Leaky Wave Antenna

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Leaky Wave Antenna

A thesis submitted in partial fulfillment of the requirements for the degree of Master of Science in Engineering

By

Pradyumna Aditya
B.E., Visvesvaraya Technological University, 2007

2016
Wright State University
WRIGHT STATE UNIVERSITY
GRADUATE SCHOOL

September 2, 2016

I HEREBY RECOMMEND THAT THE THESIS PREPARED UNDER MY SUPERVISION BY Pradyumna Aditya ENTITLED Leaky Wave Antenna BE ACCEPTED IN PARTIAL FULFILLMENT OF THE REQUIREMENTS FOR THE DEGREE OF Master of Science in Engineering.

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Dean, Graduate School
ABSTRACT

Aditya, Pradyumna. M.S., Department of Electrical Engineering, Wright State University, 2016, Leaky Wave Antenna.

The main radiation mechanism in a leaky wave antenna is a travelling wave in a guided structure. The main characteristics of a leaky wave antenna are: light weight, easy to fabricate. Leaky wave antennas have been in use since the 1940s. In this thesis, a portable and powerful leaky-wave antenna is designed, implemented, and demonstrated for scanning application. We change the guiding structure by applying a high dielectric constant material to produce a low-cost, small size, light weight, and high sensitivity leaky-wave antenna. The designed antenna can reach large scan angles with small frequency tuned. High scanning angles can be achieved by slight variation of the operating frequency. The radiation direction of the antenna can be varied with the frequency.
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1. Introduction

1.1 Definition and advantages

Having slots at proper intervals on an electromagnetic wave carrying waveguide results in a plane wave being radiated from these slots. This plane wave was called the leaky wave by N. Marcuvitz [1][2] in 1956. Waveguides, or any electromagnetic energy transmission paths, having such periodic slots which result in leaky waves are called Leaky Wave Antennas.

Leaky wave antennas are wide-band microwave antennas that radiate a narrow beam whose direction can be changed by slight variation in its operating frequency. Essentially, they are a type of phase array antennas. Initial work on array antennas began in the 1920s when a lot of work happened on array theory and its experimentation. Even during the World War II a lot of work happened on array antennas in the United States. Many leading industries like Lincoln Laboratories, General Electric and others showed renewed interest in array antennas in the early 1960s.[3] The reason for this renewed interest could be attributed to the fact that array antennas were simple to be fabricated and cost was low. However there was one major drawback with these antennas: it needed huge operating bandwidth to achieve large tuning angles.

Phase array antennas resulted in high radiation power gain. But to achieve this with continuous angle scanning, a large number of phase shifters that were to be adjusted simultaneously were needed. As the number of phase shifters increase so does the volume of the antenna and with it the cost. In order to address these drawbacks
research was conducted on leaky wave antennas. Leaky wave antennas have distinct advantages over phase array antennas: variation in frequency results in variation in scanning angle – eliminates the need to have phase shifters in the control circuit reducing the energy consumption. Leaky wave antennas are generally simple waveguide structures – reduces the fabrication cost dramatically. Variation of scanning angle with frequency is disadvantageous for point-to-point communication applications [4].

In general, leaky wave antennas follow the below equation:

$$\sin \theta = \frac{\lambda_0}{\lambda} - \frac{\lambda_0}{d} m$$

where:

- $\theta$ is the angle counted from direction perpendicular to wave propagation
- $\lambda_0$ is the wavelength in air
- $\lambda$ is the wavelength inside the waveguide
- $d$ is the perturbation space
- $m$ is any integer

### 1.2 History

The first leaky wave antenna was developed by W.W. Hansen in 1940[5] which was a rectangular closed waveguide as shown in Fig 1 which had long uniform slits that leaked, radiated, the wave propagating. But such structures cut across the current lines that made producing waves with low leakage difficult. Hines and Upson [6] came up with
what they called a “Holey Waveguide” to address this issue. In this structure, as in Fig 2, the waveguide is air-filled and the holes are placed closely.

In 1956, von Trentini introduced the concept of two dimensional leaky wave antenna [7] in which he used periodic partially reflective screen over ground plans to obtain directive beams. This structure comes under the classification of quasi-uniform leaky wave antenna.

Fig 1.1. First known leaky wave antenna – W.W. Hansen
In 1957 Rotman and Karas introduced a new structure which they called “Sandwich Wire Antenna”[8]. Its working principle was too complicated and not practical. Later Rotman along with Oliner in 1959 came up with a new theory called “Asymmetrical Trough Waveguide Antenna”[9]. In this he considered a symmetric trough open waveguide structure that did not radiate by itself but by introducing some sort of asymmetry it begins to behave like a leaky wave antenna.
In 1982 Itoh and Adelseck demonstrated a leaky wave antenna using a dielectric waveguide [10] based on the grating type leaky wave antenna created for millimeter wave integrated circuit. They altered the characteristics of the waveguide by altering the cross sectional area of the dielectric material.
In 1985, Jackson and Alexopoulos demonstrated that using materials with either permittivity, $\varepsilon$, or permeability, $\mu$, greater than 1 helped in increasing the power gain as well as narrow the beam width\[11\]. The power gain varied proportional to either $\varepsilon$ or $\mu$ depending upon the configuration. They also reported an observation that the bandwidth varied inversely to the gain.
They used two materials with different electric and magnetic properties. As shown in fig 6., material 2 with dielectric constant $\varepsilon_2$ was placed on top of material 1 which had a dielectric constant $\varepsilon_1$. The materials were so chosen that $\varepsilon_2 > \varepsilon_1$.

From fig 7, it can be observed that the gain was proportional to the difference between the dielectric constants of the two materials. The gain observed increased as the difference increased.

The beamwidth became narrow, as shown in fig 8, when the difference between the dielectric constants increased.

Dwelling on the same lines, Jackson along with Oliner and Ip came up with a multi layered dielectric structure that produced narrow beams [12]. They alternated with two
dielectric materials with different thickness and electrical characteristics as shown in fig 9. The narrow beamwidth was due to leaky waves which were attenuated very weakly. They also demonstrated the variation of beamwidth with reference to the number of layers and the material properties.

Fig 1.7. Gain vs $\varepsilon_2$. 
Fig 1.8. Beamwidth vs $\varepsilon_2$.

Fig 1.9. Narrow beam structure consisting of dielectric layers of alternate thickness and electrical properties. – Jackson, Oliner, Ip
In 1991 Guglielmi and Boccalone[12] came up with a new concept of placing metal strips periodically over a dielectric waveguide. Through their study they demonstrated that the electrical characteristics of this antenna are flexible.

Caloz and Itoh in 2003 showed the use of metamaterials in leaky wave antennas [14]. They called this a Backfire-to-Endfire leaky wave antenna – a direct application of Composite Right/Left Handed Transmission Line (CRLH-TL). This antenna in at its transition frequency in its fundamental mode of operation could scan both backward and forward. They demonstrated a scanning range between 3.1 and 6.0 GHz.
In this research work, an attempt would be made to analyze the transmission losses happening at the signal generator and antenna feed and find a solution to minimize the loss. This document is structured to give an introduction to the principle working operation of a leaky wave antenna in Chapter 1, design and simulation of the antenna in Chapter 2, fabrication and experiments in Chapter 3, and conclusion and further research in Chapter 4.
2. Modelling and Simulation

2.1 Theory of leaky wave antenna

A leaky wave antenna operates under the principle of a travelling wave leaking power along the length of the waveguide. The antenna can be designed to either radiate in a certain direction or to scan over a range of angles. The main beam of the leaky wave antenna can be changed by changing the operating frequency.

There are two basic types of leaky wave antennas: Uniform Leaky Wave Antennas and Periodic Leaky Wave Antennas [20].

2.1.1 Design principle of a Uniform Leaky Wave Antenna

A uniform leaky wave antenna is uniform along the length of the structure except for a small taper that helps to control and improve the sidelobes. In principle, uniform leaky wave antennas radiate into the forward quadrant and can scan from broadside to end-fire. In reality, how close the antenna can scan to broadside or end-fire depends on the structure of the antenna.

Uniform leaky wave antenna consists of a leaky waveguide with a length L. The leakage occurs along the length of the antenna. In the longitudinal direction, z, the propagation characteristics of the leaky mode are given by phase constant β and leakage constant α, where α is a measure of the power leaked (radiated) per unit length. The length of the antenna, L, forms the aperture of the line-source antenna and the amplitude and phase of the travelling wave along the aperture are determined by the values of β and α as a function of z. As long as the waveguide is uniform along its length without any taper, the
values of $\beta$ and $\alpha$ do not change with $z$. This would also result in the aperture distribution having an exponential amplitude variation and a constant phase due to which the sidelobe levels would be high. In order to manage the sidelobe level, $\alpha$ has to have a variation with $z$.

Beam direction and beamwidth of a uniform leaky-wave antenna are defined by:

$$\sin \theta_m \approx \frac{\beta}{k_0} \quad \text{Eqn 2.1}$$

$$\Delta \theta \approx \frac{1}{\left(\frac{L}{\lambda_0}\right) \cos \theta_m} \quad \text{Eqn 2.2}$$

Where,

- $\theta_m$ is the maximum beam angle
- $\beta$ is the phase constant
- $k_0$ is the free space wave number
- $L$ is the length of the antenna
- $\Delta \theta$ is the beamwidth of the leaky-wave antenna

As can be seen from Eqn 2.2, the beamwidth of the leaky-wave antenna is primarily determined by the length, $L$, of the antenna. It is also slightly influenced by the aperture field amplitude distribution. The beamwidth is narrowest for a constant aperture field and wider for sharply peaked distributions.

For a given value of $\alpha$, if the antenna length is chosen such that 90 percent of the power is radiated, it can be deduced that:

$$\frac{L}{\lambda_0} \approx \frac{0.18}{\left(\frac{\alpha}{k_0}\right)} \quad \text{Eqn 2.3}$$
2.1.2 Design principle of a Periodic Leaky Wave Antenna

The major difference between a periodic leaky-wave antenna and a uniform leaky-wave antenna is that the waveguide is modulated periodically along its length in a periodic leaky-wave antenna. The periodic modulation produces the leakage.

Fig 2.1 shows a uniform waveguide with an array of metal strips placed periodically along its length. Before the array of metal strips are added, the dimensions of the guide and the frequency are chosen such that $\beta > k_0$. The periodic array of strays introduces space harmonics because of the periodicity. Each of the space harmonics are characterized by phase constant $\beta_n$. The space harmonics are related to each other by:

$$\beta_n d = \beta_0 d + 2n\pi$$  \hspace{1cm} Eqn 2.4

where $d$ is the period; $\beta_0$ is the fundamental space harmonics, the original $\beta$ of the dominant mode of the uniform dielectric waveguide. $\beta_n$ can take on a large variety of values so that the space harmonics be either forward or backward in nature, and be fast or slow.
For a space harmonic to be fast, $\beta_0/k_0$ should be less than 1. Since we need only a single radiated beam, we can choose $n = -1$.

If we replace $\beta$ by $\beta_{-1}$ in equation 2.1, the beam direction of the antenna can be defined as:

$$\sin \theta_m \approx \frac{\beta_{-1}}{k_0} \quad \text{Eqn 2.5}$$

where

$$\beta_{-1} = \beta_0 - \frac{2\pi}{d} \quad \text{Eqn 2.6}$$

Using equation 2.6 in 2.5,

$$\sin \theta_m \approx \frac{\beta_0 - \frac{2\pi}{d}}{k_0} = \frac{\lambda_0}{\lambda_{g0}} - \frac{\lambda_0}{d} \quad \text{Eqn 2.6}$$

Hence, depending on how ($\lambda_0/d$) compares with ($\lambda_0/\lambda_{g0}$), the beam can point in the backward quadrant or the forward quadrant.
2.2 Design of leaky wave antenna

A leaky wave antenna was built and simulated using the simulation tool from Agilent – now called Keysight Technologies, Electromagnetic Professional. EMPro is a 3-dimensional software which supports both Finite-Difference Time Domain (FDTD) and Finite Element Method (FEM).

FEM is used for the computation and simulation here as FDTD poses certain restrictions like having extremely small grid in order to satisfy very small electromagnetic wavelength which led to extremely high computation time period. EMPro also poses a restriction for the usage of FDTD – Continuous visualization of electromagnetic field visualizations cannot be obtained using EMPro.

Radiation patterns for both near field and far field can be obtained in 3-dimensional pattern when simulating using FEM.

The basic structure of the leaky wave antenna is as shown in figure 2.2. The slot width (SW), length (SL), and distance (SD) are 1.0mm, 6.0mm, and 7.0mm, respectively. The width, length, and height of the waveguide are WW=8.0mm, LW=96.0mm, and HW=1.7mm, respectively. The figure also shows a cross sectional view of the leaky wave antenna formed by a piece of printed circuit board and PZT material.
Figure 2.2: (a) Top view of a rectangular waveguide leaky wave antenna; (b) Cross-sectional view of the rectangular waveguide leaky wave antenna. The inserted PZT film has a thickness of 0.3mm, and a relative permittivity of $\sim 1,900$

The 3-dimensional modeled structure of the antenna which was constructed for the simulation and computational purposes is shown in figure 2.3.

Figure 2.3 3-dimensional model of leaky wave antenna built using EMPro

According to the principle of operation of a leaky wave antenna, the radiation occurs through the uniformly spaced holes on the top plate of the antenna when there is a travelling wave propagating through it. From the basic principle of rectangular waveguides, for a wave to propagate through it, the frequency of the wave has to be
higher than the operational cut-off frequency. The cut-off frequency is dependent on the height and width of the waveguide and is given by:

$$f_{\text{cut-off}} = \frac{1}{2\pi\sqrt{\mu\varepsilon}} \sqrt{\left(\frac{m\pi}{WW}\right)^2 + \left(\frac{n\pi}{HW}\right)^2}$$  \hspace{1cm} \text{Eqn 2.7}$$

The propagation wavelength for the travelling wave can be obtained from the above equation for cut-off frequency as:

$$\lambda = \frac{2\pi}{\sqrt{\omega^2 \mu \varepsilon - \left(\frac{m\pi}{WW}\right)^2 - \left(\frac{n\pi}{HW}\right)^2}}$$  \hspace{1cm} \text{Eqn 2.8}$$

where $m$ and $n$ are integers, $\varepsilon$ is the dielectric constant and $\mu$ is the magnetic permeability of the material inside the waveguide.

\subsection*{2.2.1 Wave propagation in waveguide with high dielectric constant}

From Eqn 2.7, it can be seen that the dielectric constant is inversely related to the cut off frequency. The cut-off frequency depends on the dielectric constant: larger the value of $\varepsilon$, smaller the cut-off frequency. For examples, Figs 2.4 and 2.5 depict the wave propagation in an air-core rectangular waveguide at 10.0 GHz and 30.0 GHz with $WW = 8.0$ mm, $HW = 1.7$ mm, and $LW = 96.0$ mm. The thickness of the copper layer is 0.1 mm.

Fig 2.4 Evanescent wave with air as dielectric at 10.0 GHz
Analyzing Eqn. 2.8 also brings forth the relationship between the dielectric constant of the material inside the waveguide and the propagation wavelength. From the equation, it can be observed that a higher value of the dielectric material would result in a shorter propagating wavelength.

To understand the impact of various factors on the operation of the leaky wave antenna, various simulations were conducted. All the simulations were run using a 3-dimensional (3D) electromagnetic computation software Electromagnetic Professional - EMPro. The software is from Keysight Technologies which was previously called Agilent Technologies. EMPro has two options for simulating the antenna - Finite-Difference Time-Domain Method (FDTD) and Finite Element Method (FEM). FDTD requires the grid to be sufficiently small to satisfy the smallest electromagnetic wavelength and the physical length, which would consume very high computing time. Also a limitation of EMPro is that continuous visualizations of the inner waveguide cannot be obtained using FDTD. FEM, on the other hand, generates continuous visualization pattern in 3-dimensional view. Based on the above reasoning, the simulation has been done using FEM methods.
2.2.2 Effect of dielectric on the radiation pattern

To get the main radiation lobe in the vertical direction, the propagating wavelength of the wave and the slot distance of the antenna should match. That would imply that antennas with higher values of dielectric materials would need to be operated at lower frequencies. Simulations were run by giving two different values to the dielectric inside the waveguide of the leaky wave antenna. Rest of the properties of the antenna were held constant.

The first simulation was run by having the dielectric constant set to 337. The peak radiation was observed around 2.65 GHz.

The antenna was next simulated with a dielectric constant set to 813. The peak radiation in this case was observed to be around 2.43 GHz, as shown in Fig 2.6. This simulation experiment demonstrated that a material with high dielectric value lowers the central operating frequency of the leaky wave antenna.
2.2.3 Effect of distance between adjacent slots in the antenna

In this set of simulations, the total length of the antenna was held constant. The distance between the adjacent slots was varied from 6.4 mm to 16 mm. This resulted in changing the number of slots in the antenna from 6 to 15. The peak amplitude was observed to be between 2.4 GHz and 2.6 GHz for all the different configurations. The antennas with more number of slots in them, had a peak operational frequency between 2.4 GHz and 2.5 GHz while the antennas with less number of holes and more distance between adjacent holes had peak operational frequency around 2.6 GHz.
Changing the distance between adjacent slots in the antenna resulted in changing the angle of main beam radiation.

Figure 2.7 Distance between slots are varied.
3. Antenna Fabrication and Measurement

3.1 Fabrication

The leaky wave antenna that was used for the experimental measurements consisted of a copper waveguide, as shown in Fig 3.1, and a top plate which had periodic holes, as shown in Fig 3.2.

The copper waveguide was built using printed circuit board. The external dimensions of the waveguide are:

- Width of the waveguide (WW) = 9 mm
- Length of the waveguide (LW) = 107 mm
- Height of the waveguide (HW) = 4 mm

The top plate was also built using printed circuit board with copper peeled off completely. It is an established fact that higher radiation gain can be achieved by having a narrow slot length. In order to have maximum radiation gain, the slot length was determined to be 1 mm. The top plate covers the entire length of the waveguide as shown in Fig 3.3. The dimensions of the top plate are:

- Width = 8 mm
- Length = 100 mm
- Slot Length = 1 mm
- Slot Width = 8 mm
- Slot Distance = 4.8 mm
In order to feed the waveguide, an SMA connector was soldered on one end as shown in Fig 3.4.

Lead Zirconate Titanate, PZT, shown in Fig 3.5, which is #0027 on American Piezo Ceramic, Inc. having a high dielectric value of 1900 is used as the dielectric. Based on the research on the effective medium theory on piezoelectric composites, the effective dielectric constant of a multilayer is equal to the average dielectric constant of the entire multilayer, as long as the thickness of each layer is smaller than the propagation wavelength. The experiments were conducted with two values of $\varepsilon_r$: 722 and 813. Two PZT strips with width of 8 mm and thickness 0.76 mm were used to get an effective $\varepsilon_r$ of 722. For $\varepsilon_r$ of 813, in addition to the above two strips, another strip of width 2 mm and thickness of 0.76 mm was used.

![Copper Waveguide](image)

**Fig 3.1 Copper Waveguide – Part of the leaky wave antenna**
Fig 3.2 Top Plate of leaky wave antenna

Fig 3.3 Leaky Wave Antenna – Complete with Waveguide and top plate

Fig 3.4 Waveguide soldered with SMA Connector at one end
Fig 3.5 PZT Strips used

Fig 3.6 PZT Strips arranged to get effective value of 722

Fig 3.7 PZT Strips arranged to get effective value of 813
AUTOCAD was used to design the antenna layouts. The designs were converted to DXF format to use in the milling machine. Two software’s, Board Master and Circuit Cam, associated with the milling machine to mill the antennas on the printed circuit board.

### 3.2 Measurement Set up

The experimental measurement set was made up of an Agilent RF Signal Generator – N9310A, an Agilent Spectrum Analyzer – N9320B, the leaky wave antenna, and a handmade horn antenna which acted as the receiver. The horn antenna was essentially a dipole antenna enclosed in a horn shaped cavity as shown in Fig 3.9.
Fig 3.9 Handmade horn antenna consisting of a dipole antenna enclosed in a horn shaped cavity

The spatial angular resolution was found to be around 20 degrees. Between the Agilent RF Signal Generator and the Leaky Wave Antenna, a RF power amplifier, Hittite 12170 – 1, is used to amplify the signal. The receiver antenna, the horn antenna, is connected to the spectrum analyzer.
Fig 3.10 Experimental set up

Fig 3.11 Actual Measurement Set up
3.3 Measurements

Measurements are made by changing the position of the receiver antenna, as shown in Fig 3.10. The minimum angular separation between two measurement positions is 10 degrees. This gives us a measurement range between -80 to +80 degrees. The measurements were made at fixed frequencies between 2.1 GHz and 2.6 GHz. The frequency range was chosen to excite only the fundamental mode $TE_{10}$.

3.3.1 Measurement with $\varepsilon_r = 722$

Measurements were made with the dielectric constant, $\varepsilon_r$, set to 722. The experiment was conducted at 2.3527 GHz, 2.4027 GHz, 2.4527 GHz, 2.5027 GHz, and 2.5527 GHz. The results, gain against the angle of measurement, are shown in the below figures.

![Fig 3.12 f= 2.3527 GHz](image)
Fig 3.13 $f = 2.4027$ GHz

Fig 3.14 $f = 2.4527$ GHz
The main beam radiation, as observed by the measurements, is between -20 to +30 degrees. The measurements were performed over a frequency span of 200MHz. The change of angle per frequency, $\eta$, from the above data is found to be $0.25^\circ$/MHz.
3.3.2 Measurement with $\varepsilon_r = 813$

Measurements with the leaky wave antenna set up was repeated for an effective dielectric value, $\varepsilon_r$, of 813. Since the value of the dielectric used is higher than the previous experiment, with the rest of the set up being unchanged, the operating frequency was reduced to get maximum possible gain. The measurements were performed over a span of 200 MHz with the specific frequencies being 2.1027 GHz, 2.1527 GHz, 2.2027 GHz, 2.2527 GHz, and 2.3027 GHz.

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>Main beam direction</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.3527</td>
<td>-20°</td>
</tr>
<tr>
<td>2.4027</td>
<td>0°</td>
</tr>
<tr>
<td>2.4527</td>
<td>-10°</td>
</tr>
<tr>
<td>2.5027</td>
<td>10°</td>
</tr>
<tr>
<td>2.5527</td>
<td>30°</td>
</tr>
</tbody>
</table>

Table 3.1: Frequency and Main beam direction for $\varepsilon_r = 722$

Fig 3.17 f = 2.1027 GHz
Fig 3.18 $f = 2.1527$ GHz

Fig 3.19 $f = 2.2027$ GHz
Fig 3.20 $f = 2.2527$ GHz

Fig 3.21 $f = 2.3027$ GHz

From the measurements it is observed that the main beam radiates between $-30$ to $+30$ degrees for a frequency span of $200$MHz. The change of angle per frequency is $0.3^\circ$/MHz.
<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>Main beam direction</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.1027</td>
<td>-30°</td>
</tr>
<tr>
<td>2.1527</td>
<td>-30°</td>
</tr>
<tr>
<td>2.2027</td>
<td>0°</td>
</tr>
<tr>
<td>2.2527</td>
<td>30°</td>
</tr>
<tr>
<td>2.3027</td>
<td>30°</td>
</tr>
</tbody>
</table>

Table 3.2: Frequency and Main beam direction for $\varepsilon_r = 813$

### 3.3.3 Measurement data at 0°

Data collected from the above measurements were tabulated to compare the gain at 0° on the measurement setup, i.e., in front of the transmitter antenna. The measurements were conducted between 2.12 GHz and 2.27 GHz.

For $\varepsilon_r = 722$, the gain was highest around 2.24 GHz while for $\varepsilon_r = 813$, the highest gain was observed around 2.16 GHz. This experiment reaffirmed that higher permittivity lowers the central operating frequency.

Fig 3.22 shows the gain against central operating frequency for $\varepsilon_r = 722$.

Fig 3.23 shows the gain against central operating frequency for $\varepsilon_r = 813$. 
Fig 3.22: Gain vs frequency for $\varepsilon_r = 722$
3.4 Comparison between experimental and simulated results

The experimental results and the simulated results correlate with each other to a very large extent. Figures 3.24 to 3.27 show the comparison between the two sets of results at 4 distinct frequencies: 2.40 GHz (Fig 3.22), 2.45 GHz (Fig 3.23), 2.5 GHz (Fig 3.24), and 2.6 GHz (Fig 3.25) all with dielectric constant, \( \varepsilon \), set to 722. Fig 3.28 compares the radiation main beam angle between the experimental
results and simulation results. Table 3.3 tabulates the results from the experiments and simulations. The differences between the results could be attributed to the fabrication of the antenna.

Fig 3.24 Experiment vs Simulation, $\varepsilon = 722$ and $f = 2.40$ GHz
Fig 3.25 Experiment vs Simulation, $\varepsilon = 722$ and $f = 2.45$ GHz
Fig 3.26 Experiment vs Simulation, $\varepsilon = 722$ and $f = 2.50$ GHz
Fig 3.27 Experiment vs Simulation, $\varepsilon = 722$ and $f = 2.60$ GHz
Fig 3.28: Experiment vs Simulation: Main Beam Angle
### Table 3.3: Comparison between experimental and simulated results

<table>
<thead>
<tr>
<th>Dielectric constant $\varepsilon$</th>
<th>722 (simulation)</th>
<th>722 (simulation)</th>
<th>813 (simulation)</th>
<th>813 (simulation)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Angles</td>
<td>-38$^\circ$ to 35$^\circ$</td>
<td>-20$^\circ$ to 42$^\circ$</td>
<td>-30$^\circ$ to 30$^\circ$</td>
<td></td>
</tr>
<tr>
<td>Change of angle</td>
<td>73$^\circ$</td>
<td>50$^\circ$</td>
<td>72$^\circ$</td>
<td>60$^\circ$</td>
</tr>
<tr>
<td>Operating frequencies range</td>
<td>2.4 to 2.575 GHz</td>
<td>2.3527 to 2.5027 GHz</td>
<td>2.12 to 2.23 GHz</td>
<td>2.1027 to 2.3027 GHz</td>
</tr>
<tr>
<td>Change of frequency</td>
<td>175 MHz</td>
<td>200 MHz</td>
<td>110 MHz</td>
<td>200 MHz</td>
</tr>
<tr>
<td>Change of angle per frequency ($\eta$)</td>
<td>0.41$^\circ$/MHz</td>
<td>0.25$^\circ$/MHz</td>
<td>0.65$^\circ$/MHz</td>
<td>0.3$^\circ$/MHz</td>
</tr>
</tbody>
</table>

### 3.5 Matching Network

During experimentation, it was noticed that connecting the antenna to the signal generator resulted in losses. An attempt was made to analyze the amount of loss occurring and to try and build a matching network to minimize the losses. Voltage standing wave ratio was calculated to analyze the loss. To calculate the reflection coefficient, the receiver antenna was connected to a network analyzer in order to collect S11 data. VSWR was calculated using the equation:

$$VSWR = \frac{1 + |\Gamma|}{1 - |\Gamma|}$$
where $\Gamma$ is the S11 or reflection coefficient.

The standing wave ratio, VSWR, was calculated at transmitting frequencies ranging from 2 GHz to 2.28 GHz. The VSWR for these frequencies was calculated to be almost equal to 1 resulting in very high losses.

To minimize the losses, a RF Power Amplifier, HMC758LP3, was connected between the signal generator and the transmitting antenna. The receiver antenna was again collected to the network analyzer to collect S11 data to calculate VSWR.

The VSWR calculated based on this new S11 data showed marginal improvement over the previous calculation. The high value of the dielectric for PZT results in very low radiated power. Having a high relative permittivity material inserted in the waveguide resulted in designing a leaky wave antenna with high angular scan rate, low operating frequency, and narrow bandwidth but with huge loss in radiated power.
4. Conclusion and Future work

4.1 Conclusion

In the course of this research, a portable leaky wave antenna with high relative permittivity material inserted in the waveguide was designed, built, and demonstrated. The effects of having high permittivity material in building high sensitive, large angle scanning leaky wave antenna are:

- Achieve high angular scan rate
  - For $\varepsilon_{r\text{-effe}} = 337$, ASR = 273º/GHz
  - For $\varepsilon_{r\text{-effe}} = 813$, ASR = 636/GHz
- Lowers the central operating frequency
- Narrows down the radiation bandwidth

<table>
<thead>
<tr>
<th>Permittivity ($\varepsilon_r$)</th>
<th>1</th>
<th>2.55</th>
<th>337</th>
<th>813</th>
</tr>
</thead>
<tbody>
<tr>
<td>Scanning angle (º)</td>
<td>-65 ~ -35</td>
<td>10 ~ 50</td>
<td>-32 ~ 50</td>
<td>-30 ~ 40</td>
</tr>
<tr>
<td>Operating frequency (GHz)</td>
<td>15 ~ 18</td>
<td>8.1 ~ 9.1</td>
<td>2.4 ~ 2.7</td>
<td>2.12 ~ 2.23</td>
</tr>
<tr>
<td>ASR (º/GHz)</td>
<td>10</td>
<td>44</td>
<td>273</td>
<td>636</td>
</tr>
</tbody>
</table>

Table 4.1: Comparison of measured LWAs of having $\varepsilon_r = 337$ and $\varepsilon_r = 813$
• The effect of matching networks was also proven with the use of a RF Power Amplifier.

4.2 Future work

The biggest trade-off of achieving high angular scan rate and lower operating frequency by using high permittivity material in a leaky wave antenna is losing the radiation power. Further investigation needs to be made on designing matching network using a power amplifier having much higher gain than HMC758LP3. Operating the antenna at much higher frequencies could result in obtaining better angular scan rates and narrower bandwidths.
References:

1. N. Marcuvitz: On field representation in terms of leaky modes or eigenmodes, IRE Trans., AP-4, 3 (July 1956).


