one. The measured S-parameters show a slight discrepancy with the simulated results; however, the measured values rendered acceptable performance ($|S_{11}| < -10$ dB between 8.3-10.5 GHz and $|S_{21}| < -10$ dB between 8.65-8.8 GHz). As for the measured radiation patterns, they match well with their corresponding simulated radiation patterns with maximum main beam angle difference of 4° between measured and simulated results. Also, the measured radiation patterns show side-lobe levels lower than -10 dB just like the simulated results. Moreover, the beam-scanning behavior is observed in the measured

radiation patterns at different frequencies. As for the measured peak gain, even though it is lower than the simulated result, it exhibits similar curve trend as the simulated one. The measured peak gain also reaches values higher than 10 dBi between 8.35-8.75 GHz.

Figure 5.13: Fabricated antenna on turntable inside the anechoic chamber

Figure 5.14: Rotation direction of turntable for chamber measurements (rotation in xz -plane at $\phi = 0^\circ$ and varied θ values)

Figure 5.15: Measured normalized radiation pattern of single LW antenna in E-plane (xz-plane) at 8.2 GHz and $R = 3.68$ m

Figure 5.16: Measured normalized radiation pattern of single LW antenna in E-plane (xz-plane) at 8.4 GHz and $R = 3.68$ m

Figure 5.17: Measured normalized radiation pattern of single LW antenna in E-plane (xz-plane) at 8.6 GHz and $R = 3.68$ m

Figure 5.18: Measured peak gain of single LW antenna

5.3 Fabricated LW Array Antenna Prototype and Measurement

This section presents the measured results of the fabricated LW array antenna prototype. The pictures of the LW array antenna prototype are shown in Figures 3.3 and 3.4. The addition of another leaky line is shown in Figure 3.3 and forms an array antenna of two leaky lines compared to the single LW antenna with one leaky line in Figure 5.7. Again, the microstrip edges at each end of the fabricated prototype are cut off to enable valid connector-microstrip interface as explained in Section 5.1. Figure 5.19 shows the measured return loss of fabricated array antenna compared with simulated result. Figure 5.19 shows that the measured $|S_{11}| < -10$ dB between 8.3-10.5 GHz and at 8.2 GHz measured $|S_{11}| = -9.97$ dB. Therefore, the measured return loss is better than the simulated result since the latter shows return loss lower than 10 dB at a few frequency points as seen in Figure 5.19.

Figure 5.20 shows the measured versus simulated insertion loss of the LW array antenna. The measured insertion loss is well matched with the simulated result. The measured $|S_{21}| < -10$ dB between 8.5-8.8 GHz indicating high leakage/radiation at these frequencies. After 8.8 GHz, the measured insertion loss is lower than 10 dB where it reaches values lower than 5 dB indicating most energy is transmitted to the matched load port and the array structure is not working in the leaky mode anymore.

As for the radiation pattern measurement of the LW array antenna, the same measurement setup explained in Section 5.2.1 for the single LW antenna is used here. Figures 5.21, 5.22 and 5.23 show the measured normalized radiation patterns of LW array antenna in E-plane at 8.2, 8.4 and 8.6 GHz respectively. Also, the simulated results are shown for comparison purposes.

At both 8.2 and 8.4 GHz, it is observed that the main beams (both measured and simulated results) are fairly matched. Also, at 8.2 GHz the side-lobe level behavior in the measured result is similar to that of the simulated result; that is both measured and simulated results show side-lobe levels slightly higher than -10 dB. However, at 8.4 GHz the measured

Figure 5.19: Measured Return Loss of LW Array Antenna Prototype

Figure 5.20: Measured Insertion Loss of LW Array Antenna Prototype

Table 5.2: Main Beam Angle Comparison (Measured vs. Simulated) of LW Array Antenna with $R = 3.68$ m

Frequency (GHz)	θ_m (Measured)	θ_m (HFSS)
8.2	29°	26°
8.4	41°	43°
8.6	59°	68°

normalized radiation pattern shows better performance of side-lobe levels where it is lower than -10 dB compared to the simulated result as shown in Figure 5.22. At 8.6 GHz, Figure 5.23 shows better performance of the measured normalized radiation pattern compared to simulated one. As mentioned earlier in Section 4.3, at 8.6 GHz the simulated normalized pattern of the array antenna shows an odd-shaped beam with two peaks. The reason for this peculiar phenomenon is unknown; however, Figure 5.23 shows that the measured normalized pattern does not manifest the odd-shaped beam with two peaks seen in simulated result. Again, the beam-scanning phenomenon is evident as frequency changes just like in single LW antenna. Table 5.2 shows a comparison between the measured and simulated main beam angles at different frequencies.

In addition to radiation pattern measurements at $R = 3.68$ m, pattern measurements of the LW array antenna at $R = 1$ m are conducted. This extra set of measurements at $R = 1$ m is based on the proposed system setup in Section 3.1 where the array antenna would be placed at a vertical height of 1 m above the patient's bed as shown in Figure 3.1. Figures 5.24, 5.25 and 5.26 show the measured normalized radiation patterns of LW array antenna in E-plane at 8.2, 8.4 and 8.6 GHz respectively and $R = 1$ m.

At both 8.2 and 8.4 GHz, the main beams of both measured and simulated patterns are well matched. As for the side-lobe levels, the measured results show better performance than the simulated ones; measured side-lobe levels are lower than -10 dB at both 8.2 and 8.4 GHz whereas simulated side-lobe levels reach values slightly higher than -10 dB at a few θ angles as shown in Figures 5.24 and 5.25. Figure 5.26 shows measured versus simulated normalized patterns at 8.6 GHz. Once more, the odd-shaped main beam, with two peaks, appears in

Figure 5.21: Measured normalized radiation pattern of LW array antenna in E-plane (xz-plane) at 8.2 GHz and $R = 3.68$ m

Figure 5.22: Measured normalized radiation pattern of LW array antenna in E-plane (xz-plane) at 8.4 GHz and $R = 3.68$ m

Figure 5.23: Measured normalized radiation pattern of LW array antenna in E-plane (xz-plane) at 8.6 GHz and $R = 3.68$ m

simulated result but not in measured one with $R = 1$ m. Additionally, the measured side-lobe levels are lower than -10 dB compared to simulated result as shown in Figure 5.26.

Table 5.3 shows a comparison between the measured and simulated main beam angles at different frequencies and $R = 1$ m. Both measured and simulated main beam angles are close in value with 1° difference except at 8.6 GHz where the difference between measured and simulated main beam angles is 4° .

As for the gain measurement of the LW array antenna, the same measurement method (gain-transfer method) explained in Section 5.2.1 for the single LW antenna is used here. Figure 5.27 shows the measured peak gain of LW array antenna between 8.2-8.8 GHz. Also, the simulated result is shown for comparison purposes. It is observed from Figure 5.27 that the measured peak gain closely matches the simulated result with values higher than 10 dBi between 8.2-8.8 GHz.

Table 5.3: Main Beam Angle Comparison (Measured vs. Simulated) of LW Array Antenna with $R = 1$ m

Frequency (GHz)	θ_m (Measured)	θ_m (HFSS)
8.2	27°	28°
8.4	42°	43°
8.6	57°	53°

Figure 5.24: Measured normalized radiation pattern of LW array antenna in E-plane (xz-plane) at 8.2 GHz and $R = 1$ m

Figure 5.25: Measured normalized radiation pattern of LW array antenna in E-plane (xz-plane) at 8.4 GHz and $R = 1$ m

Figure 5.26: Measured normalized radiation pattern of LW array antenna in E-plane (xz-plane) at 8.6 GHz and $R = 1$ m

Figure 5.27: Measured peak gain of LW array antenna

5.4 Measurement of Simulated Chest Movements using LW Array Antenna

As mentioned earlier, phase variation of the reflected signal from the patient's moving chest is the key parameter used to monitor respiratory activity in EM sensing. For this reason, the fabricated LW array antenna is tested to show how well it detects those phase variations due to the moving chest during breathing. The phase variation measurement is simplified by using a metallic surface (600 mm $\times$ 670 mm) placed in front of the array antenna to reflect the signal. Then the reflecting surface is moved closer to the array antenna at various incremental distances. These incremental distances are achieved by placing plates of different thicknesses underneath the reflecting surface in order to move it closer to the array antenna. This simulates the breathing mechanism where the chest moves up and down. Figure 5.28 shows the setup for measuring phase variation where the LW array antenna is suspended above the reflecting surface using the arm of the turntable inside the anechoic chamber. For each new plate placed underneath the reflecting surface, a frequency sweep between 8-9 GHz is performed and the reflection signal (S_{11}) is measured using a VNA. Due to the frequency sweep, the main beam scans the reflecting surface area. Within the frequency sweep, there exists one frequency point where the main beam is normal to the reflecting surface (as shown in Figure 5.28 in yellow), whereas at other frequency points the main beam is at oblique angles to the reflecting surface. The frequency point within the sweep, at which the main beam is normal to the reflecting surface, can be chosen beforehand by rotating the turntable's arm to an angular position equal to the main beam's angle (θ_m) at the chosen frequency point (as shown in Figure 5.28 in orange). This can be achieved using Table 5.3 which shows measured θ_m of LW array antenna at 8.2, 8.4 and 8.6 GHz. As an example, if the turntable's arm is rotated to an angular position of 42° , the main beam will be normal to the reflecting surface at frequency point 8.4 GHz and at oblique angles to the reflecting surface at other frequency points during the frequency sweep. It is also acknowledged that the array antenna is suspended differently than what is explained earlier in Section 3.1 and shown in Figure

3.1. This is because it is desired for the main beam to be normal to the reflecting surface at desired frequency points within the sweep based on the chosen angular position of the turntable's arm. This ensures maximum signal reflection off the reflecting surface at desired frequency points which again simplifies the initial phase variation measurements. In this section, phase variation measurements at angular positions (for the turntable's arm) equal to θ_m of 27° , 42° and 57° , which correspond to main beam normal to reflecting surface at frequency points 8.2, 8.4 and 8.6 GHz respectively during frequency sweep, will be shown. For comparison purposes, a simplified setup of Figure 5.28 (only the LW array antenna and reflecting surface) is simulated in HFSS with main beam normal to reflecting surface at frequency point 8.4 GHz and oblique at other frequency points during the sweep. Moreover, the measured and simulated phase variations are validated by theoretical phase variations. The theoretical phase variations can be computed by measuring the thickness of the different plates placed underneath the reflecting surface and using the following equation:

$$\delta = 2 \times \frac{2\pi}{\lambda_0} \times t, \quad (5.5)$$

where t is the thickness of the plate placed underneath the reflecting surface. It should be noted that the factor of "2" in equation 5.5 is used since it is a reflection measurement (S_{11}) and the signal travels back and forth.

A detailed description of the phase variation measurement at $\theta_m = 42^\circ$ (i.e. main beam normal to reflecting surface at 8.4 GHz during frequency sweep) is provided along with its results. The measured phase variation results of the other two main beam angles (27° and 57°) are provided after that. With the measurement setup used in Figure 5.28, S_{11} is measured using the VNA first when there is only the reflecting surface; i.e. no plates placed underneath. After that, plates of different thicknesses are placed underneath the reflecting surface one at a time and S_{11} is recorded each time. Table 5.4 shows the measured thicknesses (using a caliper) of the different plates placed underneath the reflecting surface. To obtain the phase variation due to the placement of a certain plate, the measured S_{11} phase

Figure 5.28: Phase Variation Measurement Setup

of the previous measurement is used as reference and subtracted from the next (following) measurement. For example, the phase variation due to Plate #1 is computed by subtracting the measured S_{11} phase of only the reflecting surface (previous measurement) from the measured S_{11} phase after placing Plate #1 underneath the reflecting surface (next measurement). Same concept can be used to obtain the cumulative phase variation; i.e. phase variation due to , for example, Plate #1 + #2. Figure 5.29 shows the measured S_{11} phase traces corresponding to the plates in Table 5.4 placed underneath the reflecting surface. Table 5.5 shows a comparison of measured versus theoretical (using equation 5.5) phase variations where $\theta_m = 42^\circ$ (corresponding to main beam normal to reflecting surface at frequency point 8.4 GHz during frequency sweep) for the different plates used. It is noticed from Table 5.5 that the measured phase variations are off by more than 20° for the different cases when compared to the theoretical values. It is suspected that the intrinsic input reflection of the LW array antenna (due to feed transition from microstrip to SIW) might be too high that it is

overshadowing the signal reflected off the reflecting surface back to the antenna. To verify this, the measured data at $\theta_m = 42^\circ$ (main beam normal to reflecting surface at 8.4 GHz during frequency sweep) is transformed to time domain as shown in Figure 5.30. It is noticed in Figure 5.30 that between 0-5 *ns* there are three high amplitude peaks for all five traces corresponding to the different plates used. Also, when zooming in it is noticed that the five traces align exactly with each other between 0-5 *ns*. An example of this is Figure 5.31 where it shows a zoomed-in view of Zone 2 in Figure 5.30. After 5 *ns*, the traces start to have variations until slightly after 10 *ns* as shown in Figure 5.32 (zoomed-in view of Zone 1 in Figure 5.30) due to the plates of various thicknesses. The traces in Figure 5.33 are gated where only responses within the time window of interest; in this case between 5-10 *ns*, are kept and responses outside the time window are zeroed. After that, the gated time domain signals are transformed back to frequency domain and phase variations are computed once again. Table 5.6 shows once again a comparison of measured (after time gating) versus theoretical phase variations at $\theta_m = 42^\circ$ (main beam normal to reflecting surface at 8.4 GHz during frequency sweep) for the different plates used. The measured phase variations in Table 5.6 are significantly improved after time gating since the maximum discrepancy between measured and theoretical is now 26° compared to reaching 100° in some cases before time gating as shown in Table 5.5. As a result, the time gating approach is shown to improve measurement results since the undesired intrinsic input reflections are removed and only the information pertaining to the plates is kept. With that being said, the time domain gating approach would consequently add a required step of calibration in a practical system due to the distance between the antenna and patient playing a role in the signal's time of arrival. Hence, affecting the location of the time window of interest in time axis where signal variations due to plates' placement occur.

The time gating approach is also used at both $\theta_m = 27^\circ$ and 57° (main beam normal to reflecting surface at frequency points 8.2 and 8.6 GHz respectively during frequency sweep)

Table 5.4: Thicknesses of Plates for Phase Variation Measurements

Plate	Thickness t (mm)
Plate #1	19.32
Plate #2	4.43
Plate #3	10.89
Plate #4	5.95

Figure 5.29: Phase Variation Measurement at $\theta_m = 42^\circ$ (8.4 GHz) using LW Array Antenna

Table 5.5: Phase Variation Comparison (Measured vs. Theoretical) at $\theta_m = 42^\circ$ (8.4 GHz) using LW Array Antenna

Plate	δ (Measured)	δ (Theoretical)	$ Error $
Plate #1	8.3°	29.4°	21.1°
Plate #2	1.1°	89.2°	88.1°
Plate #3	348.3°	219.5°	128.8°
Plate #4	14.8°	119.9°	105.1°
Plate #1+#2	9.4°	118.7°	109.3°
Plate #1+#2+#3	357.8°	338.2°	19.6°
Plate #1+#2+#3+#4	12.6°	98.1°	85.5°

Figure 5.30: Time Domain of Measured Signal at $\theta_m = 42^\circ$ (8.4 GHz) using LW Array Antenna

Table 5.6: Phase Variation Comparison (Measured vs. Theoretical) at $\theta_m = 42^\circ$ (8.4 GHz) using LW Array Antenna (Measured data gated)

Plate	δ (Measured)	δ (Theoretical)	$ Error $
Plate #1	55.3°	29.4°	25.9°
Plate #2	67.1°	89.2°	22.1°
Plate #3	218.6°	219.5°	0.9°
Plate #4	126.7°	119.9°	6.8°
Plate #1+#2	122.4°	118.7°	3.7°
Plate #1+#2+#3	341.1°	338.2°	2.9°
Plate #1+#2+#3+#4	107.8°	98.1°	9.7°

Figure 5.31: Zoomed-in Zone 2 of Time Domain Measured Signal at $\theta_m = 42^\circ$ (8.4 GHz) using LW Array Antenna

Figure 5.32: Zoomed-in Zone 1 of Time Domain Measured Signal at $\theta_m = 42^\circ$ (8.4 GHz) using LW Array Antenna

Figure 5.33: Gated Time Domain of Measured Signal at $\theta_m = 42^\circ$ (8.4 GHz) using LW Array Antenna

with Tables 5.7 and 5.8 showing measured phase variations versus theoretical values. Table 5.7 shows that the maximum error between measured and theoretical phase variations is 100° at $\theta_m = 27^\circ$ (main beam normal to reflecting surface at frequency point 8.2 GHz during frequency sweep) even after time gating the measured data. The reason for this is likely because the measured input reflection of the LW array antenna in the setup shown in Figure 5.28 is -9.3 dB at frequency point 8.2GHz (with no reflecting surface placed in front of array antenna) which is higher than -10 dB, whereas the measured input reflections at frequency points 8.4 and 8.6 GHz are -15.2 dB and -13.4 dB respectively. Consequently, at frequency point 8.2 GHz less power is transmitted through the antenna structure; hence, less energy is radiated, compared to frequency points 8.4 and 8.6 GHz. As for Table 5.8, it shows that the maximum error between measured and theoretical phase variations is 16° at $\theta_m = 57^\circ$ (main beam normal to reflecting surface at frequency point 8.6 GHz during frequency

Table 5.7: Phase Variation Comparison (Measured vs. Theoretical) at $\theta_m = 27^\circ$ (8.2 GHz) using LW Array Antenna (Measured data gated)

Plate	δ (Measured)	δ (Theoretical)	Error
Plate #1	4.9°	20.1°	15.2
Plate #2	27.7°	87.1°	59.4
Plate #3	314.7°	214.2°	100.5
Plate #4	37.7°	117°	79.3
Plate #1+#2	32.7°	107.3°	74.6
Plate #1+#2+#3	347.4°	321.5°	25.9
Plate #1+#2+#3+#4	25.1°	78.6°	53.5

Table 5.8: Phase Variation Comparison (Measured vs. Theoretical) at $\theta_m = 57^\circ$ (8.6 GHz) using LW Array Antenna (Measured data gated)

Plate	δ (Measured)	δ (Theoretical)	Error
Plate #1	35.7°	38.7°	3°
Plate #2	104.3°	91.4°	12.9°
Plate #3	208.8°	224.7°	15.9°
Plate #4	139.1°	122.7°	16.4°
Plate #1+#2	140.1°	130.1°	10°
Plate #1+#2+#3	349°	354.9°	5.9°
Plate #1+#2+#3+#4	128.1°	117.7°	10.4°

sweep); smallest error compared to $\theta_m = 27^\circ$ and 42° (frequency points 8.2 and 8.4 GHz of the sweep).

After comparing measured versus theoretical phase variations, the measured phase variations are converted to displacement measurements and compared with theoretical displacement values (plate thicknesses in Table 5.4). The measured displacement values are computed using the following equation:

$$Displacement = \frac{\delta}{2} \times \frac{\lambda_0}{2\pi}, \quad (5.6)$$

where δ is obtained from Tables 5.6, 5.7 and 5.8 (measured δ column). The factor of "2" in equation 5.6 is used again because it is a reflection measurement (S_{11}) and the signal travels back and forth. It should be noted that the measured δ values are between 0° and 360° even in cases where the thickness is larger than $\lambda_0/2$, where measured δ should be greater than 360° . This is because the VNA used to measure S_{11} signal displays phase shifts (corresponding

to the different plates placed underneath the reflecting surface) between 0° and 360° where it wraps around when 360° is reached. As a result, integer multiple of 360° is added to measured δ in cases where the thickness, whether for a single plate or combination of plates, is larger than $\lambda_0/2$ (based on the frequency point at which the main beam is normal to reflecting surface during frequency sweep) before calculating measured displacement using equation 5.6. This phase shift compensation step is done for each θ_m ; 27° , 42° and 57° (corresponding to main beam normal to reflecting surface at frequency points 8.2, 8.4 and 8.6 GHz during frequency sweep). For example, in the case of $\theta_m = 42^\circ$ (main beam normal to reflecting surface at frequency point 8.4 GHz during frequency sweep), $\lambda_0/2 = 17.85$ mm. As a result, the measured phase variations (δ) in Tables 5.5 (before time gating) and 5.6 (after time gating) have 360° added to their values for the following cases: Plate #1, Plates #1+#2 and Plates #1+#2+#3, where the total thickness in each case is larger than 17.85 mm but smaller than 35.71 mm (λ_0 at 8.4 GHz). For the case Plates #1+#2+#3+#4 in Tables 5.5 (before time gating) and 5.6 (after time gating), 720° is added to the measured phase variation (δ) since the total thickness in this case is larger than 35.71 mm (λ_0) but smaller than 53.56 mm ($1.5\lambda_0$ at 8.4 GHz). Figures 5.34, 5.35 and 5.36 show measured displacement (both before and after time domain gating) versus theoretical displacement (plates' thicknesses in Table 5.4). It is expected for the measured displacement data to have a linear trend that increases with theoretical displacement (thickness). This is true for Figures 5.35 and 5.36 after applying time domain gating. Before time domain gating, the measured displacement data did not have a distinguishable linear trend. This again shows how time domain gating improves results. As for Figure 5.34, the measured displacement data does not have a distinguishable linear trend for both cases; before and after time domain gating. As mentioned earlier, this is because the array antenna's measured intrinsic input reflection (without having reflecting surface in front of the antenna) is higher than -10 dB at $\theta_m = 27^\circ$.

Figure 5.34: Displacement Comparison at $\theta_m = 27^\circ$ (8.2 GHz)

Figure 5.35: Displacement Comparison at $\theta_m = 42^\circ$ (8.4 GHz)

As mentioned earlier, a simulation in HFSS is performed of the LW array antenna and reflecting surface at $\theta_m = 43^\circ$ from Table 5.3 (main beam normal to reflecting surface at frequency point 8.4 GHz during frequency sweep) for comparison between simulated and theoretical phase variations. Figure 5.37 shows the layout of the simulation in HFSS. Several iterations of the simulation are performed where the reflecting surface is moved incremental distances closer to the array antenna. These incremental distances imitate the placement of the different plates in Table 5.4 underneath the reflecting surface. For each iteration of the simulation, a frequency sweep between 8-9 GHz is performed with the main beam being normal to reflecting surface at frequency point 8.4 GHz and oblique to reflecting surface at other frequency points during the sweep. Figure 5.38 shows the time domain signals for the different iterations of simulation. The time domain signals from simulation have similar characteristics to the measured ones in Figure 5.30. Figure 5.38 shows high amplitude values at the beginning with traces closely aligning each other between 0-5 ns. For example, Figure 5.39 shows a zoomed-in view of Zone 2 in Figure 5.38 where there is not much variation between the different traces. However, after 5 ns the variations between the traces become significant as shown in Figure 5.40 (due to moving the reflecting surface incremental distances closer to antenna). Table 5.9 shows a comparison of simulated versus theoretical phase variations. Time gating is also used here to compute the simulated phase variations. It is noticed from Table 5.9 that the simulated phase variations closely match the theoretical values with a maximum error of 3° ; much smaller error compared to the measured results. This verifies that the time gating approach is a valid method to mitigate the effects of the unwanted feed-transition reflections in overshadowing the reflected signal from the reflecting surface. The simulation in Figure 5.37 is performed without the connectors in Figure 5.1. It is acknowledged that Plate #1 is not used in the simulation as shown in Table 5.9. There is no reason for not including it except the fact that the simulations for the different plates are quite lengthy in time and the results for Plates #2, #3 and #4 suffice in validating the time domain

Table 5.9: Phase Variation Comparison (Simulation vs. Theoretical) at $\theta_m = 43^\circ$ (8.4 GHz) using LW Array Antenna (Simulated data gated)

Plate	δ (Simulation)	δ (Theoretical)	$ Error $
Plate #2	91°	89.2°	1.8°
Plate #3	219.1°	219.5°	0.4°
Plate #4	121.2°	119.9°	1.3°
Plate #2+#3	310.1°	308.7°	1.4°
Plate #2+#3+#4	71.3°	68.6°	2.7°

gating approach as a preliminary step to accurately compute phase variations. Additionally, Figure 5.41 shows simulated displacement (both before and after time domain gating) versus theoretical displacement (the incremental distances equivalent to plates' thickness in Table 5.4) at $\theta_m = 43^\circ$ (main beam normal to reflecting surface at frequency point 8.4 GHz during frequency sweep). After time domain gating, the simulated displacement data showed a linear trend that increased with theoretical displacement (thickness) unlike the data before time gating.

Figure 5.36: Displacement Comparison at $\theta_m = 57^\circ$ (8.6 GHz)

Figure 5.37: Phase Variation Simulation Setup in HFSS

Figure 5.38: Time Domain of Simulated Signal at $\theta_m = 43^\circ$ (8.4 GHz) using LW Array Antenna

Figure 5.39: Zoomed-in Zone 2 of Time Domain Simulated Signal at $\theta_m = 43^\circ$ (8.4 GHz) using LW Array Antenna

Figure 5.40: Zoomed-in Zone 1 of Time Domain Simulated Signal at $\theta_m = 43^\circ$ (8.4 GHz) using LW Array Antenna

Figure 5.41: Displacement Comparison at $\theta_m = 43^\circ$ (8.4 GHz)

5.5 Chapter Summary

Three different prototypes were fabricated: SIW structure, single LW antenna and LW array antenna. Then each of these fabricated prototypes was measured and compared to simulation results. Measured $|S_{11}|$ of both SIW and single LW antenna prototypes was higher than simulated results. However, the measured $|S_{11}|$ of LW array antenna prototype was slightly better compared to the simulated result. One main reason for differences between simulated and measured results is the microstrip-connector interface having uneven tapered edges. Moreover, the measured $|S_{21}|$ of SIW prototype showed a slight discrepancy with the simulated result. However, the measured $|S_{21}|$ for both the single and array antenna prototypes had similar plots to the simulated results. Also, the measured $|S_{21}|$ for both antenna prototypes reached values lower than -10 dB during leaky mode operation demonstrating high leakage/radiation.

The E-plane radiation patterns of both single and array antenna prototypes were measured at three frequency points (8.2, 8.4 and 8.6 GHz) using known reference horn antenna inside an anechoic chamber. For the single LW antenna prototype, the measured radiation patterns were well matched with simulated results at all frequency points. Also, the maximum difference between measured and simulated main beam angles was 4° at 8.6 GHz. As for the LW array antenna prototype, the measured radiation patterns were also well matched with simulated results at 8.2 and 8.4 GHz. However, at 8.6 GHz the measured radiation pattern was better than the simulated result where the measured main beam had no odd shape with two peaks like the simulated main beam. The reason for this better performance of measured result than simulated one was unknown since sources of disagreement could be introduced during measurement such as cable movement, multi-path reflections and system noise. Additionally, the maximum difference between measured and simulated main beam angles was 4° with $R = 1$ m and 9° with $R = 3.68$ m, both at 8.6 GHz. Furthermore, the measured peak gain of single LW antenna was slightly lower than the simulated result but

it reached values higher than 10 dBi between 8.35-8.75 GHz. As for the measured peak gain of LW array antenna, it closely matched the simulated result with values higher than 10 dBi between 8.2-8.8 GHz.

The LW array antenna was also tested for remote detection of simulated chest movements. A metallic surface was used as the simulated chest since it would reflect back the signal to the array antenna when placed in front of it. Chest movements were mimicked by placing plates of different thicknesses underneath the reflecting surface to move it closer to the array antenna. For every new plate placed, frequency sweep between 8-9 GHz was performed and S_{11} was measured and recorded. The signal reflected from the reflecting surface was weak and hidden in much stronger reflections coming from different system components. This resulted in having huge discrepancy between measured and theoretical phase variations. Time gating method was used on measured data and phase variations were computed at $\theta_m = 27^\circ$, 42° and 57° (corresponding to main beam normal to reflecting surface at frequency points 8.2, 8.4 and 8.6 GHz respectively during frequency sweep). Using time gating method on measured data as a preliminary step for phase variation computations, the results were significantly improved. For the different plates used, the maximum errors between measured and theoretical phase variations were 100° , 26° and 16° at $\theta_m = 27^\circ$, 42° and 57° respectively. The maximum error at $\theta_m = 27^\circ$ was the highest even after time gating which was likely because of the higher input reflection (above -10 dB) to start with compared to the other two frequency points (lower than -10 dB). On the other hand, the phase variation maximum errors at $\theta_m = 42^\circ$ and 57° were much lower after time gating. Then a conversion from phase variations to displacement was also shown where the measured displacement data showed a linear trend that increased with thickness after time domain gating was applied except at $\theta_m = 27^\circ$ due to input reflection being higher than -10 dB. Moreover, a simulation of the reflecting surface, LW array antenna and plates' placement was performed at $\theta_m = 43^\circ$ (main beam normal to reflecting surface at frequency point 8.4 GHz during frequency sweep). The

simulated phase variations were compared to theoretical values. After applying time gating on simulated data, the simulated phase variations closely matched the theoretical values with a maximum error of 3° . Also, displacement variations were computed where simulated displacement data showed a linear trend that increased with thickness (simulated as moving reflecting surface incrementally closer to antenna) after applying time domain gating. This proved the validity of the time gating approach in considerably reducing the effects of strong stationary reflectors.

Chapter 6

Conclusions and Future Work

This thesis proposed an SIW based antenna to be specifically used in a remote respiratory monitoring system. Respiratory activity is one of the vital signs that if measured, could reduce mortality incidents in a hospital environment. For example, continuous monitoring of respiratory activity may prevent central sleep apnea (CSA) incidence in patients moved from intensive care unit. Existing respiratory monitoring systems involve direct contact with the patient which is inconvenient for critically-ill and burn-injured patients. Therefore, there is a need for monitoring systems that remotely measure respiratory activity without direct contact with the patient. Different approaches have been exploited to achieve remote respiratory sensing such as optical and EM methods. EM sensing has the advantage of measuring through clothes and fabric unlike the optical method. EM sensing systems have been developed showing successful measurement of respiratory activity using EM waves. One important component in EM sensing systems is the antenna. The EM sensing systems found in literature used horn and patch antennas. With horn antennas, they are bulky structures which is a disadvantage in terms of sensing system placement since the antenna component must be suspended above the patient's body for respiratory measurement. As for patch antennas, they do not offer wide bandwidth which is a disadvantage if operating at multiple frequencies is desired. Consequently, there is a necessity for an antenna specifically designed for remote respiratory sensing. Also, it is desired for this antenna to have the following attributes:

- Low profile and compact
- Simple and inexpensive fabrication process
- Beam-scanning ability to scan the patient's bed area and locate the chest

- Directive beam with a beam width that sufficiently covers the patient's bed area

In this thesis a LW array antenna that incorporates SIW technology is designed and fabricated. The operating frequency range is between 8-9 GHz. The designed antenna is proposed to be suspended 1 m above the patient's bed of size 1.1 m $\times$ 2.3 m. Additionally, the proposed antenna has three main components: SIW, LW antenna and array structure. Collectively, these three components achieve the goals of this thesis. Firstly, it is desired to have a sensing system that scans the patient's bed and locates the moving chest. This is achieved by designing a LW antenna that provides an inexpensive way to scan the patient's bed through beam scanning with frequency. The LW antenna used in this thesis is a transverse slotted SIW [10] which provides a simple way to radiate power along the guiding structure. Secondly, it is desired to have a directive antenna beam with a beam width that sufficiently covers the patient's bed area. This is achieved by incorporating the LW antenna into an array structure. Two leaky lines (two lines of transverse slotted SIW) are used in this thesis producing a directive beam that is narrow enough to cover the width of patient's bed. Thirdly, it is desired to design an antenna that is low profile and compact. This is achieved by using SIW technology which enables antenna fabrication using inexpensive PCB fabrication processes.

Simulated results of transverse slotted SIW structure (single LW antenna) showed operation in leaky mode up to 8.8 GHz. Moreover, simulated LW array antenna showed $|S_{11}| < -10$ dB between 8.3-8.8 GHz (leaky mode operation) and at 8.2 GHz $|S_{11}| = -9.3$ dB (close to -10 dB). As for simulated $|S_{21}|$, it was less than -5 dB between 8.2-8.8 GHz where it reached values below -10 dB between 8.5-8.8 GHz indicating high leakage/radiation of the array structure. Furthermore, the simulated main beam scanned in E-plane with frequency; $28^\circ \rightarrow 53^\circ$. The simulated peak gain of the LW array antenna was higher than 10 dBi between 8.2-8.8 GHz (leaky mode range). Lastly, the simulated radiation efficiency achieved values higher than 50% between 8.3-8.8 GHz with a value of 43% at 8.2 GHz (close to 50%).

The measured LW array antenna prototype showed $|S_{11}| < -10$ dB between 8.3-8.8 GHz and at 8.2 GHz measured $|S_{11}| = -9.97$ dB. Also, the measured insertion loss was well matched with the simulated result. The measured $|S_{21}| < -10$ dB between 8.5-8.8 GHz indicating high leakage/radiation at these frequencies. After 8.8 GHz, the measured insertion loss was lower than 10 dB and it reached values lower than 5 dB indicating high energy transmission to the matched load port and the array structure was not working in the leaky mode anymore. Furthermore, the measured main beam scanned in E-plane with frequency; $27^\circ \rightarrow 57^\circ$. As for the measured peak gain of LW array antenna, it reached values higher than 10 dBi between 8.2-8.8 GHz.

The half-power beam projection onto bed's plane was also shown using simulated 3-D radiation patterns of LW array antenna. Beam projection showed the beam-scanning phenomenon of the array antenna along the bed's length (x' -axis) as frequency changed. Also, the width of the beam projection in the y' -axis was approximately 1 m which is close to the bed's width of 1.1 m. Therefore, the array structure of two leaky lines produced a directive beam with a beam width that would sufficiently cover the bed's width.

The LW array antenna was also shown to successfully detect simulated chest movements as displacement variations (computed from phase variations of the reflected signal from the reflecting surface) in both simulation and measurement. Time domain gating was used as a necessary preliminary step to have both measured and simulated displacement variations in agreement with theoretical values. In measurements, $\theta_m = 42^\circ$ and 57° (main beam normal to reflecting surface at frequency points 8.4 and 8.6 GHz respectively during frequency sweep) showed much improved results after applying time gating to measured data. Specifically, the measured displacement data had a linear trend that increased with thickness after applying time gating. That was not the case at $\theta_m = 27^\circ$ (main beam normal to reflecting surface at frequency point 8.2 GHz during frequency sweep) where the measured displacement data did not display a linear trend even after time gating. This was attributed to the higher input

reflection of the LW array antenna measured as -9.3 dB at $\theta_m = 27^\circ$ compared to -15.2 dB and -13.4 dB at $\theta_m = 42^\circ$ and 57° respectively. As for simulation, it was performed at $\theta_m = 43^\circ$ (main beam normal to reflecting surface at frequency point 8.4 GHz during frequency sweep) and the results were excellent after using time domain gating. The simulated displacement data had a linear trend that increased with thickness after time gating was applied.

Altogether, it appears that antennas for the purpose of remote respiratory monitoring have to be designed with much smaller input reflection; e.g. $|S_{11}| < -15$ dB. This is because the returning signal from the patient's chest is very small and can be overwhelmed by the reflection of the forward signal (traveling from antenna to patient's chest) due to intrinsic input reflection as well as weakened by reflection as it travels through the feed-transition section again on the way back (from patient's chest to antenna).

6.1 Contributions

Overall, the fabricated LW array antenna shows promise in remote respiratory monitoring. The contributions made in this thesis are summarized as follows:

- Designed and fabricated a low profile and compact antenna structure with size of 300 mm $\times$ 55 mm $\times$ 0.78 mm
- Used SIW technology for simple and inexpensive fabrication process
- Showed proposed antenna's beam-scanning ability on bed's plane
- Showed proposed antenna's beam width sufficiently covering bed's width
- Showed proposed antenna's ability in detecting simulated chest movements as displacement computed from phase variations of back reflected signal
- Produced and presented an international conference paper at IEEE-APS 2015 in Vancouver

Figure 6.1: Addition of Phase Shifters to Array Antenna [49]

6.2 Future Work

A number of improvements can be implemented to extend the functionality of the fabricated array antenna. Firstly, phase shifters could be added to each leaky line in the array structure. The addition of phase shifters enables electronic scanning of the main beam in the y direction as shown in Figure 6.1. Specifically, the phase shifters change the phase of excitation currents in each leaky line of the array antenna becoming a phased array antenna. Therefore, with beam-scanning ability in both x and y directions due to frequency change and phase change respectively, the main beam can now scan other beds. In this case, the array antenna could be placed at the middle of a hospital room's roof and directing the beam towards the patient that requires monitoring.

Another suggested improvement is to increase the number of leaky lines in the array structure. For example, the array antenna could have four leaky lines instead of two. The result of adding more leaky lines would create a more directive beam with narrower beam width. This could be beneficial if smaller bed sizes are used in the hospital or in the case

of new-born babies inside incubators. However, instead of building multiple array antennas where each one has different number of leaky lines, array antennas with fixed number of leaky lines and leaky-line-activation ability could be implemented. For example, a 4-leaky-line array antenna could be fabricated that includes control electronics where activation of 1, 2, 3 or all 4 of the leaky lines is enabled. This ability to control how many leaky lines to be active in the array antenna allows one to control the directivity and beam width. For instance, a hospital bed of size $1.1 \text{ m} \times 2.3 \text{ m}$, it would suffice to activate only two leaky lines out of the four available ones. However, for an incubator which is smaller in size, all four leaky lines can be activated to generate a much narrower beam width for efficient scanning of the incubator area without including the surroundings.

A minor modification of the proposed array antenna is to remove the matched load part and close the end side with a row of metal vias. Finally, other architectures of LW antennas could be explored for the application of respiratory monitoring in which a comparison study could be conducted to choose the design that outperforms.

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