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Compact Leaky Wave Antenna Using Ferroelectric Materials

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

By

Hyung Min Jeon
B.S.N.E., Chosun University, 2007

2012
Wright State University

WRIGHT STATE UNIVERSITY

GRADUATE SCHOOL

December 13, 2012

I HEREBY RECOMMEND THAT THE THESIS PREPARED UNDER MY SUPERVISION BY Hyung Min Jeon ENTITLED Compact Leaky Wave Antenna Using Ferroelectric Materials BE ACCEPTED IN PARTIAL FULFILLMENT OF THE REQUIREMENTS FOR THE DEGREE OF Master of Science in Engineering.

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ABSTRACT

Hyung Min Jeon. M.S.Egr Department of Electrical Engineering, Wright State University, 2012. Compact Leaky Wave Antenna Using Ferroelectric Materials.

Leaky wave antenna continuously attracts a lot interests for tuning the radiation direction over a large angle. Various types of leaky wave antenna have been developed since 1940. However, most of these leaky wave antennas suffered from their excessive volume and the required wide range of operating frequency (bandwidth) to achieve a large tuning angle. To circumvent these challenges, research in this thesis is focused on implementation of material (like PZT) with high dielectric constant (k) in a rectangular waveguide leaky wave antenna. Due to the high dielectric constant of PZT $\sim 1,900$, the propagation wavelength in the waveguide can be dramatically shortened. This allows the leaky wave antenna to be operated at much lower frequency, or to have a significant volume reduction. In addition, using high k material enables bandwidth narrowing. In experiments, 70 degree tuning angle was observed by changing frequency from 2.1217 GHz to 2.235 GHz in a proto-type leaky wave antenna filled with material of $k = 813$. The tuning angle can be further adjusted by the slot length and distance.

TABLE OF CONTENTS

Chapter 1: INTRODUCTION.....	1
1.1: Introduction.....	1
1.2: History.....	3
Chapter 2: DESIGN AND MODELING.....	13
2.1: Basic structure of waveguide.....	13
2.2: Wave propagation in waveguide with high dielectric constant.....	15
2.3: Design of leaky wave antenna.....	18
2.3.1: Impact on the slot distance.....	18
2.3.2: Impact on the slot length.....	22
2.3.3: Impact on the slot length.....	23
Chapter 3: DEVICE FABRICATION AND CHARACTERIZATION.....	27
3.1: Device fabrication.....	27
3.2: Characterization of leaky wave antennas.....	31
3.2.1: Measurement set-up.....	31
3.2.2: Leaky wave antenna with inserted dielectric material $\epsilon = 722$	32

3.2.3: Leaky wave antenna with inserted dielectric material $\epsilon = 813$	35
3.3 Comparison between simulation and measurement.....	38
Chapter 4: CONCLUSIONS.....	40
4.1: Summary.....	40
4.2: Future works.....	40
Bibliography.....	41

LIST OF FIGURES

Fig 1.1: first example of leaky wave antenna (L.O. Glodstone and A.A. OLINER, 1959).	3
Fig 1.2: full views of asymmetric structure (W.ROTMAN and A.A OLINER, 1959).....	4
Fig 1.3: holey waveguide (L.O.Glodstone and A.A. OLINER, 1959).....	5
Fig 1.4: Trapped image guide antenna (a) and Cross section of trapped image guide (b) (T. Itoh and B. Adelseck, 1980).....	6
Fig 1.5: Superstrate-substrate geometry (DAVID R. JACKSON and NICOLAOS G. ALEXOPOULOS, 1985).....	7
Fig 1.6: a) Gain vs. dielectric constant of top layer ϵ_2 . b) Beam width vs. dielectric constant of top layer ϵ_2 . (DAVID R. JACKSON and NICOLAOS G. ALEXOPOULOS, 1985).....	8
Fig 1.7: a metal-strip-loaded dielectric leaky-wave antenna (Min Chen, Bijan Houshmand, and Tatsuo Itoh, 1997).....	8
Fig 1.8: narrow-beam multiple layers structures (D.R. Jackson, A.A. Oliner, and Antonio, 1993).....	9
Fig 1.9: ferroelectric substrate leaky-wave antenna (YevhenYashchyshyn and Jozef Modelski, 2002).....	10

Fig 1.10: Leaky wave antenna with meta-material (Christophe Caloz, and Tatsuo Itoh, 2003).....	11
Fig 2.1: The 3D modeling structure of leaky wave antenna (a) and detail parameters (b)	14
Fig 2.2: Evanescence wave in empty (dielectric constant =1) waveguide at 13.5 GHz...	15
Fig 2.3: travelling wave in empty (dielectric constant =1) waveguide at 25 GHz.....	15
Fig 2.4: travelling wave in dielectric constant 10 material waveguide at 13.5 GHz.....	16
Fig 2.5: travelling wave in dielectric constant 10 material waveguide at 25 GHz.....	16
Fig 2.6: no travelling wave in dielectric constant 10 material waveguide at 2 GHz.....	17
Fig 2.7: travelling wave in dielectric constant 722 material waveguide at 2 GHz.....	17
Fig 2.8: Radiation pattern at 2.4GHz (a), 2.475GHz (b), and 2.575 GHz (c).....	19
Fig 2.9: Radiation pattern with various slot distance 4.7mm (a), 4.8mm (b), and 4.9mm (c).....	20
Fig 2.10: Radiation patterns with slot distance 9.5238mm at 2.4 GHz (a), 2.475 GHz (b), and 2.575 GHz (c).....	22
Fig 2.11: Scanning radiations with dielectric constant 362 at 3.34286 GHz (a), 3.51429 GHz (b), and 3.64286 GHz (c).....	24

Fig 2.12: Scanning radiations with dielectric constant 813 at 2.225 GHz (a), 2.33125 GHz (b), and 2.4375 GHz (c).....	25
Fig 2.13: main radiation direction patterns with various dielectric constant materials....	26
Fig 3.1: Designed waveguide by printed circuit board.....	28
Fig 3.2: Connected one port to waveguide.....	28
Fig 3.3: Calculated dielectric constant $\epsilon = 722$ (a) and dielectric constant $\epsilon = 813$ (b).....	29
Fig 3.4: PZT films to construct $\epsilon = 722$ (a) and $\epsilon = 813$ (b).....	29
Fig 3.5: Fabricated top plate.....	30
Fig 3.6: Assembled leaky wave antenna (a) and inserted the material the cross view(b).....	30
Fig 3.7: Hand-made horn antenna (a) and special angular resolution (b).....	31
Fig 3.8: Setup diagram.....	32
Fig 3.9: Main radiation direction patterns with $\epsilon = 722$ by increasing frequencies 2.323 GHz (a), 2.353 GHz (b), 2.489 GHz (c), and 2.54 GHz (d).....	35
Fig 3.10: Main radiation direction patterns with $\epsilon = 813$ by increasing frequencies 2.1217 GHz (a), 2.1588 GHz (b), 2.2217 GHz (c), and 2.235 GHz (d).....	38
Fig 3.11: Comparison between simulation and measurement for $\epsilon = 722$ (a) and $\epsilon = 813$ (b).....	39

List of Tables

Table 1: Radiation direction and power gain in 2.4GHz, 2.475GHz, and 2.575GHz.....	19
Table 2: Radiation direction and power gain vs. slot distance.....	21
Table 3: Radiation direction and power gain for $SD=2\lambda$	22
Table 4: Beam Width and Gain vs. Slot Length.....	23
Table 5: Radiation direction and power gain for dielectric constant 362.....	24
Table 6: Radiation direction and power gain for dielectric constant 813.....	26
Table 7: Change of angle per frequency with $\epsilon = 362$, $\epsilon = 722$, and $\epsilon = 813$	26
Table 8: Compare between simulation and measurement with $\epsilon = 722$ and $\epsilon = 813$	38

ACKNOWLEDGEMENT

First of all, I would like to thank Dr. Yan Zhuang, my thesis director. He supports and guides through this whole thesis. His experience, research, and great idea help me to complete this thesis. Also, I would like to thank Dr. Kefu Xue and Dr. Kuan-Lun Chu who were the thesis committee members.

Secondly, In Joo Jeon and Jung Nim Choi encouraged and supported me to overcome entire hard situation all the time. Without my parent, I would not succeed to complete my thesis. Also, I like to thank Lucy Fu. She takes care and understands me while I work this research.

Lastly, I like to thank my neighbors, friends, and colleagues.

Compact Leaky Wave Antenna Using Ferroelectric Materials

Chapter 1- Introduction

1.1 Introduction

Leaky-wave antenna consisting of a number of leaky waves with constant phase differences exhibits frequency controllable radiation direction. The concept has been demonstrated using a slotted rectangular waveguide by W. W. Hansen in 1940, which is similar as phase array antenna. Instead of electrically/magnetically tuning the phase shift in phase array antenna, leaky wave antenna modifies the phase by changing the frequencies.

The major advantages of phase array antenna are the quick response and high radiation power gain. However, in order to achieve continuous angle scanning and coordinate phase difference for maximum power gain, a large number of phase shifters are required and needed to be adjusted simultaneously. This imposes formidable challenges in making compact, low cost, and energy efficient phase array antennas. The large number of phase shifters increases the volume and fabrication cost dramatically. In addition, tuning of the large number of phase shifter array consumes significant energy.

Alternative approach capable of large angle scanning is the so-called leaky wave antenna. Compared with phase array antenna, leaky wave antenna has a number of advantages: 1) Leaky wave antenna only needs the frequency modulation for scanning beam angle. This allows fast response; 2) generally leaky wave antenna is based on simple wave guide structure, assuring low cost manufacture; 3) leaky wave antenna does not need complicate control circuit as required in phase array antenna to coordinate the phase difference; 4) the simplified control circuit and less number of components (e.g. eliminating of phase shifter) in leaky wave antenna reduce significantly the energy consumption. However, the most popular leaky wave antenna suffers from the large size on one hand and large frequency change to reach large angle scan. In this work, I propose to implement ferroelectric materials with high permittivity in leak wave antenna to circumvent the challenges. The high permittivity effectively reduces the wave propagation wavelength, consequently scale down the antenna size. In addition, the high permittivity helps achieve the large scan angle and narrow down the operating bandwidth. This is specific important for a point to point communication because of the required narrow operating bandwidth. For example, Wireless LAN by IEEE 802.11 has bandwidth around 0.05GHz between 2.412 to 2.462 GHz in North America. Also, the bandwidth f or Bluetooth is around 0.4GHz between 2.4 to 2.48GHz. Beyond the operating bandwidth, devices cannot function well in the communication.

1.2 History

The first leaky wave antenna (shown in Fig. 1.1) was the slotted rectangular waveguide by W. W. Hansen in 1940 [7].

Figure 1.1 first example of leaky wave antenna (L.O. Glodstone and A.A. OLINER, 1959)

This antenna has a long slot on the side of a rectangular waveguide, from which the travelling electromagnetic wave leaks out. Generally, leaky wave antenna follows this expression.

$$\sin \theta = \frac{\lambda_0}{\lambda} - \frac{\lambda_0}{d} \quad \text{Eq.(1-1)}$$

where

θ is the angle counted from direction perpendicular to wave propagation;

λ_0 is the wavelength in air;

λ is the wavelength inside the waveguide;

d is perturbation space;

m is an integer.

Another example is an asymmetrical waveguide antenna [1]. W. Rotman and A. A. Oliner tried to design a waveguide with two symmetrical spaces using middle center fin in 1959. However, they did not observe any leaky radiation because the symmetrical structure functioned as a short circuit. As the result, the symmetrical structure dissipates all the power in the structure. Later, they redesigned an asymmetrical antenna, as shown in fig. 1.2. The unbalanced inner space helped the leaky wave radiation. This observation explains that the open wall of the asymmetric structure could generate radiation.

Figure 1.2 full views of asymmetric structure (W.ROTMAN and A.A OLINER, 1959).

In 1954, Stegen and Reed [14] replaced the long slot by an array of small round holes with a short interval to enhance the radiation power and narrow down the radiated beam (Fig. 1.3). However both the radiated power and beam width are not good enough for practical application. In order to achieve higher power gain and large scanning angle in

leaky wave antenna, people have spent tremendous efforts in trying various ideas [11][8][9]. Using dielectric material is one of the solutions.

Figure 1.3 holey waveguide (L.O.Glodstone and A.A. OLINER, 1959)

Figure 1.4 Trapped image guide antenna (a) and Cross section of trapped image guide (b)
(T. Itoh and B. Adelseck, 1980)

T. Itoh and B. Adelseck demonstrated leaky wave antenna using dielectric material in 1980 [17], by inserting dielectric material into the waveguide as shown in figure 1.4. By varying the cross-sectional area filled by dielectric material, the waveguide characteristic were changed. They discussed various leaky wave antenna structures including fully filled, and partially filled of dielectric materials in the waveguides. In their experiments, the scanning angle of the main beam reached from -50 degree to -10 degree.

In the middle of 1980s, Alexopoulos and Jackson reported in their work that using dielectric material (Fig. 1.5) to improve the radiation power gain and narrow down the beam width [6]. Figure 1.6(a) shows the gain was getting increased by choosing high ϵ_2 of the top layer, e.g. When the dielectric constant ϵ_2 changed from 0 to 90, the gain was increased from 6 to 20 dB. The correlation between the beam width and the dielectric constant is shown in figure 1.6 (b). By properly choosing the dielectric constant of the bottom layer ($\epsilon_1 \sim 12.5$), the beam angle varied from 80 to 160 degree.

Figure 1.5 Superstrate-substrate geometry (DAVID R. JACKSON and NICOLAOS G. ALEXOPOULOS, 1985)

Figure 1.6 a) Gain vs. dielectric constant of top layer ϵ_2 . b) Beam width vs. dielectric constant of top layer ϵ_2 . (DAVID R. JACKSON and NICOLAOS G. ALEXOPOULOS, 1985)

In 1990s, the concept of metal-strip dielectric leaky wave antenna was studied in [12][13][16]. In general, the leaky wave antenna was made of periodic metal strips on the dielectric waveguide, as shown in figure 1.7. It turned out that by utilizing Duroid material ($\epsilon_r = 2.33$), this type of antenna achieved proper gain with large scanning beam angle (30 degree). However, the operation frequency bandwidth is a little large from 75 to 85 GHz.

Figure 1.7 a metal-strip-loaded dielectric leaky-wave antenna (Min Chen, Bijan Houshmand, and Tatsuo Itoh, 1997)

Encouraged by these results, tremendous efforts have been spend on investigation of the correlation between dielectric material and radiation beam width. In 1993, D. R. Jackson et al designed narrow beam on multiple-layer dielectric structure [5], as shown in figure 1.8. The structure consisted of alternative dielectric layers with different permittivity.

Due to the excitation of a pair of TE and TM waves, the radiation beam width was significantly narrowed.

Figure 1.8 narrow-beam multiple layers structures (D.R. Jackson, A.A. Oliner, and Antonio, 1993).

Instead of dielectric material, Yevhen Yashchyshyn and Jozef Modelski used ferroelectric material [18]. They changed the dielectric constant from 85 to 115 by applying DC voltage into the ferroelectric material below figure 1.9. Different permittivity led to different main beam angles. Eventually, they can control the main beam angle by applying different DC voltage.

Figure 1.9 ferroelectric substrate leaky-wave antenna (Yevhen Yashchyn and Jozef Modelski, 2002)

Recently, meta-material was used to build leaky wave antenna [3][4][15]. In 2003, Christophe Caloz and Tatsuo Itoh reported that implementation of meta-material can increase the scanning beam angle [2]. Since the dielectric constant and permeability of the meta-material are both negative, resulting in a reverse direction of the propagation. Compared to traditional leaky wave antenna using positive material, which has limited beam scanning angle because the beam angle cannot reach to backward direction, experimentally, using meta-material significantly enlarged the beam scanning angle (Fig. 1.10).

Figure 1.10 Leaky wave antenna with meta-material (Christophe Caloz, and Tatsuo Itoh, 2003).

In summary, the leaky wave antenna provides an alternative method to tune the direction of the main radiation beam. Compared to phase array antenna, leaky wave antenna has compact volume and less complicate control circuit. The main concern is that significantly wide operating frequency bandwidth is required to reach large scanning angle for main radiation. For example, though leaky wave antenna is able to communicate with customer at different location, due to a large frequency change needed (~ a few GHz) to cover a wide angle, the receiver antenna must have ultra-wide bandwidth. This imposes a severe challenge for receivers. Using dielectric material, leaky wave antenna has been demonstrated with reduced volume, improved power gain, enlarged scanning angle, and narrowed beam width and operating frequency range.

In this thesis, I focused on applying ferroelectric material for its high dielectric constant to improve the performance of leaky wave antenna at radio frequencies. The thesis is organized as the followings: Chapter 1 gives the introduction of leaky wave antenna and a review of history. Chapter 2 presents the modeling and simulation results of leaky wave antenna based on high dielectric material. Chapter 3 discusses the fabrication, measurement, and characterization of leaky wave antenna. Chapter 4 outlines some of the future works.

Chapter 2- Design and Modeling

2.1 Basic structure of waveguide.

Radiation pattern of leaky wave antenna was simulated by using Agilent EMPro, which is 3-dimensional (3D) electromagnetic computation software. It includes Finite-Difference Time-Domain Method (FDTD) and Finite Element Method (FEM). FDTD requires the grid sufficient small to satisfy the smallest electromagnetic wavelength and the physical length, which in turn consume dramatic computing time. In EMPro, FDTD cannot provide continuous visualization of electromagnetic field inner waveguide. FEM can provide 3D radiation pattern with continuous visualization in both the near field and the far field. Therefore, I used FEM to compute near- and far- field antenna radiation pattern.

The basic structure of the leaky wave antenna is similar as a rectangular waveguide with a number of periodic holes cut into a side wall, shown in Figs 2.1.

Figure 2.1 The 3D modeling structure of leaky wave antenna (a) and detail parameters (b).

During wave propagation in the waveguide, radiation occurs through the holes. This requires wave travelling from one end to the other end of the waveguide. For travelling wave in the rectangular waveguide, the frequency should be higher than the so-called cutoff frequency, below which the wave cannot travel along the waveguide. The cut-off frequency depends on the width (a) and height (b) of the waveguide:

$$f_{cut-off} = \frac{1}{2\pi\sqrt{\mu\epsilon}} \cdot \sqrt{\left(\frac{m\pi}{WW}\right)^2 + \left(\frac{n\pi}{HW}\right)^2} \quad \text{Eq.(2-1)}$$

And the propagation wavelength is

$$\lambda = \frac{2\pi}{\sqrt{\omega^2\mu\epsilon - \left(\left(\frac{m\pi}{WW}\right)^2 + \left(\frac{n\pi}{HW}\right)^2\right)}} \quad \text{Eq.(2-2)}$$

Where m and n are integer number, ϵ is dielectric constant and μ is the magnetic permeability of the material filled in the waveguide.

2.2 Wave propagation in waveguide with high dielectric constant

According to Eq. (2-1), the cut-off frequency depends on the dielectric constant: larger the value of ϵ , smaller the cut-off frequency. For examples, Figs 2.2 and 2.3 depict the wave propagation in an air-core rectangular waveguide at 13.5 GHz and 25 GHz with WW=8mm, HW=4mm, and LW=100mm. The thickness of the copper layer is 0.1mm. The cutoff frequency is 18.75 GHz of the rectangular waveguide by Eq. (2-1).

Figure 2.2 Evanescent wave in empty (dielectric constant =1) waveguide at 13.5 GHz.

Figure 2.3 travelling wave in empty (dielectric constant =1) waveguide at 25 GHz.

From Eq. (2-2), insertion of dielectric material with larger value of ϵ , the propagation wavelength becomes shorter. This can be seen by comparing wave propagation in air-core waveguide (Fig. 2.5) and high k material filled waveguide with $\epsilon = 10$ (Fig. 2.2). The distance between two adjacent maxima reflects half wavelength. In Fig. 2.5, the

distance between adjacent maxima is much shorter than in Fig. 2.2, indicating a much larger wavelength in Fig. 2.2.

In addition, the larger dielectric constant can also reduce the cut-off frequency, allowing wave propagation in waveguide at lower frequency. Comparing Figs. 2.3 and 2.4, there is no traveling wave in the air-core waveguide (Fig. 2.3) at $f=13.5\text{GHz}$, while there is a clear wave propagation at 13.5 GHz as the waveguide filled with high k material $\epsilon = 10$.

The slightly increased dielectric constant which is 10 times is shown below Figs 2.5 and 2.2.

Figure 2.4 travelling wave in dielectric constant 10 material waveguide at 13.5 GHz

Figure 2.5 travelling wave in dielectric constant 10 material waveguide at 25 GHz

Nowadays, most antennas are operated at high frequency for the high gain, narrow beam width, and compact volume. However, limited by the facilities in our lab, both the signal generator and spectrum analyzer can only work below 3GHz. Therefore, in this work, I aimed for demonstrating the impacts on the performance of leaky wave antenna by using

high k materials. All leaky wave antennas were designed to work below 3 GHz.

However, the technology is readily to upgrade to high frequency.

To lower down the operating frequency below 3GHz, the dielectric constant k of the filled material in waveguide must be significantly increased. For example, as shown in Fig. 2.6, there is no traveling wave observed at 2GHz, while the same device conducts wave propagation at 13.5GHz (Fig. 2.4). It means 2 GHz is below the cutoff frequency. Figure 2.7 computed wave propagation in the same waveguide as in Fig. 2.6, but using a high k material with dielectric constant equal 722.

Figure 2.6 no travelling wave in dielectric constant 10 material waveguide at 2 GHz.

Figure 2.7 travelling wave in dielectric constant 722 material waveguide at 2 GHz

2.3 Design of leaky wave antenna

The travelling wave can radiate into space through holes array on the sidewall of the waveguide. The radiation pattern depends on many parameters including operating frequency, as well as configuration of the holes array like width, length, and spacing.

2.3.1 Impact on the slot distance (SD)

According to phase array antenna theory [19][20][21][23][26], to achieve vertical radiation (perpendicular to the waveguide), the spacing between adjacent antenna should equal to the propagation wavelength. Following this argument, the spacing between adjacent holes (SD in Fig. 2.1) of leaky wave antenna was designed equal to the propagation wavelength in the waveguide. From simulation at 2.475 GHz, the wavelength (λ) is about 4.8mm for a waveguide filled dielectric material of $\epsilon = 722$.

Figure 2.8 computed radiation pattern of a rectangular waveguide – based leaky wave antenna as a function of frequencies. By changing the frequency from 2.4GHz to 2.575GHz, the main beam of the radiation rotates from -38° to 35° , i.e. $0.417^\circ/\text{MHz}$. The radiation intensity along the main beam is about 3.25~5.48dB in table 1. In addition, the radiation power reaches the minimum as the main radiation beam approaching to the vertical direction. The explanation is depicted in Fig. 2.8 (d)-(f) by computing the electric field distribution along the top and bottom metallic wall of the waveguide.

At $f = 2.475\text{GHz}$, the excited electric wave on both the top and bottom metallic walls are stronger than those in Figs. 2.8 (d) and (f), where $f = 2.4\text{GHz}$ and $f = 2.575\text{GHz}$, respectively. This results in a uniform radiation pattern in the space, and consequently weak radiation in the main beam radiation.

Figure 2.8 Radiation pattern at 2.4GHz (a), 2.475GHz (b), and 2.575 GHz (c), and electric field distribution on the top and bottom metallic wall along the rectangular waveguide. The arrow reflects the main beam direction.

Table 1 Radiation direction and power gain in 2.4GHz, 2.475GHz, and 2.575GHz.

Frequency	2.4 GHz	2.475 GHz	2.575 GHz
Angle	-38 °	5 °	35°
Gain	5.1517 dB	3.258 dB	5.4843 dB

Figure 2.9 shows the radiation patterns simulated at 2.475 GHz with various slot distances. Radiation directivity and power gain are listed in table 2. Decrease the SD,

equivalent to reduce frequency, leads to the negative angle of the main radiation beam (Fig. 2.9 (a)). While increasing the SD, the main beam turns to the positive angle (Fig. 2.9 (c)). As the slot distance changes from 4.7mm to 4.9mm, the direction of the main radiation beam rotates from -20° to 23.8° in table 2.

Figure 2.9 Radiation pattern with various slot distance 4.7mm (a), 4.8mm (b), and 4.9mm (c), and electric field distribution on the top and bottom metallic wall along the rectangular waveguide.

Table 2 Radiation direction and power gain vs. slot distance

Slot distance	4.7mm	4.8mm	4.9mm
Angle	-20 °	5.0 °	23.8 °
Gain	2.5852 dB	3.258 dB	2.6738 dB
Frequency	2.475 GHz		

Further increasing the slot distance to twice of the original slot distance, i.e. $SD=2\lambda$, the radiation pattern is close to Fig. 2.8 where $SD=\lambda$, as shown in figure 2.10. According to the phase array antenna theory, the radiation patterns are essentially no difference as long as the distance between two adjacent antennas equal to $n\lambda$, where n is an integer, and the amplitude of electromagnetic wave from each antenna is the same. In our calculations, due to the attenuation from either the metal loss or dielectric loss, the amplitude of the electromagnetic waves arriving at each hole decreases. This results in the difference between Figs. 2.8 and 2.10.

Figure 2.10 Radiation patterns with slot distance 9.5238mm at 2.4 GHz (a), 2.475 GHz (b), and 2.575 GHz (c), and electric field distribution on the top and bottom metallic wall along the rectangular waveguide.

Table 3 Radiation direction and power gain for $SD=2\lambda$.

Frequency	2.4 GHz	2.475 GHz	2.575 GHz
Angle	-26.3 °	13 °	45°
Gain	4.287 dB	3.2182 dB	5.529 dB

2.3.2 Impact on the slot length (SL)

According to [24][25], narrower slot length leads to larger gain enhancement. Following this argument, the slot length was designed from 1 mm to 2 mm. From simulation at 2.575 GHz with $\varepsilon = 722$, narrower slot length leads to the narrower beam width.

Computation of beam width and radiation power gain with various slot widths is compared in table 4.

Table 4 Beam Width and Gain vs. Slot Length.

Slot Length	1 mm	1.5 mm	2 mm
Beam Width	46 °	46.2 °	46.419 °
Gain	5.4842 dB	5.1913 dB	4.7836 dB
Frequency	2.575 GHz		

2.3.3 Impact on radiation pattern by dielectric constant

Varying the dielectric constant of materials filled in the waveguide changes the radiation pattern, as shown in Figs. 2.11 and 2.12. In the simulations, the structural parameters of the waveguides are the same except the dielectric constant of the filled materials, $\varepsilon = 362$ in Fig. 2.11 and $\varepsilon = 813$ in Fig. 2.12, respectively. In Eq.2.2, larger value of dielectric constant leads to a shorter wavelength. However, as discussed in Sec. 2.3.1, the slot distance should match to the propagation wavelength for the vertical main radiation direction. As a result, waveguide filled with material having larger dielectric constant will have to be operated at lower frequencies, which is manifested by our simulations in Figs. 2.11 and 2.12. Compared Figs. 2.11, 2.8 and 2.12 in table 7, increased dielectric constant leads to lower operating frequency, and large ratio (η) of scanning angle respect to required frequency change. For example, the operating frequency range decreases from

300MHz to 160 MHz, and η increases from $0.25^\circ/\text{MHz}$ to $0.45^\circ/\text{MHz}$ as ϵ changes from 362 to 900. However, the power of using high k materials is the lower radiation power. Figs 2.13 shows simulation result that main radiation direction patterns with various dielectric constant $\epsilon = 137, 187, 287, 722$, and 813.

Figure 2.11 Scanning radiations with dielectric constant 362 at 3.34286 GHz (a), 3.51429 GHz (b), and 3.64286 GHz (c), and electric field distribution on the top and bottom metallic wall along the rectangular waveguide

Table 5 Radiation direction and power gain for dielectric constant 362.

Frequency	3.34286 GHz	3.51429 GHz	3.64286 GHz
Angle	-45.085 °	7.3157 °	31.282°
Gain	6.2189 dB	4.6166 dB	5.7944 dB

Figure 2.12 Scanning radiations with dielectric constant 813 at 2.225 GHz (a), 2.33125 GHz (b), and 2.4375 GHz (c), and electric field distribution on the top and bottom metallic wall along the rectangular waveguide

Figure 2.13 main radiation direction patterns with various dielectric constant materials

Table 6 Radiation direction and power gain for dielectric constant 813.

Frequency	2.225 GHz	2.33125 GHz	2.4375 GHz
Angle	-55 °	3 °	42°
Gain	3.7011 dB	3.1098 dB	5.7413 dB

Table 7 Change of angle per frequency with $\epsilon = 362$, $\epsilon = 722$, and $\epsilon = 813$.

Dielectric constant ϵ	362	722	813
Change of angle	76.28 °	72 °	97 °
Change of frequency	300 MHz	175 MHz	212 MHz
Change of angle per frequency (η)	0.25° /MHz	0.41° /MHz	0.457° /MHz

Chapter 3

- DEVICE FABRICATION AND CHARACTERIZATION

3.1 Device fabrication

The waveguide was fabricated on printed circuit board, as shown in Fig. 3.1 with $WW=8\text{mm}$, $HW=4\text{mm}$, and $LW=100\text{mm}$. SMA connector was soldered on the rectangular waveguide to feed the leaky wave antenna (Fig. 3.2). To construct material with $\varepsilon = 722$ and $\varepsilon = 813$, commercial available lead zirconate titanate (PZT) (#0027 in American Piezo Ceramics, Inc.) with $\varepsilon = 1900$ was used. The sketch of layers with $\varepsilon = 722$ and $\varepsilon = 813$ are shown in Fig. 3.3. According to effective medium theory[22], the effective dielectric constant of a multilayer equals to the average of the dielectric constant of the entire multilayer, as long as the thickness of each layer smaller than the propagation wavelength. In details, the width and thickness of PZT film are 8mm and 0.76mm, respectively. Material with $\varepsilon = 722$ was made of insertion of two PZT films in the waveguide, and $\varepsilon = 813$ is made of insertion of two PZT films together with a 2mm-width PZT strip.

As discussion 2.3.2, narrow slot length leads to higher radiation power gain. Limited by the facilities in our lab, the minimum slot length is 1mm. To coincide with the PZT films, the rectangular waveguide was designed with the following parameters: $WW=8\text{mm}$, $LW=100\text{mm}$, $WS = 8\text{mm}$, $SL=1\text{mm}$, $SD=4.8\text{mm}$, as shown in Figs 3.5.

Figure 3.1 Designed waveguide by printed circuit board.

Figure 3.2 Connected one port to waveguide.

Figure 3.3 Calculated dielectric constant $\epsilon = 722$ (a) and dielectric constant $\epsilon = 813$ (b).

Figure 3.4 PZT films to construct $\epsilon = 722$ (a) and $\epsilon = 813$ (b).

Figure 3.5 Fabricated top plate.

The antenna layout was designed using AUTOCAD and then converted to DXF format to input to the milling machinery. Board Master and Circuit CAM program was employed to transfer the layout structure into the printed board. After soldering all pieces and inserting PZT films, a proto-type leaky wave antenna was assembled as shown in Figs 3.6.

Figure 3.6 Assembled leaky wave antenna (a) and inserted the material the cross view (b)

3.2 Characterization of leaky wave antennas

3.2.1 Measurement set-up

The measurement set-up consists of a RF signal generator (Agilent N9310A 9 kHz – 3 GHz), a spectrum analyzer (Agilent N9320B), and a hand-made horn antenna. Figure 3.7 shows a photograph of the hand-made antenna, consisting of horn-shape cavity, and a dipole antenna. The spatial angular resolution is determined by $\vartheta = \frac{\phi}{l}$ (Fig. 3.7(b)). In our experiment, ϕ and l were set to 105mm and 300mm, respectively, achieving a spatial resolution ~ 20 degree. To magnify the signal, a RF power amplifier (Hittite 121701-1) was used. Agilent N9310A was connected with the leaky wave antenna to transmit the signal. Agilent N9320B was connected with horn antenna to receive the signal.

Figs 3.7 Hand-made horn antenna (a) and special angular resolution (b)

Figure 3.8 Setup diagram.

3.2.2 Leaky wave antenna with inserted dielectric material $\epsilon = 722$

The measurements were done by rotating the receiving antenna at a number of fixed operating frequencies. As shown in Fig. 3.8, this set-up allows locating the receiver antenna from -80 degree to $+80$ degree. During the measurements, the angular difference between each step of the receiving antenna is 10 degree. Measurement results obtained at $f=2.323, 2.352, 2.489, 2.54$ GHz are shown in figures 3.9 (a), (b), (c), and (d). From the measurements, the main radiation beam rotates from -5 to 50 degree, reaching $\eta=0.253^\circ/\text{MHz}$

Figure 3.9 Main radiation direction patterns with $\epsilon = 722$ by increasing frequencies 2.323 GHz (a), 2.353 GHz (b), 2.489 GHz (c), and 2.54 GHz (d)

3.2.3 Leaky wave antenna with inserted dielectric material $\epsilon = 813$

To understand the impact on radiation pattern by the dielectric constant, measurements were performed on the same rectangular waveguide but filled with material of having $\epsilon = 813$. As discussed in chapter 2, increase of dielectric constant will lead to reduction of propagation wavelength. Since the configuration of the holes array remains the same, i.e. slot width, length and distance, in order to equal the propagation wavelength to the slot distance to have vertical radiation, the operating frequency has to be reduced.

Experimentally, it was observed that the large angle tuning can be achieved from $f=2.1217$ GHz to 2.235GHz. The measurement results are shown in figures 3.10 (a), (b), (c), and (d) with $\eta=0.617^\circ/\text{MHz}$. Compared to $\epsilon = 722$, Higher dielectric constant results in larger value of η .

Figure 3.10 Main radiation direction patterns with $\epsilon = 813$ by increasing frequencies
2.1217 GHz (a), 2.1588 GHz (b), 2.2217 GHz (c), and 2.235 GHz (d)

3.3 Comparison between simulation and measurement

Comparison with theoretical modeling is shown in Table 8. In general, our measurement results coincide well with the modeling in view of scanning angle, angle range, operating frequencies, and variation of operating frequency to reach large angle scan. The difference might stem from fabrication uncertainties, like slot length, distance, and dielectric constant of the PZT materials. In comparison between simulation and measurement, the main radiation direction patterns of experiments showed the similar pattern as simulation in Figs 3.11.

Table 8 Compare between simulation and measurement with $\epsilon = 722$ and $\epsilon = 813$.

Dielectric constant ϵ	722 (simulation)	722	813 (simulation)	813
Angles	-38 ° to 35 °	-5 ° to 50 °	-55 ° to 42 °	-30 ° to 40 °
Change of angle	72 °	55 °	97 °	70 °
Operating frequencies range	2.4 to 2.575 GHz	2.323 to 2.54 GHz	2.225 to 2.4375 GHz	2.1217 to 2.235 GHz
Change of frequency	175 MHz	217 MHz	212 MHz	113 MHz
Change of angle per frequency (η)	0.41° /MHz	0.253° /MHz	0.456° /MHz	0.617° /MHz

Figure 3.11 Comparison between simulation and measurement for $\varepsilon = 722$ (a) and $\varepsilon = 813$ (b).

Chapter 4 - Conclusion

4.1 Summary

In this thesis, the leaky wave antennas with inserted high k material was demonstrated for working at low frequencies, narrowing operating frequency bandwidth, and reducing the volume. To reach the 55 degree scanning angles with $\varepsilon = 722$, the operating frequency bandwidth was 217 MHz from 2.323 to 2.54 GHz. In the leaky wave antenna with $\varepsilon = 813$, the operating frequency bandwidth was tuned 113 MHz from 2.1217 to 2.235 GHz to achieved 70 degree scanning angle. In the other word, the scanning angle per frequency with higher dielectric material was achieved to larger value.

4.2 Future works

Even if the leaky wave antennas with high k material was working successfully, the gain, systematic scanning angle, and beam width need an optimized design for better performance. Increasing the gain, expanding the scanning angles, and narrowing main beam width were remained to be solved in future. In addition, the technic about changing slot distance at fixed frequency leads to the tunable direction of radiation. Optimizing the challenges at low frequency leads to connect to a lot of successful wireless applications such as high efficiency reverse parking sensor, coaxial cable type for leak wave antenna, and wireless device with WiFi and Bluetooth.

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