DESIGN AND PACKAGING OF ULTRA BROADBAND LITHIUM NIOBATE MODULATOR FOR MILLIMETER-WAVE APPLICATIONS

by

Julien Macario

A dissertation submitted to the Faculty of the University of Delaware in partial fulfillment of the requirements for the degree of Doctor of Philosophy in Major

Fall 2014

© 2014 Julien Macario
All Rights Reserved
DESIGN AND PACKAGING OF ULTRA BROADBAND LITHIUM NIOBATE MODULATOR FOR MILLIMETER-WAVE APPLICATIONS

by

Julien Macario

Approved:

_________________________________________________________

Kenneth E. Barner, Ph.D.
Chair of the Department of Electrical and Computer Engineering

Approved:

_________________________________________________________

Babatunde A. Ogunnaike, Ph.D.
Dean of the College of Engineering

Approved:

_________________________________________________________

James G. Richards, Ph.D.
Vice Provost for Graduate and Professional Education
I certify that I have read this dissertation and that in my opinion it meets the academic and professional standard required by the University as a dissertation for the degree of Doctor of Philosophy.

Signed:

Dennis W. Prather, Ph.D.
Professor in charge of dissertation

I certify that I have read this dissertation and that in my opinion it meets the academic and professional standard required by the University as a dissertation for the degree of Doctor of Philosophy.

Signed:

Keith W. Goossen, Ph.D.
Member of dissertation committee

I certify that I have read this dissertation and that in my opinion it meets the academic and professional standard required by the University as a dissertation for the degree of Doctor of Philosophy.

Signed:

Mark S. Mirotznik, Ph.D.
Member of dissertation committee

I certify that I have read this dissertation and that in my opinion it meets the academic and professional standard required by the University as a dissertation for the degree of Doctor of Philosophy.

Signed:

Ahmed Sharkawy, Ph.D.
Member of dissertation committee
ACKNOWLEDGMENTS

I would like to take this opportunity to express my gratitude to those who have made the work presented in this dissertation possible. First, I would like to thank my advisor Dr. Dennis W. Prather for his guidance and his tremendous support throughout this journey. I also would like to thank him for taking a chance on me when I was just a French student in Europe whom he had no connection with, and for welcoming me to join his group with so much enthusiasm.

I would also like to thank my other committee members Dr. Keith Goossen, Dr. Mark Mirotznik and Dr. Ahmed Sharkawy for taking the time to review my dissertation, attending my defense and their valuable feedback. My special thanks go to Dr. Peng Yao for teaching me so much about fabrication over the years, for listening to my ideas and guiding me on my research. I am grateful to Dr. Shouyuan Shi for helping me with designs, data processing and for his overall support. I would like to thank Dr. Christopher Schuetz for his intellectual advices and mentorship.

I would also like to thank all the colleagues and all the group members that I had the chance to collaborate with throughout my years in Dr. Prather’s group. I would like to thank Andrew Mercante, Alicia Zablocki, Charles Harrity, Dr. Richard Martin and Dr. Rownak Shireen for their precious help in the labs, Dr. Elton Marchena, Dr. Timothy Creazzo and Dr. Mathew Zablocki for their assistance with all the fabrication equipment, Jian Bai, Yifei Zhang, Dr. James Mutitu and Dr. Benjamin Olbricht for being such great office mates. I would like to acknowledge Defense
Advanced Research Projects Agency (DARPA) and Office of Naval Research (ONR) for generously supporting various aspects of this work.

I want to say a big thank you to my longtime partner in life, Dr. Muhua Yang, for her love and support over all those years. I would also like to thank my dear friend Jean-Remy Bonnefoy for always being there for me when needed and for making life outside of work so much more enjoyable.

Finally, I would to thank my late mother, who I know would be proud, for teaching me the right values in life and for being such a great example. I would like to thank my sister Delphine for her encouragement and for helping me become who I am. I would like to thank my step-father Jean-Jacques Pouètre for helping raising me and for his mentorship.
# TABLE OF CONTENTS

LIST OF FIGURES .................................................................................................................. ix  
ABSTRACT .................................................................................................................................. xiii

Chapter

1 INTRODUCTION .................................................................................................................... 1  

1.1 Millimeter-Wave Imaging ................................................................................................. 2  

1.1.1 Millimeter-Waves Characteristics ................................................................................. 2  
1.1.2 Passive vs. Active Imaging ............................................................................................ 3  
1.1.3 Potential Millimeter-Wave Imaging Applications ......................................................... 4  
1.1.4 Blackbody Radiation ...................................................................................................... 5  
1.1.5 Atmospheric Windows for Passive Imaging ................................................................. 6  

1.2 Passive Millimeter-Wave Imaging System Architecture .................................................. 9  

1.2.1 Distributed Aperture Technology ................................................................................ 9  
1.2.2 Optical Upconversion Detection Technique ............................................................... 11  

1.3 Motivation and Dissertation Outline ................................................................................. 12  
1.4 List of Original Contributions .......................................................................................... 15

2 MODULATOR DESIGN FOR MILLIMETER-WAVE APPLICATIONS .... 18  

2.1 Pockels Electro-Optic Effect ............................................................................................ 20  
2.2 Phase Modulation ............................................................................................................. 24  

2.2.1 DC Operation ............................................................................................................... 24  
2.2.2 Low Frequency Operation ........................................................................................... 26  
2.2.3 RF Operation ............................................................................................................... 28  

2.3 Broadband Operation Challenges .................................................................................... 32  

2.3.1 Index Matching ............................................................................................................ 33  
2.3.2 RF Propagation Losses ............................................................................................... 35  

2.4 Modulator Design .............................................................................................................. 37
LIST OF FIGURES

Figure 1.1: Naturally emitted blackbody radiation and radiation through 1 km of fog for an emissive object at temperatures of 77 K, 300 K and 6000 K [2]........................................................................................................................................6

Figure 1.2: Atmospheric attenuation curves from 10 GHz to 1 THz under various levels of relative humidity (RH) and fog [2]............................................................................7

Figure 1.3: Architecture of the distributed aperture imaging system based on optical upconversion technique [8].................................................................10

Figure 1.4: Optical upconversion technique .........................................................................11

Figure 2.1: Electro-optic modulation concept. .................................................................20

Figure 2.2: LiNbO₃ ellipsoid with principal axes aligned with coordinate axes. ....21

Figure 2.3: Refractive index ellipsoid configuration of LiNbO₃ when an electric field is applied along its z-axis................................................................................24

Figure 2.4: Phase modulation of an optical mode traveling through LiNbO₃. ........25

Figure 2.5: Intensity of the modulated optical field at the LN optical waveguide output. A lower sideband and an upper sideband are created at an offset frequency of ωₘ and ωₘ, respectively, from the optical carrier frequency ω₀. .................................................................................................................................28

Figure 2.6: Traveling-wave modulator structure using microstrip. .........................29

Figure 2.7: Cross-section design of the traveling-wave LiNbO₃ phase modulator. ..38

Figure 2.8: Design parameters impact on the RF behavior of the modulator and on the overall modulation efficiency .................................................................39

Figure 2.9: Modulator’s cross-section numerical analysis for (a) optical field and (b) electrical and magnetic field..............................................................................40

Figure 2.10: Simulated half-wave voltage \( V_π(f_m) \) .................................................................42
Figure 3.1: Fabrication process flow chart. ................................................................. 45

Figure 3.2: 3D representation of a processed LiNbO$_3$ wafer containing sixty modulators. ................................................................. 52

Figure 3.3: SEM pictures of fabricated LiNbO$_3$ phase modulators at different scales with views of (a) wafer with modulator array, (b) CPW in the launch region and (c) ridged CPW central electrode on top of diffused Ti optical waveguide. ................................................................. 53

Figure 3.4: SEM picture of the modulator’s cross-section. ........................................ 54

Figure 3.5: 3D representation of LiNbO$_3$ modulators after (a) wafer dicing and polishing and (b) individual chip dicing. ................................................................. 55

Figure 3.6: 3D representation of modulator’s cross-section after SiO$_2$ etching for millimeter-wave index tuning. ................................................................. 57

Figure 3.7: 3D representation of an optically integrated LiNbO$_3$ phase modulator. 58

Figure 4.1: Transmission parameter S$_{21}$ of a LiNbO$_3$ modulator suffering from strong substrate mode coupling in millimeter-wave region. .................. 62

Figure 4.2: Dicing process to eliminate substrate modes in LiNbO$_3$ substrate. ....... 65

Figure 4.3: SEM pictures (a) RF launch region and interaction region of modulator thinned down to 100 µm and (b) close-up look at the RF launch region of a modulator thinned down to 20 µm for full millimeter-wave spectrum operation. ................................................................. 66

Figure 4.4: Effect of substrate thickness on substrate mode coupling cutoff frequency. ................................................................................................................. 67

Figure 4.5: Impact of launch region on substrate mode coupling for straight CPW structure. ................................................................................................................. 69

Figure 4.6: Impact of launch region on substrate mode coupling with 90-degree bend in CPW structure at launch. ................................................................. 71

Figure 4.7: SEM pictures of surface roughness of the CPW structure of (a) modulator with 90-degree bend in launch region and (b) modulator with a straight CPW structure. ................................................................. 72
Figure 5.1: Measured transmission parameter $S_{21}$ over the 280 GHz bandwidth. The transmission parameter $S_{21}$ confirms that the substrate modes have been suppressed for a substrate thickness of 30 µm. .................. 77

Figure 5.2: Sidebands normalization process. The optical power at the modulator output is measured (a), then normalized to the optical carrier power (b), and finally normalized to the RF power (c) where the modulation efficiency can be finally read in the logarithm domain. ...................... 82

Figure 5.3: Optical modulation sidebands test setup. ..................................................... 83

Figure 5.4: 300 GHz optical modulation spectrum. Each pair of sidebands centered around the optical carrier represents the mmW energy for a given mmW frequency upconverted to optical energy using the EO effect of the LiNbO$_3$. ................................................................. 84

Figure 5.5: RF power delivered by the PNA over the entire mmw spectrum. ....... 85

Figure 5.6: Measured and calculated modulator half-wave voltage $V_\pi$. The measured $V_\pi$ extracted from the sidebands measurements is in agreement with the $V_\pi$ calculated using the S-parameters. ...................... 86

Figure 5.7: Half-wave voltage DC-$V_\pi$ measurement on oscilloscope. ................. 87

Figure 6.1: Module parts. ........................................................................................... 90

Figure 6.2: Broadband module's integration design. ................................................. 91

Figure 6.3: Ribbon bonding transition design............................................................ 93

Figure 6.4: Packaged module.................................................................................... 95

Figure 6.5: Packaged module connected to the RF source through a 1.0 mm cable. 97

Figure 6.6: S-parameter characteristics of the modulator. ...................................... 99

Figure 6.7: Setup for module's RF insertion loss measurement. ......................... 100

Figure 6.8: Simulated and measured S-parameter characteristics of the 1.0 mm connector-Al$_2$O$_3$ chip transition. ................................................................. 101

Figure 6.9: S-parameter characteristics of modulator and module. ....................... 102

Figure 6.10: Integrated 1.0 mm module connected to the RF source (PNA) through a 1.0 mm RF connector. ................................................................. 104
Figure 6.11: Optical modulation spectrum. ................................................................. 105
Figure 6.12: Half-wave voltage $V_{\pi}$. ........................................................................ 106
Figure 7.1: Sample imagery from passive millimeter-wave perception study of small watercraft. ........................................................................................................ 110
Figure 7.2: LiNbO$_3$ modulator integrated into Optical Upconversion Module. ...... 111
ABSTRACT

Millimeter-waves have grown in popularity for imaging applications in recent years due to their unique properties. In the electromagnetic spectrum, millimeter-waves can be seen as the frontier between radio waves and optical waves and thus benefit to some extent from both worlds. On one hand, they are long enough to penetrate through obscurants and thin dielectric and experience low atmospheric attenuation. On the other hand, they are small enough to be relatively convenient for imaging by providing good resolution for a manageable aperture size. In addition, millimeter-wave imaging has also benefited directly from the recent emergence of new components in the millimeter-wave region, an effort mainly driven by the booming of the telecommunication market and the quest for bigger and faster networks.

At the University of Delaware, we have spent the last decade developing passive millimeter-wave imaging systems at 35 GHz and in W band at 77 GHz and 94 GHz. The architecture of these systems is based on the optical upconversion of the native blackbody radiations emitted by the scene and detected by an antenna or a distributed aperture antenna array. The modulator converting the detected millimeter-wave radiations into the optical domain on a carrier is considered the heart of the detection technique. However, there are no commercially available modulators capable of operating in W band, which led us to develop our own.

In this regard, I present an ultra broadband LiNbO₃ electro-optic phase modulator capable of operating in W band and well beyond. The main challenges in
designing broadband LiNbO$_3$ are high RF propagation losses and the intrinsic mismatch in LiNbO$_3$ between the millimeter-wave index and the optical index. I present a design based on ridge coplanar waveguide that allows precise index matching and minimizes propagation losses. In addition to the design, I also introduce the main processing techniques developed to fabricate the LiNbO$_3$ modulator designed.

To enhance the efficiency of the LiNbO$_3$ modulator, I present a novel LiNbO$_3$ micromachining process that eliminates substrate mode coupling, which is a strong source of RF attenuation in the millimeter-wave region. As a result, optical sidebands generation were observed over the full millimeter-wave spectrum.

Modulation in the millimeter-wave region is interesting not only for imaging applications but also for the telecommunication industry. In that regard, I developed a fully packaged, low RF insertion loss, module integrating a LiNbO$_3$ modulator with a 1 mm coaxial connector for modulation operation over the 0-110 GHz band.
When incident on a material, electromagnetic radiations can be either transmitted, reflected, absorbed or experience a combination of these three phenomena. The nature of this interaction is determined by the physical properties of the material and by the radiating frequency (wavelength) of the incident wave. A classic example illustrating the wavelength sensitivity of materials is the human eye, which can only detect electromagnetic radiations whose wavelength is comprised between 390 nm and 750 nm. Those wavelengths define the visible spectrum, which corresponds to the 400-770 THz frequency bandwidth. With the eyes acting as filter, humans are limited in their ability to see in the electromagnetic spectrum. A clear benefit of detecting only a narrow range of wavelength is the ease for our brain to process the information captured. The downside is the blindness in interesting bandwidths. In an effort to capture information invisible to the human eye, advanced imaging technologies have been developed in most of the regions of interest. For instance, infra-red (0.75-1000 µm) detection is commonly used to image the difference of temperature in a scene. X-rays (0.01-10 nm) imaging allows, among many applications, medical radiography and the study of celestial objects. Radar systems in the 3-12,000 MHz range have been developed for long distance transmission and detection due to the low atmospheric attenuation in that bandwidth. One interesting band of the electromagnetic spectrum for imaging that remains
relatively unexplored is the millimeter-wave region, where the wavelength in free
space is in the 1-10 mm range (30-300 GHz). Millimeter-wave radiations are very
attractive for imaging because of their ability to penetrate through fog, cloud, smoke
and thin dielectric while allowing high-resolution imaging. The lack of good
millimeter-wave sources and detectors has been for a long time the main limitations
for imaging in this region. The University of Delaware, in collaboration with Phase
Sensitive Innovations Inc., is developing passive millimeter-wave imaging systems
using distributed aperture approach combined with millimeter-wave optical
upconversion technique.

The first section of this chapter introduces the millimeter-wave imaging
technology and presents its advantages over other imaging systems, its potential
applications and the principle governing passive millimeter-wave detection. Section
1.2 reveals the millimeter-wave detector architecture developed at the University of
Delaware to overcome the technological barriers faced by current imaging systems.
The third section discusses the motivation for my work and presents the outline of this
dissertation. Finally, my original contributions to the field and my publications are
listed in section 1.4.

1.1 Millimeter-Wave Imaging

1.1.1 Millimeter-Waves Characteristics

Millimeter-waves are electromagnetic radiations whose wavelengths are
comprised in the 1-10 mm (30-300 GHz) region. They are located between the
microwave band (300 MHz -30 GHz) and IR domain (300 GHz – 430 THz). In a
sense, millimeter-waves can be considered as the outer limit of the radio frequency
spectrum. The millimeter-wave spectrum draws the border between radio waves and light.

One attractive advantage that radio waves (long wavelengths) possess over shorter wavelengths (IR, visible…) is their relatively low atmospheric attenuation. Therefore, they have been used for years and proven convenient for long distance aerial communication and remote sensing. For instance, the long-distance radio broadcasting techniques based on amplitude modulation (AM) and frequency modulation (FM) typically emit in the 535-1605 KHz and 87.5-108 MHz bands, respectively. Modern telecommunication systems such as cell phone networks or Bluetooth operate in the 1.6-2.4 GHz band. In addition to experiencing low atmospheric attenuation, radio waves have the special ability to penetrate thin dielectrics and many atmospheric obscurants such as fog, sand brownout, smoke or clouds. However, with the size of the antennas and detectors increasing with wavelength and resolution, imaging systems in the microwave region would prove to be too large and impractical for most of the potential applications. Millimeter-waves offer a good compromise with a still relatively low atmospheric attenuation, smaller aperture size, higher resolution, and the ability to propagate through obscurants and thin dielectric materials.

### 1.1.2 Passive vs. Active Imaging

Two methods are generally considered for millimeter-wave imaging, active and passive. In active imaging a source radiates millimeter-waves at the scene, and either the transmitted or the reflected waves are detected. In this case, the millimeter-waves illuminating the scene are usually of known wavelength with a significant radiant power, making the detection relatively easy. In contrast, passive imaging relies
on the detection of the native blackbody radiations of the objects of the scene and of the reflected blackbody radiations emitted by external sources and reflected from those objects [1]. No illuminating is needed in this case, but only a very low power is available for detection. As a result, severe constraints are imposed on the detectors.

In addition to the nature of the millimeter-wave source, more features differentiate the two imaging methods. Since passive systems rely solely on native radiation, they are non-invasive and completely covert. Those traits are highly desirable for military applications. Passive imaging also prevents from any health risk or material damage. Moreover, passive systems often offer higher image contrast for outdoor imaging by benefitting from the highly reflective nature of certain object. For example, reflective materials such as metal will appear to have an effective temperature equivalent to that of the cold sky, which is lower than 100K in the millimeter-wave region. Therefore, metallic objects are easily discernable from dielectrics for instance. Finally, passive systems are usually less cumbersome than active system due to the absence of illuminating source. This is a key element as some potential applications are very stringent on volume and weight. For all the relative advantages of passive imaging over active imaging, we have decided at the University of Delaware to focus our effort on developing passive millimeter-wave imaging systems.

### 1.1.3 Potential Millimeter-Wave Imaging Applications

Passive millimeter-wave imaging has potentials in both commercial and military applications provided that it is built in a cost-effective manner, light and compact. The imagers can assist aircrafts in taking off, landing and taxing in low-visibility conditions. They can also enhance obstacle avoidance capabilities for low
flying aircrafts or be deployed for covert aerial surveillance and reconnaissance under any atmospheric conditions, under daylight or at night. If compact and light enough, passive millimeter-wave imagers could even be deployed on drones, who have the added advantage of being hard to detect by conventional radars. Because millimeter-wave radiations have the ability to penetrate through thin dielectrics like clothing, those imagers can detect concealed weapons and explosives in airports and other security checkpoints. They also can have applications in all-weather search and rescue missions, like in foggy or smoky conditions.

1.1.4 Blackbody Radiation

The main benefit of millimeter-wave radiation for imaging is their ability to penetrate fog, cloud, smoke cover and thin dielectric with very low attenuation. The naturally emitted blackbody radiation of the sun (6000 K), of a ground object (300 K) and of the cold sky (77 K) through clear atmospheric conditions and through 1 km of fog is shown in Fig. 1.1. As expected, the figure reveals that the blackbody radiations emitted at frequencies beyond the millimeter-wave region are quickly absorbed under foggy conditions. The figure also shows the amplitude of the blackbody radiations increases as a function of frequency. Therefore, it would be highly beneficial, from a sensitivity and resolution standpoint, to be able to detect blackbody radiations at the highest frequency possible. According to the figure, the frequency that provides the highest blackbody radiation energy for imaging through obscurants is located around 300 GHz, the border of the millimeter-wave region.
Figure 1.1: Naturally emitted blackbody radiation and radiation through 1 km of fog for an emissive object at temperatures of 77 K, 300 K and 6000 K [2].

1.1.5 Atmospheric Windows for Passive Imaging

The level of atmospheric attenuation plays a major role in determining the system performance for outdoor applications. Although atmospheric attenuation is almost negligible for long wavelengths such as radio waves or microwaves, it becomes an issue in the millimeter-wave spectrum due to absorption from water vapor and oxygen. Curves of atmospheric attenuation for various levels of relative humidity condition in the 10 GHz–1 THz range are presented in Fig. 1.2.
Figure 1.2: Atmospheric attenuation curves from 10 GHz to 1 THz under various levels of relative humidity (RH) and fog [2].

Millimeter-wave imaging systems need to operate in low atmospheric attenuation frequency windows in order to collect a maximum amount of energy and optimize the image contrast. This is especially true for passive imaging because the level of native blackbody radiation is already rather low. The benefits of operating at high frequency are the increased image resolution for a given aperture size and the larger amount of blackbody radiation available for detection, as seen in Fig. 1.1. However, this comes at the price of lower image contrast, for two reasons. The first is the increased atmospheric absorption. Since the radiation process is reciprocal, the absorbing atmosphere also emits radiation according to its physical temperature,
therefore obscuring the scene. The atmosphere also reflects the cold sky radiation and increases its effective temperature, which as a result lowers the contrast of reflective objects in the scene. The second reason is the lower performance metrics of receivers at higher frequencies, which reduces the resolvable temperature.

The relatively high transmission windows in the millimeter-wave region are centered around 35 GHz, 94 GHz, 140 GHz and 220 GHz, as seen in Fig. 1.2. Ultimately, the application and the technology available in a particular frequency band play a major role in defining the frequency of operation for which to design the imaging system. So far, most millimeter-wave imaging systems that have been built are operating at 35 GHz because of the wide array of components commercially available at this frequency at a reasonable cost. At the University of Delaware, we have been developing for several years passive millimeter-wave imaging systems using a novel detection technique based on optical upconversion that eases the system requirements for high frequency of operation. A first proof of concept was successfully realized at 35 GHz using commercially available [3]. A second system based on single-pixel scanning was built at 94 GHz [4]. Recently, a third passive millimeter-wave system that provides real-time, video rate imagery has been designed and built for 77 GHz operation by partially using components developed in-house for this specific application [5]. A fourth real time, video rate imaging system at 94 GHz is at the initial phase of development. In the future, more systems should be designed for operation in the 140 GHz and 220 GHz windows. The next section discusses the novel architecture developed at the University of Delaware to passively detect millimeter-waves for imaging.
1.2 Passive Millimeter-Wave Imaging System Architecture

1.2.1 Distributed Aperture Technology

The imaging systems developed at the University of Delaware rely on the distributed aperture technique to passively capture the blackbody radiation emitted in the millimeter-wave region and reconstruct the image of the scene [6]. In the past, the usefulness of millimeter-wave imagers has often been limited by the large aperture sizes required to obtain images of sufficient resolution due to the diffraction limit. Traditional millimeter-wave imaging techniques, such as focal plane arrays or scanned systems, require a volumetric increase in imager size and, consequently, weight to improve imager resolution. Thus, such imaging systems are largely impractical for many applications like aerial surveillance and all-weather navigation, when they have to be placed on aircrafts where space is limited and added weight prohibitively costly. Distributed aperture techniques could help extend the range of applications by significantly scale down the volume and weight requirements of high-resolution imagers. Unfortunately, traditional distributed aperture techniques for millimeter-wave imaging, such as RF detection and calorimetric detection, encounter serious complications due to significant detection, routing, and processing requirements of such distributed arrays. In RF style detection systems, the detected millimeter-wave signal is amplified then downconverted, or the millimeter-wave signal is directly detected via non-linear effects provided by semiconductor technology. For each antenna, a system often needs local oscillator, mixers and coaxial cables or metal waveguides, which considerably drives up the cost, the weight and the spatial requirements. Calorimetric detectors use thermally dependent material properties to detect the radiated energy, but the sensitivity of that technique is insufficient for
passive millimeter-wave imaging. A new approach based on optical upconversion technique has been developed at the University of Delaware that solves the space, weight and sensitivity issues encountered when using conventional techniques [7-9].

The architecture of the passive millimeter-wave imaging system based on distributed aperture detection is presented in Fig. 1.3.

![Architecture of the distributed aperture imaging system based on optical upconversion technique](image)

Figure 1.3: Architecture of the distributed aperture imaging system based on optical upconversion technique [8].

In this architecture, the millimeter-waves are detected by the distributed antenna array and guided to a modulator where they are upconverted into the optical
domain. The modulated optical signal is then routed through flexible and lightweight fiber optics to an optical filter to conserve only the millimeter-wave energy. Next, the fibers containing the filtered optical signals are directed toward a lens that focuses the light on a low frequency optical detector array.

1.2.2 Optical Upconversion Detection Technique

The optical upconversion technique employed to detect and process the millimeter-wave radiations emitted by a scene is presented in Fig. 1.4 [10,11].

![Optical upconversion technique](image)

**Figure 1.4:** Optical upconversion technique.

In this architecture, a single monochromatic optical source is split and fed to each element of the array. The incoming millimeter-wave radiation of frequency \( \omega_m \) is detected by an antenna and guided to an electro-optic or electro-absorption modulator.
The millimeter–wave signal induces a change in refractive index in the modulator’s substrate, which creates sidebands on each side of the optical carrier of frequency $\omega_m$ traveling through the modulator. The sidebands are located at a frequency offset from the carrier equal to the frequency of the millimeter-wave detected, corresponding to $\omega_m \pm \omega_o$, with an intensity that is linearly dependent on the power of the millimeter-wave coupled to the modulator. An optical filter then suppresses the optical carrier and eventually one of the sidebands. Subsequently, the optical sideband is focused through a lens on a low frequency photodetector that performs square law detection in the optical domain to convert the sideband to a near DC electrical current, or photocurrent. A high-gain transimpedance amplifier may then be used to provide low noise conversion of photocurrent to a voltage signal.

1.3 Motivation and Dissertation Outline

The primary factor limiting the development of high resolution passive millimeter-wave imaging systems is the lack of suitable technology for fabricating large array of millimeter-wave detectors with sufficient sensitivity. The optical upconversion technique developed at the University of Delaware circumvents the stringent sensitivity requirements placed on conventional millimeter-wave detectors by shifting the detection technique from the millimeter-wave region to the optical domain. In this architecture, the sensitivity of the imaging system is highly dependent on the modulator’s ability to convert the millimeter-wave energy into an optical signal [11]. The lack of commercially available devices capable of efficient modulation deep in the millimeter-wave region at the time the system was designed motivated us to design and fabricate our own modulator. Moreover, the modulator needs to be deeply integrated into the imaging system to convert the millimeter-wave radiation into the
optical domain as early as possible after the detection by the antenna. Vendors typically only provide fully integrated devices. By minimizing the length of the RF transition between the antenna and the modulator, the RF losses are limited and more millimeter-wave energy is coupled into the modulator. Optimizing the amount of millimeter-wave radiation available for modulation is of critical given the small amount that is passively emitted by the scene and then detected by the antenna. This dissertation reports the design, fabrication, characterization and packaging of an ultra broadband external phase modulator developed for passive millimeter-wave imaging.

In Chapter 2, I present the design of LiNbO$_3$ electro-optic phase modulator designed to operate over the full millimeter-wave spectrum. The modulation concept and the advantages of LiNbO$_3$ over other technologies for millimeter-wave modulation are first presented, followed by a theoretical description of phase modulation in LiNbO$_3$. The modulator design challenges are then discussed before finally revealing our design, accompanied with simulation results.

Chapter 3 focuses on the modulator’s fabrication process. The different steps of the fabrication are discussed, from the optical waveguide fabrication all the way to the optical packaging with optical fibers.

In Chapter 4, I present a novel LiNbO$_3$ micromachining technique that prevents the coupling of the millimeter-wave signal into substrate modes and extend considerably the modulation bandwidth of LiNbO$_3$ modulators. As the frequency of operation increases, the millimeter-wave wavelength becomes small enough to couple into substrate modes supported by the modulator. By reducing the LiNbO$_3$ substrate’s thickness, it is possible to eliminate substrate modes coupling in the entire millimeter-wave region. Experimental measurements of substrate mode coupling for different
LiNbO$_3$ substrate thicknesses and different machining configurations are presented in the 0-220 GHz band.

The performance of the LiNbO$_3$ modulator fabricated in our facilities over the full millimeter-wave spectrum is presented in Chapter 5. Excellent modulator transmission properties have been measured over the 0-280 GHz including perfect matching between RF effective index and optical effective index, a critical condition for broadband operation. Optical modulation sidebands have been observed for the first time in the full 0-300 GHz band, demonstrating that LiNbO$_3$ remains responsive to the electro-optic effect over the entire millimeter-wave region. The conversion efficiency of the modulator in terms of half-wave voltage is reported.

In chapter 6, the LiNbO$_3$ phase modulator packaged with a 1 mm coaxial connector in a module for operation over the 0-110 GHz band is reported. The design of the low-loss RF transition between the 1 mm connector and the modulator and the integration process are discussed in details. Simulation results showing low insertion loss in the 1-2.5 dB range have been experimentally verified in the 70-110 GHz band through S-parameters characterization. The module’s low insertion loss was later confirmed by measuring the optical modulation spectrum over the full 0-110 GHz band.

In chapter 7, I conclude with a summary of the work accomplished in this dissertation and comments on its integration in the passive millimeter-wave imaging systems developed at the University of Delaware. I also discuss new designs on LiNbO$_3$ modulators that is being conducted that could improve bandwidth and modulation efficiency.
1.4 List of Original Contributions

The most significant aspects of my work as a Ph.D. student are presented in this dissertation. The research on LiNbO$_3$ modulator technology and packaging has led to the following contributions:

1. Design, fabrication and characterization of LiNbO$_3$ electro-optic modulator over full millimeter-wave spectrum
2. Implementation of millimeter-wave effective index tuning technique using silicon dioxide buffer layer wet etching
3. Design and implementation of LiNbO$_3$ modulator micromachining technique for substrate mode suppression in millimeter-wave region
4. Experimental demonstration of substrate mode coupling resonance and substrate mode suppression over full millimeter-wave region in LiNbO$_3$ modulators region
5. Design, fabrication and characterization of fully packaged LiNbO$_3$ modulator with 1 mm coaxial connector for 110 GHz bandwidth

Most of the contributions were published in peer-reviewed journals (Appendix A) and/or presented at the conferences. The following is a list of my publications:

Peer Reviewed Journal Publications:


**Conference Publications and Presentations:**


“Development of a high-speed modulator for a W-band millimetre-wave
imaging system” Proceedings of the SPIE - The International Society for

5. P. Yao, C.A. Schuetz, S. Shi, J. Macario, R. Shireen and D.W. Prather,
"Development of high speed modulator for W-band," in Passive Millimeter-
Wave Imaging Technology XII, April 16, 2009 - April 16, pp. The
International Society for Optical Engineering (SPIE), 2009.

W-band transition from coplanar waveguide to rectangular waveguide,” IEEE
International Symposium on Antennas and Propagation and USNC/URSI

GHz millimeter wave imaging system implementing optical upconversion,”
Proceedings of SPIE - The International Society for Optical Engineering, v
7117, Millimetre Wave and Terahertz Sensors and Technology, p 71170T
(2008).

8. P. Yao, R. Shireen, J. Macario, C. A. Schuetz, S. Shi, and D. W. Prather
“Design, fabrication and characterization of LiNbO3 optical modulator for
high-sensitivity mmW imaging system,” Proceedings of SPIE - The
International Society for Optical Engineering, Passive Millimeter-Wave
Chapter 2

MODULATOR DESIGN FOR MILLIMETER-WAVE APPLICATIONS

The millimeter-wave optical upconversion technique combined with the distributed aperture approach developed at the University of Delaware for passive imaging impose several constraints on the type of optical modulator that can be integrated into the imaging system. In this architecture, the phase of the incoming millimeter-wave radiations must be transferred into the optical domain to reconstruct the image on the back end of the system. Moreover, each optical modulator must be fed from a common optical source to maintain the phase coherence of the optical fields at the detector. Therefore, external modulation is required. The amplitude of the millimeter-wave must also be preserved during the conversion into the optical domain to reproduce the contrast of the scene. Optical phase modulation technique accomplishes these requirements by imprinting on each side of the carrier optical sidebands that conserve the phase of the modulating millimeter-wave signal and whose power is linearly related to the power of the millimeter-wave coupled into the device.

To perform an efficient phase modulation, a substrate material must possess several key characteristics. The material must display low optical loss at the wavelength emitted by the laser source, preferably at 1550 nm to benefit from the wide array of cheap and efficient photodetectors developed at this wavelength for the telecommunication market. The substrate material must also be able to handle the high optical power necessary for modulation sidebands detection and have low optical
coupling loss. Moreover, the substrate material must provide strong modulation response at 77 GHz and 94 GHz for the systems currently being developed, and at 140 GHz and 220 GHz for the next generations. The material should also be available in high quality at an adequate size and be stable over the operating temperature range. Finally, the substrate material must remain highly reliable over time.

LiNbO$_3$ is currently the material that offers the most upside for broadband external modulation. LiNbO$_3$ is transparent over a wide wavelength range, from 0.33 µm to 4.5 µm, and can therefore support laser light emitted at 1550 nm. Comparing to electro-absorption materials, the optical propagation loss is very low and the optical coupling is efficient and relatively easy to implement. The typical optical insertion loss for LiNbO$_3$ modulators is in the 2-4 dB range. In addition, LiNbO$_3$ can handle very high optical power with an optical damage threshold at 250 MW/cm$^2$ [12]. Moreover, LiNbO$_3$ possesses a strong electro-optic coefficient along one of its axis, $z$, at $30.9 \times 10^{-12}$ V/m [13]. It is also very stable over a wide temperature range and extremely reliable over time. LiNbO$_3$ modulators have been used by the telecommunication industry for many years and have proved to be extremely durable. Moreover, LiNbO$_3$ modulation at frequency as high as 110 GHz has already been reported [14], thus demonstrating good potential for modulation deeper into the millimeter-wave region.

This chapter discusses the design of a LiNbO$_3$ phase modulator developed for millimeter-wave applications. In section 2.1, the Pockels electro-optic effect in LiNbO$_3$, allowing an electric field applied onto the crystal to change its refractive index, is reviewed. Section 2.2 introduces the concept of optical phase modulation in LiNbO$_3$ using the Pockels effect and demonstrates the impact of the modulating signal
frequency on the design requirements. Section 2.3 discusses the LiNbO$_3$ phase modulator design and the design techniques employed to achieve ultra broadband operation. Finally, section 2.4 presents the design of the LiNbO$_3$ modulator and simulated modulation efficiency of the device.

2.1 Pockels Electro-Optic Effect

The LiNbO$_3$’s crystalline structure geometry is trigonal and lacks inversion symmetry. Due to the non-central symmetry of the its crystal structure, LiNbO$_3$ demonstrates birefringence and electro-optic effect and displays ferroelectricity with large self-polarization along its z-axis as well as piezoelectric effect. The electro-optic effect corresponds to the mechanism by which the refractive index of a material is modified when an external electric field is applied to that material such that:

\[ n(E) = n_0 + \alpha_1 E + \alpha_2 E^2 + \cdots, \]  

(2.1)

where \( n \) is the refractive index of the material as a function of the applied electric field \( E \) and \( \alpha_i \) (\( i=1, 2, \ldots \)) are the electro-optic coefficients. The linear electro-optic effect \( \alpha_1 \) in equation (2.1), called Pockels effect, dominates the electro-optic response in LiNbO$_3$ and will therefore be used to modulate the optical signal.

Figure 2.1: Electro-optic modulation concept.
The index of refraction of LiNbO$_3$ can be represented as an ellipsoid whose equation, when the coordinates are chosen to lie along the LiNbO$_3$ principal axes, can be described as:

\[
\frac{x^2}{n_x^2} + \frac{y^2}{n_y^2} + \frac{z^2}{n_z^2} = 1 ,
\] (2.2)

where $n_x$, $n_y$ and $n_z$ are the index of refraction of LiNbO$_3$ along the $x$, $y$ and $z$-axes (Fig. 2.2).

![LiNbO3 ellipsoid with principal axes aligned with coordinate axes.](image)

**Figure 2.2:** LiNbO$_3$ ellipsoid with principal axes aligned with coordinate axes.

When an electric field is exerted on the LiNbO$_3$, the Pockels effect induces an index change by rotating the optical axis of the material. The general equation of the ellipsoid, non aligned anymore with the LiNbO$_3$ principal axes, becomes:

\[
\left(\frac{1}{n_x^2}\right)x^2 + \left(\frac{1}{n_y^2}\right)y^2 + \left(\frac{1}{n_z^2}\right)z^2 + \left(\frac{1}{n_x^2}\right)2yz + \left(\frac{1}{n_x^2}\right)2xz + \left(\frac{1}{n_x^2}\right)2xy = 1 ,
\] (2.3)
where the first three terms are the modified index changes from the original indices and the last three terms are related to the change in refractive index $\Delta n$. Without electric field, those last three terms are null. The effect of the electric field on the change of refractive can be describes as:

$$\Delta \left( \frac{1}{n^2} \right)_i = \sum_{j=1}^{3} r_{ij} E_j,$$  (2.4)

where $r_{ij}$ is the electro-optic coefficient along the $j$-axis of the material and $E_j$ is the electric field along that axis. With three electric field components orthogonal to the LiNbO$_3$ principal axes, the refractive index change induced in LiNbO$_3$, by an electric field can be represented by the following matrix system:

$$\begin{bmatrix}
\Delta \left( \frac{1}{n^2} \right)_1 \\
\Delta \left( \frac{1}{n^2} \right)_2 \\
\Delta \left( \frac{1}{n^2} \right)_3 \\
\Delta \left( \frac{1}{n^2} \right)_4 \\
\Delta \left( \frac{1}{n^2} \right)_5 \\
\Delta \left( \frac{1}{n^2} \right)_6
\end{bmatrix} = 
\begin{bmatrix}
0 & -r_{22} & r_{13} \\
0 & r_{22} & r_{13} \\
0 & 0 & r_{33} \\
0 & r_{51} & 0 \\
r_{51} & 0 & 0 \\
-r_{22} & 0 & 0
\end{bmatrix}
\begin{bmatrix}
E_x \\
E_y \\
E_z
\end{bmatrix},$$  (2.5)

where $r_{13}, r_{22}, r_{33},$ and $r_{51}$ are respectively equal to $9.6 \times 10^{-12}$ V/m, $6.8 \times 10^{-12}$ V/m, $30.9 \times 10^{-12}$ V/m and $28 \times 10^{-12}$ V/m. For optimum modulation efficiency, the electric field should be applied along the axis that offers the strongest electro-optic response. For LiNbO$_3$, the highest electro-optic coefficient is $r_{33}$. Thus, the electric field should be applied along the principal z-axis of the LiNbO$_3$ crystal according to equation (2.5). If the electric field components along the x and y-axis are considered null, the refractive index changes in equation (2.4) become:

$$\Delta \left( \frac{1}{n^2} \right)_1 = \Delta \left( \frac{1}{n^2} \right)_2 = r_{13} E_z,$$  (2.6)
\[
\Delta \left( \frac{1}{n^2} \right)_3 = r_{33} E_z ,
\]

\[
\Delta \left( \frac{1}{n^2} \right)_4 = \Delta \left( \frac{1}{n^2} \right)_5 = \Delta \left( \frac{1}{n^2} \right)_6 = 0 .
\]

The refractive indices in a LiNbO\(_3\) crystal when no electric field applied to it are:

\[
n_x = n_y = n_o ,
\]

\[
n_z = n_e ,
\]

where \(n_o\) and \(n_e\) are the ordinary index and the extraordinary index of LiNbO\(_3\), respectively. At the telecommunication wavelength of 1550 nm, \(n_o\) equals to 2.214 while \(n_e\) equals to 2.138 [15]. The modified indices of a LiNbO\(_3\) crystal, when an electric field is applied along its z-axis, can therefore be expressed as:

\[
\left( \frac{1}{n^2} \right)_1 = \left( \frac{1}{n^2} \right)_2 = \frac{1}{n_o^2} + r_{13} E_z ,
\]

\[
\left( \frac{1}{n^2} \right)_3 = \frac{1}{n_e^2} + r_{33} E_z ,
\]

\[
\left( \frac{1}{n^2} \right)_4 = \left( \frac{1}{n^2} \right)_5 = \left( \frac{1}{n^2} \right)_6 = 0 .
\]

Because the changes in refractive index \(r_{ij} E_j \ll \frac{1}{n^2}\), we can replace \((1 + n^2 r_{ij} E_j)^{-1/2}\) in equations (2.11) and (2.12) by the approximation \(1 - \frac{1}{2} n^2 r_{ij} E_j\). Thus, the new refractive indices of LiNbO\(_3\), when an electric field is applied along the z-axis of the crystal, can be approximated as:

\[
n_x = n_y \approx n_o - \frac{1}{2} n_o^3 r_{13} E_z ,
\]

\[
n_z \approx n_e - \frac{1}{2} n_e^3 r_{33} E_z .
\]
The refractive index ellipsoid representing the change of indices along the principal axes of the LiNbO$_3$ crystal is represented in Fig. 2.3.

2.2 Phase Modulation

2.2.1 DC Operation

The phase of an optical signal can be modulated when traveling through LiNbO$_3$ by changing the index of refraction of the material. As discussed in the previous section, this is accomplished by applying an electric field across the LiNbO$_3$ crystal and taking advantage of the Pockels effect. The phase of a light beam propagating along the $y$-axis of a LiNbO$_3$ crystal and polarized in the $z$-cut direction (TM mode) can be described at the output of the device as:
\[
\Phi = k_0(n_e + \Delta n)L = \frac{2\pi}{\lambda_0} \left( n_e - \frac{1}{2} n_e^2 r_{33} E_z \right)L, 
\]

where \( \Phi, k_0, \lambda_0 \) and \( L \) are respectively the phase, the phase constant, the optical light wavelength and the length of the interaction region between the electric field \( E_z \) and the optical field along the \( y \)-direction. From equation (2.16), the phase change induced by the Pockels effect can be expressed as:

\[
\Delta \Phi = \frac{1}{2} n_e^2 r_{33} E_z \frac{2\pi L}{\lambda_0}, 
\]

(2.17)

At low frequency, the electric field wavelength is much greater than the interaction length \( L \). Thus, the optical signal will experience a uniform voltage along \( L \) and equation (2.16) can be written as a function of the voltage:

\[
\Phi = \Phi_0 + \frac{\pi}{\lambda_0} n_e^2 r_{33} \frac{V}{d} L = \Phi_0 + \frac{V}{V_n}, 
\]

(2.18)

where \( V \) and \( d \) are the voltage across the LiNbO\(_3\) and the inter-electrode spacing between the signal electrode and the ground electrode, respectively (Fig. 2.4).

Figure 2.4: Phase modulation of an optical mode traveling through LiNbO\(_3\).
The half-wave voltage $V_\pi$, which corresponds to the voltage required to shift the phase $\phi$ by 180°, is a common figure of merit for modulators and is expressed for $z$-cut LiNbO$_3$ modulators as:

$$V_\pi = \frac{d\lambda_0}{n_e^2 r_{33} L}.$$  \hfill (2.19)

### 2.2.2 Low Frequency Operation

Assuming now that the electric field is sinusoidal in time, but at a frequency low enough that the electric field can still be considered uniform across the length $L$ of the interaction region, the phase of the optical mode becomes:

$$\Phi = \Phi_0 + \frac{\pi}{\lambda_0} n_e^2 r_{33} \frac{V}{d} L \sin(\omega_m t),$$  \hfill (2.19)

where $\omega_m$ is the angular frequency of the sinusoidal electric field, also called modulation frequency. At the modulator output, the electric field component of the modulated optical carrier can be derived as:

$$E = E_0 e^{j(\omega_0 t + \Phi_0)} e^{\frac{\pi}{\lambda_0} n_e^2 r_{33} \frac{V}{d} L \sin(\omega_m t)},$$  \hfill (2.20)

where $\omega_0$ is the angular frequency of the optical mode. By defining the modulation index $\Gamma_m$ as:

$$\Gamma_m = \frac{\pi}{\lambda_0} n_e^2 r_{33} \frac{V}{d} L = \frac{\pi}{V_\pi},$$  \hfill (2.21)

the electric field in equation (2.20) can be simplified as:

$$E = E_0 e^{j(\omega_0 t + \Phi_0)} e^{j\Gamma_m \sin(\omega_m t)}. \hfill (2.22)$$

By applying the Jacobi-Anger expansion of exponential function to equation (2.22), the electric field can be expressed as a series of Bessel functions of the first kind that represents a superposition of harmonics of frequency $\omega_m$:

$$E = E_0 e^{j(\omega_0 t + \Phi_0)} \sum_{n=-\infty}^{+\infty} J_n(\Gamma_m) e^{jn\omega_m t}. \hfill (2.23)$$
When developed, this last equation becomes:

\[ E = E_0 e^{j \Phi_m} \left[ J_0 (I_m) e^{j \omega_0 t} + \sum_{n=1}^{\infty} J_n (I_m) e^{j (\omega_0 - n \omega_m) t} + (-1)^n e^{j (\omega_0 + n \omega_m) t} \right] \]  \hspace{1cm} (2.24)

If the electric field is small enough to satisfy \( I_m \ll 1 \), only the dominant first order of the Bessel function needs to be taken into account as the higher orders can be considered null and equation (2.24) reduces down to:

\[ E = E_0 e^{j \Phi_m} \left[ J_0 (I_m) e^{j \omega_0 t} + J_1 (I_m) e^{j (\omega_0 - \omega_m) t} - e^{j (\omega_0 + \omega_m) t} \right] \]  \hspace{1cm} (2.25)

which can be developed into the form:

\[ E = E_0 J_0 (I_m) e^{j (\omega_0 t + \Phi_m)} + E_0 J_1 (I_m) e^{j \Phi_m} \left[ e^{j (\omega_0 - \omega_m) t} - e^{j (\omega_0 + \omega_m) t} \right] \]  \hspace{1cm} (2.26)

By adding an arbitrary phase constant term \( \Phi_m \) to the modulating signal such that:

\[ E_x = \frac{V}{d} e^{j (\omega_m t + \Phi_m)} \]  \hspace{1cm} (2.27)

the general expression of the phase modulated signal at the output of the LiNbO\(_3\) becomes:

\[ E = E_0 J_0 (I_m) e^{j (\omega_0 t + \Phi_m)} + E_0 J_1 (I_m) e^{j \Phi_m} \left[ e^{j (\omega_0 - \omega_m) t - j \Phi_m} - e^{j (\omega_0 + \omega_m) t + j \Phi_m} \right] \]  \hspace{1cm} (2.28)

Equation (2.28) shows that two new spectral components have been imprinted onto the original optical carrier. Those sidebands, namely upper sideband and lower sideband, are located at a frequency offset from the optical carrier \( \omega_0 \) equal to the modulating frequency \( \omega_m \). The optical power in the sidebands is scaled down to the square of \( J_1(I_m) \) when compared to the optical carrier. Because the voltage applied to the modulator is considered small \( (I_m \ll 1) \), \( J_1(I_m) \) can be approximated to \( I_m/2 \) while the power of the optical carrier remains almost the same with \( J_0(I_m) \sim 1 \). An illustration of the modulated optical spectrum, which contains the optical carrier with
lower and upper sidebands located at $\omega_0 - \omega_m$ and $\omega_0 + \omega_m$, respectively, is shown is Fig. 2.5.

![Intensity of the modulated optical field at the LN optical waveguide output. A lower sideband and an upper sideband are created at an offset frequency of $-\omega_m$ and $+\omega_m$, respectively, from the optical carrier frequency $\omega_0$.](image)

**Figure 2.5**: Intensity of the modulated optical field at the LN optical waveguide output. A lower sideband and an upper sideband are created at an offset frequency of $-\omega_m$ and $+\omega_m$, respectively, from the optical carrier frequency $\omega_0$.

### 2.2.3 RF Operation

To this point the optical mode was assumed to experience a constant electric field when traveling across the interaction region, although the electric field varies in time. This is because the optical transit time is a lot shorter than the electric field switch time at low frequency. This assumption remains true as long as the wavelength of the modulating electric field, $\lambda_m$, is much greater than $L$. However, when $\lambda_m$ becomes of the same order as $L$ like in the millimeter-wave region, the optical mode stops seeing a uniform electric field. The electric field in the RF spectrum is a propagating wave that most likely travels at a different speed than the optical mode. This velocity mismatch induces a strong decrease in interaction between the two signals. A particular electrical structure, the traveling-wave electrode configuration,
allows the RF effective index and the optical effective index to be matched to optimize the modulation efficiency.

In a traveling-wave modulator, the RF electrodes are designed to allow the RF signal to propagate down the modulator (y-direction) at a velocity identical to the optical mode’s velocity. A basic example of a traveling-wave modulator structure is shown in Fig. 2.6.

Figure 2.6: Traveling-wave modulator structure using microstrip.

At high frequency, the electrical signal of frequency $f_m$ is not uniform along the device. Instead, it propagates down the electrode as a wave at a certain velocity. Since the signal is a propagating wave, it is subject to frequency-dependent propagating losses $\alpha_m(f_m)$ inherent to RF signals. The general expression of the electric field $E_z$ propagating in a traveling-wave modulator is:

$$E_z(f_m, y, t) = E_0 e^{-\alpha_m(f_m)y} e^{-j2\pi f_m t}, \quad (2.29)$$

29
By taking into account the non-uniformity of the RF signal along $L$ and the RF losses, the expression of the phase change $\Delta \phi$ from equation (2.17) becomes at high frequency:

$$\Delta \Phi (f_m) = \frac{1}{2} n_e^3 r_{33} \lambda_0 \frac{2\pi}{\lambda_0} \int_0^L E_z (y, t, f_m) \, dy.$$  (2.30)

By replacing $E_z$ in equation (2.30) by its expression in equation (2.29), the phase change can be expressed as:

$$\Delta \Phi (f_m) = \frac{1}{2} n_e^3 r_{33} \lambda_0 \frac{2\pi}{\lambda_0} e^{-j 2\pi f_m t} E_0 \int_0^L e^{-\alpha_m (f_m) y} \, dy.$$  (2.31)

However, equation (2.31) does not account for the possible velocity mismatch between the RF signal and the optical mode. The voltage seen at any point along the electrode by a photon that enters the LiNbO$_3$ substrate at any time $t_0$ can be written as:

$$V (f_m, y) = V_0 e^{-\alpha_m (f_m) y} e^{j k_m (n_m - n_0) y} e^{-j 2\pi f_m t_0},$$  (2.32)

where $n_m$ and $n_0$ are the RF effective and the optical effective index, respectively, and $V_0$ is the amplitude of the RF wave at the coupling to the modulator input. By introducing the relative index mismatch $\delta$ between $n_m$ and $n_0$ such that $\delta = 1 - n_0/n_m$, equation (2.32) can be expressed as:

$$V (f_m, y) = V_0 e^{-\alpha_m (f_m) y} e^{j k_m n_m \delta y} e^{-j 2\pi f_m t_0}.$$  (2.33)

The induced optical phase change at the output of the modulator is derived by integrating the voltage over the interaction region:

$$\Delta \Phi (f_m) = \frac{2\pi}{\lambda_0} \left( \frac{1}{2} n_e^3 r_{33} \frac{1}{d} \right) V_0 e^{-j 2\pi f_m t_0} \int_0^L e^{-\alpha_m (f_m) y} e^{j k_m n_m \delta y} \, dy.$$  (2.34)

By computing the integral in equation (2.34) and by defining the mode overlap integral factor $\Gamma$ as the induced change in refractive index for a unit applied voltage across the electrode gap $d$ over a unit length such that $\Gamma = \left( \frac{1}{2} n_e^3 r_{33} \frac{1}{d} \right)$, the expression of the induced phase change becomes:
\[ \Delta \Phi(f_m) = \frac{2\pi}{\lambda_0} \Gamma V_0 e^{-j2\pi f_m t_0} \left[ e^{(-a_m + j k_m n_m \delta)k - 1} \right], \] (2.35)

The spatial distribution of the electric field and of the optical mode is one more factor affecting the modulation, as both signals are spatially varying functions in the optical waveguide cross section. For a given applied voltage \( V_0 \) over the transmission line, the field distribution can be described as \( E_m(x, z) \) for the RF signal and \( E_o(x, z) \) for the optical mode. The mode overlap integral factor \( \Gamma \) between the electric field and the optical mode is given in this case by:

\[ \Gamma = \frac{1}{V_0} \frac{\int \int \int \Re \{E_m(x,z)E_o(x,z)\} dx \, dz}{\int \int |E_o(x,z)|^2 dx \, dz}. \] (2.36)

To characterize the high-speed response of the modulator, we define the electro-optic response \( m(f_m) \), also called modulation depth, as the phase change \( \Delta \Phi(f_m) \) normalized to the DC phase change \( \Delta \Phi(0) \):

\[ m(f_m) = \frac{|\Delta \Phi(f_m)|}{|\Delta \Phi(0)|} = e^{-\frac{\alpha_0(f_m) L}{2}} \sqrt{\frac{\sinh (a_m L/2) + \sin^2 (k_m n_m \delta L / 2)}{(a_m L/2)^2 + (k_m n_m \delta L/2)^2}}. \] (2.37)

This last equation shows that the electro-optic response \( m(f_m) \) is strongly dependent on the RF loss coefficient \( \alpha_m(f_m) \) and on the index mismatch between the optical mode and the RF modulating signal. A strong electro-optic response will be obtained if index matching is achieved and if the RF propagating losses are kept low. The induced phase change at DC can easily be deducted from equation (2.34) as \( \Delta \Phi(0) = \frac{2\pi}{\lambda_0} \Gamma L V_0 \). The induced phase at high frequency can then be written from equation (2.37) as:

\[ |\Delta \Phi(f_m)| = \frac{2\pi}{\lambda_0} \Gamma L V_0 m(f_m) \] (2.38)

By substituting the induced phase \( \Delta \Phi(f_m) \) in equation (2.38) by \( \pi \), the half-wave voltage \( V_{\pi}(f_m) \) at high frequency can be expressed as:
Although a longer interaction length $L$ helps lowering the DC half-wave voltage $DC-V_{\pi}$, it has the opposite effect on $V_{\pi}(f_m)$ at high frequencies because it reduces the electro-optic response $m(f_m)$ due to the RF losses. One of the challenges is to determine the optimal length that results in relatively low $DC-V_{\pi}$ and a strong modulation depth. It can also be seen in equation (2.39) that a strong mode overlap between the electric field and the optical mode, described by $\Gamma$, increases the modulator’s efficiency at high frequency.

2.3 Broadband Operation Challenges

The modulator must be designed to optimize the electro-optic response $m(f_m)$ over the largest bandwidth possible. At a minimum, the modulator must be optimized over the 0-100 GHz frequency band to offer the best optical upconversion efficiency possible to the current passive millimeter-wave imaging systems that we are developing. The limiting factors can be of both optical and electrical nature.

From an optical design standpoint, the important parameters are optical mode confinement, optical propagation loss and optical coupling loss. They are directly related to the optical properties of LiNbO$_3$ and to the optical waveguide structure design. A lot of work on LiNbO$_3$ optical waveguide was conducted in the 1970s and 1980s and high quality optical waveguides in LiNbO$_3$ were developed [16-18]. The diffusion of titanium in the LiNbO$_3$ substrate is a popular optical waveguide fabrication method due to the friendly fabrication process and the low optical propagation loss achieved.

The real challenges for designing ultra broadband LiNbO$_3$ modulators have stemmed from the electrical properties of LiNbO$_3$ in the millimeter-wave domain. In
particular, the high effective index disparity of LiNbO$_3$ between the optical domain, with $n_o = 2.14$ at 1550 nm, and the RF domain, with $n_m = 6$, is of great concern at high frequency. This high index mismatch causes the RF signal to travel at a much lower pace than the optical mode. As seen in the previous section, a good index matching is critical to obtain high electro-optic response. The optical effective index of LiNbO$_3$ can only be increased slightly without dramatically increasing the optical propagation loss. Therefore, the only viable solution to match the velocity of the two modes is to lower the RF effective index.

RF propagation loss is also an important limiting factor for the modulator. The design techniques employed to lower the RF effective index and limit the RF propagation loss are discussed in detail in this section.

2.3.1 Index Matching

It has been established that the RF modulating signal and the optical mode need to propagate at the same speed for the modulator to operate efficiently in deep in the millimeter-wave spectrum. Co-propagation is achieved by matching the RF effective index with the optical effective index. For optimal operation, it was determined by that the RF index must actually be matched to the optical effective group index of the optical mode, which is 2.189 at 1550 nm, as opposed to the optical effective phase index of 2.138 used so far for LiNbO$_3$ modulators [19]. The strong disparity between RF index (~6) and optical index stems from the high dielectric constants of LiNbO$_3$ in the millimeter-wave domain ($\varepsilon_x = 44$, $\varepsilon_z = 28$) relative to the IR domain ($\varepsilon_x = 4.6$, $\varepsilon_z = 4.9$). This index mismatch has been for a long time the main limiting factor for the development of LiNbO$_3$ modulator in the millimeter-wave region. However, design techniques based on coplanar waveguide (CPW) ridged
structure have allowed the RF effective index to be reduced all the way down to the optical effective index [20].

Characteristics such as flexibility in design and integration, strong electric field confinement, straightforward fabrication process and highly tunable RF effective index give CPW the edge over microstrip as the structure of choice for guiding the millimeter-waves in LiNbO$_3$ traveling-wave modulators. The CPW structure consists of a central electrode, also called hot electrode or signal electrode, located between two ground electrodes. For z-cut LiNbO$_3$ traveling-wave modulators, the hot electrode needs to be located on top of the optical waveguide to align the electric field crossing the optical mode with the LiNbO$_3$ z-axis. A buffer layer between the CPW and the substrate is necessary to prevent optical mode attenuation caused by metal absorption from the hot electrode located directly above the optical waveguide. The strength of the electric field across the optical waveguide is mainly dictated by the thickness of the buffer layer and by the gap between the hot electrode and the ground electrodes. A narrow gap guarantees a high electrical field confinement. However, it also negatively impacts the RF conduction losses and the characteristic impedance of the CPW structure. The characteristic impedance of the CPW should be equal to 50 Ω to limit RF reflection loss at the modulator input when connected to RF sources, which are typically designed for 50 Ω impedance.

There are three design techniques that can be combined to lower the RF effective index from 6 down to 2.189. Each of them serves the purpose of reducing the impact of LiNbO$_3$, a high dielectric material, on the RF effective index. The first technique consists in introducing a thick buffer layer with low dielectric constant. That buffer layer must be thick enough to limit optical mode attenuation and to
considerably lower the RF effective index. However, it should also be as thin as possible to maintain a strong electric field across the optical waveguide. Silicon dioxide (SiO$_2$) is typically the material of choice due its low dielectric constant (3.9) and its mature deposition process. A SiO$_2$ layer of at least 600 nm is necessary to isolate the optical mode from the CPW structure. The second technique consists in designing tall electrodes to pull some electric field out of the LiNbO$_3$ and into the air, which can be seen as a material with a dielectric constant equal to unity. Increasing the CPW thickness also helps reducing the RF propagation losses. The downside of this technique is the lowering of the line impedance, which in turn increases the RF reflection losses. The third and final technique consists in etching away some LiNbO$_3$ material between the CPW electrodes to form a ridged CPW structure. In this configuration, some high dielectric constant material, LiNbO$_3$, is replaced by air. Therefore, the deeper and narrower the ridge is and the lower the RF effective index will be. However, the ridge should remain wide enough to support the hot electrode and preserve the optical mode.

2.3.2 RF Propagation Losses

Maintaining low RF propagation loss is critical to keep a strong interaction between the RF field and the optical mode along the modulator at high frequency. In addition to the RF substrate mode coupling loss, which will be discussed in details in Chapter 4, there are two more dominant RF propagation loss mechanisms that impact the modulator’s electro-optic response: the conduction loss and the dielectric loss. The propagation constant $\gamma$ of a propagating electromagnetic wave is described as:

$$\gamma = \alpha + j\beta,$$  \hspace{1cm} (2.40)
where $\alpha$ and $\beta$ are the real part and the imaginary part of the complex $\gamma$, namely the attenuation constant and the phase constant, respectively. The conduction loss and the dielectric loss are the two parameters defining the attenuation constant of a propagating wave. The attenuation constant is defined as a function of the RF frequency $f_m$ such that:

$$\alpha(f_m) = \alpha_c \sqrt{f_m} + \alpha_d f_m ,$$  \hfill (2.41)

where $\alpha_c$ (dB/cm/GHz$^{1/2}$) and $\alpha_d$ (dB/cm/GHz) represent the conduction loss and the dielectric loss, respectively.

As seen from equation (2.41), the conduction loss is proportional to the square root of $f_m$ whereas the dielectric loss varies linearly with $f_m$. Although the conduction loss does not escalate as fast as the dielectric loss when the modulating frequency increases, it typically dominates the attenuation loss in the millimeter-wave region, especially at the lower end of the spectrum. Several parameters affect the conduction loss: the width and height of the CPW structure, the gap between the electrodes and the CPW roughness due to skin effect. The conduction loss decreases as the gap increases. Therefore, although a narrow gap is needed to get a strong mode overlapping between RF mode and optical mode, its size will be limited by the conduction loss it induces. A trade-off between strong electrical field confinement and low conduction loss must be reached. Thick CPW electrodes combined with wide central electrode also help maintaining low conduction loss. Surface roughness in the CPW can also generate strong conduction losses due to the skin effect. Surface roughness becomes very critical at high frequency (smaller wavelength), as the defects appear more pronounced relative the RF signal’s wavelength and therefore generate more scattering losses.
The RF dielectric loss, low relative to the conduction loss for low frequency operation, can become a strong limiting factor when designing a broadband modulator due to its linear relation with frequency. Deep in the millimeter-wave spectrum, the dielectric loss has about the same negative impact on the RF propagation as the conduction losses. Dielectric loss can be minimized by pulling some of the propagating electric field component out of the LiNbO$_3$ and of the SiO$_2$ dielectric materials to force it to travel more through the air. Typically, this is achieved by combining three techniques. The first technique consists in simply increasing the buffer layer thickness, the second in increasing the electrode height, and the third in increasing the height and narrowing the width of the ridge. Increasing the buffer layer thickness must be performed with caution, as it comes at the expense of lowering the electric field confinement in the LiNbO$_3$ across the optical mode. It also reduces the RF effective index. Increasing the electrodes thickness results in lowering the impedance, and therefore increases the RF reflection losses. Regarding the ridge, it needs to support the CPW central electrode and must avoid cutting off the optical waveguide mode. Therefore, it can only be raised and narrowed by a certain amount. A trade-off between acceptable conduction loss, dielectric loss, electric field confinement and impedance matching must be reached.

2.4 Modulator Design

The modulator has been designed using a numerical simulator based on the finite-element method (FEM) [21]. The design of the modulator is optimized to provide the highest electro-optic response $m(f_m)$, also called modulation depth, over the largest bandwidth. The half-wave voltage $V_{\pi}(f_m)$ is another popular figure of merit used to characterize the modulation efficiency of modulators. In this study, we decide
to use mostly $V_{\pi}(f_m)$ to comment the modulator’s performance. However, $m(f_m)$ can easily be calculated from $V_{\pi}(f_m)$ using equation (2.39) if $DC-V_{\pi}(f_m)$ is known.

Figure 2.7: Cross-section design of the traveling-wave LiNbO$_3$ phase modulator.

The CPW electrodes thickness $T$, the center electrode width $S$, the gap $G$ between the hot electrode and the ground electrodes, the ridge width $R$, the ridge height $H$, the buffer layer thickness $B$ and the LiNbO$_3$ substrate thickness $D$ are the main design parameters affecting the RF properties of the modulator. Those
parameters can be easily modified in the simulator to adjust the modulator’s geometry and determine the combination offering the highest modulation efficiency.

The impact of the design parameters on the modulator’s electrical characteristics and modulation efficiency is summarized in Fig. 2.8. The priority is to match the RF effective index with the optical phase index at 2.189. The characteristic impedance of the CPW transmission line must be as close as possible to 50 Ω to match with the impedance of the RF source and limit the RF reflection losses. A thick CPW helps lowering the RF effective index and the conduction loss but also lowers the impedance. The conduction loss can be minimized with a large gap between electrodes. However, the electric field must be remained confined in the optical mode region as much as possible. Therefore, the gap should remain narrow enough for strong mode overlap. The ridge structure must be wide enough to prevent cutting off the optical mode but also narrow enough to allow index matching and low dielectric loss. Finally, the buffer layer must be thin to maintain strong mode overlap but also thick enough to prevent optical mode attenuation and to reach index matching.

<table>
<thead>
<tr>
<th></th>
<th>Conduction loss $\alpha_{\text{con}}$</th>
<th>Dielectric loss $\alpha_{\text{die}}$</th>
<th>Impedance $Z$</th>
<th>RF index $n_{\text{CPW}}$</th>
<th>Half-wave voltage $V_{\text{hes}}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Increase buffer thickness $B$</td>
<td>$\downarrow$</td>
<td>$=$</td>
<td>$\uparrow$</td>
<td>$\downarrow$</td>
<td>$\uparrow$</td>
</tr>
<tr>
<td>Increase CPW thickness $T$</td>
<td>$\downarrow$</td>
<td>$\downarrow$</td>
<td>$\downarrow$</td>
<td>$\downarrow$</td>
<td>$\uparrow$</td>
</tr>
<tr>
<td>Decrease CPW gap $G$</td>
<td>$\uparrow$</td>
<td>$=$</td>
<td>$\downarrow$</td>
<td>$\downarrow$</td>
<td>$\downarrow$</td>
</tr>
<tr>
<td>Increase ridge height $H$</td>
<td>$=$</td>
<td>$=$</td>
<td>$\uparrow$</td>
<td>$\downarrow$</td>
<td>$\downarrow$</td>
</tr>
</tbody>
</table>

Figure 2.8: Design parameters impact on the RF behavior of the modulator and on the overall modulation efficiency.
The design of the optical waveguide should allow the optical mode to be confined at the surface of the LiNbO$_3$, where the electric field is the strongest. The mode profile should also match the optical mode of single mode polarization maintained (PM) optical fibers to minimize the optical coupling losses at the modulator’s optical input and output. Parameters such as the width and thickness of the titanium (Ti) strip to be diffused in LiNbO$_3$ and the diffusion profile define the optical waveguide mode profile.

Independent electrical and optical analyses are performed to simulate the behavior of the electrical field and of the optical mode. The electrical analysis computes the electric field profile, the RF effective index, the RF propagation loss and the CPW characteristic impedance. The optical analysis computes the optical propagation mode. The overlap between the RF mode and the optical mode is then calculated from those analyses to yield the overall modulation efficiency of the device. The simulated electrical field and optical mode profile are presented in Fig. 2.9.

Figure 2.9: Modulator’s cross-section numerical analysis for (a) optical field and (b) electrical and magnetic field.
Ultimately, the final design is a trade-off between electric field confinement, impedance matching, RF propagation losses, optical mode confinement and mode overlapping between the RF field and the optical mode. It was determined through simulations that the effective index matching is the most critical factor for broadband modulation. Therefore, index matching at 2.189 has been made a requirement in terms of design and cannot be sacrificed to improve other characteristics.

According to the simulation, the highest modulation efficiency over the largest bandwidth is achieved when \( T, S, G, R, H, B \) and \( L \) are equal to 25 \( \mu \)m, 8 \( \mu \)m, 25 \( \mu \)m, 10.5 \( \mu \)m, 3.6 \( \mu \)m, 0.9 \( \mu \)m and 2 cm, respectively. The substrate is considered thin enough that substrate mode coupling does not occur. The modulator’s electrical parameters corresponding to this design, namely the RF effective index \( n_m \), the conductor loss \( \alpha_m \), the dielectric loss \( \alpha_m \) and the characteristic impedance \( Z_m \), are equal to 2.189, 0.29 dB/(cm·GHz\(^{1/2}\)), 0.0145 dB/(cm·GHz) and 45.6 \( \Omega \). The simulated modulator efficiency in terms of half-wave \( V_{\pi}(f_m) \) is represented in Fig. 2.10 over the DC-110 GHz band. The \( V_{\pi}(f_m) \) curve shows that the modulator achieves a DC- \( V_{\pi} \) of about 6.2 V and a 3 dB electrical bandwidth of 47 GHz. The 3 dB optical bandwidth can also be estimated at about 150 GHz by projecting the \( V_{\pi}(f_m) \), which is almost linear, beyond 110 GHz. If the modulator designed can be fabricated properly, it would become the fastest and most efficient LiNbO\(_3\) modulator ever reported.
Figure 2.10: Simulated half-wave voltage $V_\pi(f_m)$.

### 2.5 Notes on Launch Region Design

Two types of modulators have been designed for two different purposes. One was designed for operation over the full millimeter-wave spectrum, with the millimeter-wave signal launched onto the modulator through RF probes only. Another one was designed for wire bonding with an external RF source in an integrated module. An RF launch region has been designed for each type of modulator to maximize millimeter-wave efficiency. Typically, a launch region has larger central electrode and larger gaps between electrodes to fit with the test probe or allow wire bonding. For the modulator operating over the full millimeter-wave spectrum, the launch region was designed with $S$, $G$ and $R$ equal to 25 $\mu$m, 75 $\mu$m and 30 $\mu$m, respectively, for smooth RF mode transition from the probe to the CPW. The launch
region also includes a 75 µm long tapered CPW section between the end of the landing pad just described and the narrow interaction region. For the modulator to be integrated, a 90-degree bend was introduced in the launch region to provide an CPW access point for wire bonding. A wider central electrode is also required to match with the 50 µm wide ribbon. For this modulator, $S$, $G$ and $R$ in the launch region are equal to 50 µm, 50 µm and 55 µm, respectively, with a 350 µm-long tapered section. $T$, $H$ and $B$ are the same as in the interaction region for both types of modulator.

2.6 Conclusion

The design of an ultra broadband LiNbO$_3$ phase modulator was introduced in this chapter. The LiNbO$_3$ electro-optic properties and phase modulation principles were first studied to layout the design requirements for millimeter-wave modulation. The main challenges faced when designing an efficient broadband modulator have been presented. The biggest concerns for high frequency operation are the velocity mismatch between the modulating RF signal and the optical and the RF propagations losses. A design based on a ridged CPW structure with thick buffer layer and tall electrodes has been proposed to overcome those challenges. The design was optimized using numerical simulation, allowing a $DC-V_{π}$ of about 6.2 V. A Slow roll off in efficiency has resulted in a 3 dB electrical bandwidth of 47 GHz and a 3 dB optical bandwidth of 145 GHz
Chapter 3

MODULATOR FABRICATION PROCESS

The LiNbO$_3$ modulator fabrication process is presented in this chapter. The modulators are fabricated from 3 inch wide and 500 µm thick LiNbO$_3$ wafers. By optimizing the space available on a wafer, more than sixty modulators can be fabricated at a time. A summary of the fabrication process is presented in Fig. 3.1.

Titanium strips are first formed on top of a bare LiNbO$_3$ wafer (1) through UV lithography, Ti evaporation and lift-off (2), before being diffused into the substrate (3). Afterwards, the LiNbO$_3$ substrate around the diffused Ti optical waveguide is etched away to form a ridge structure (4). Subsequently, a buffer layer is deposited on top of the LiNbO$_3$ substrate and annealed (5). Metal seed layers are then deposited on the wafer in anticipation of the CPW structure fabrication (6). A thick CPW pattern is defined by lithography (7) before growing the electrode structure through electroplating (8). After removing photoresist seed layers (9), the LiNbO$_3$ wafer is diced at both end of the modulator array (10, 11). The optical inputs and outputs of the modulators are then carefully polished for optimum light coupling (12). Afterward, the modulators are diced into single chips (13) and each individual modulator’s backside substrate is eventually machined to prevent substrate mode coupling (14). Finally, optical fibers are pigtailed to the modulators on both ends (15).
Figure 3.1: Fabrication process flow chart.
3.1 LiNbO₃ Wafer Cleaning

The first step of the fabrication process is the cleaning and conditioning of the LiNbO₃ wafer in piranha solution (H₂SO₄:H₂O₂, 3:1) to remove any organic compound present on the wafer surface. Mastering this process is critical to the success of the overall fabrication process due to the risk of shattering the wafer and the number of times this process needs to be repeated. The risk of shattering comes from the piezoelectric nature of LiNbO₃. The chemical reaction between sulfuric acid (H₂SO₄) and hydrogen peroxide (H₂O₂) is quick and highly exothermic, with the temperature exceeding 100 °C in few seconds. If a LiNbO₃ wafer at room temperature is directly immersed in the hot piranha solution, the abrupt temperature change in LiNbO₃ induces a strong mechanical stress inside the crystal. This stress then leads to a rapid accumulation of electric charges in the crystal due to the piezoelectric nature of LiNbO₃. This combination of high stress and charge accumulation often results in the shattering of the LiNbO₃ wafer. To prevent this high stress, the temperature of the LiNbO₃ must be slowly ramped up to the piranha solution temperature. This is achieved using a hot plate initially set at room temperature and heated up to 90°C and with the LiNbO₃ wafer on top. When removed from the piranha solution, the LiNbO₃ wafer must be gradually cooled down in hot de-ionized (DI) water and not be rinsed directly under DI water at room temperature. The wafer typically needs to stay immersed in the piranha solution for about 20 min to obtain a pristine surface.

3.2 Optical Waveguide Diffusion

The optical waveguide is formed by thermally diffusing a Ti strip into the LiNbO₃ crystal. Several fabrication steps are necessary before the actual Ti diffusion can be performed. First, the waveguide structure needs to be defined. Second, a
precise amount of Ti has to be evaporated on top of the LiNbO$_3$ wafer to create the Ti strips. Finally, the photoresist is dissolved to leave only the Ti strip to be diffused on the top surface the LiNbO$_3$ substrate.

The UV contact lithography process employed to define the optical waveguide is the following. First, a 1 µm thick NR7-1500-PY negative lift-off photoresist layer from Futurrex Inc. is spun onto the wafer. Second, the photoresist is soft baked to evaporate the solvent and harden the film. The soft bake is a three-step process requiring two different hotplates to limit the stress in LiNbO$_3$ caused by its piezoelectric nature. This baking process is used throughout the entire LiNbO$_3$ modulator fabrication process, with parameters such temperature and time varying depending on the photoresist and the type of baking. The first step consists in a gradual but quick heating on a small hotplate, set originally at room temperature with the LiNbO$_3$ wafer on top and then heated up to the soft baking temperature of 135°C. The LiNbO$_3$ wafer is then transferred to a bigger, more accurate and more stable hotplate previously set a 135°C, where the photoresist soft bakes for 1 min, before being finally transferred back to a small hotplate for cooling. After soft bake, the wafer is aligned with the lithography mask under a mask aligner to guarantee that the optical waveguide is aligned along the $\gamma$-axis of the LiNbO$_3$ crystal. Once aligned, the mask is put into contact with the wafer and the photoresist is exposed. A post exposure bake identical to the soft bake is then performed before developing the photoresist in RD-6 Developer solution and rinsing the wafer in DI water.

Next, a 100 nm thick Ti layer is evaporated on top of the LiNbO$_3$ wafer. A lift off process is then used to remove the photoresist and form the Ti strips. This process consists in dipping the wafer into a solvent, typically acetone, to dissolve the
photoresist and detach the unwanted Ti from the wafer. After completion, only a 100 nm thick Ti strip remains on the LiNbO$_3$ surface.

The final step of the optical waveguide fabrication is the diffusion of the Ti strip into the LiNbO$_3$ substrate. The same heating principle as for the photoresist baking applies for the diffusion. The LiNbO$_3$ wafer is loaded into a furnace at room temperature. The temperature of the furnace is then slowly ramped up to about 1000 °C and stay there for about 10h. During the diffusion, oxygen is continuously flowing in the furnace tube to create a dry oxygen environment. After 10h, the furnace is slowly cooled down to room temperature. After removing the LiNbO$_3$ wafer from the furnace, the quality of the diffused waveguides is checked under microscope. If the waveguides are continuous in the regions of interest, the wafer is cleaned and conditioned in piranha solution for the next fabrication process step.

3.3 LiNbO$_3$ Ridge Etching

Once the optical waveguides have been diffused, the LiNbO$_3$ material surrounding the waveguides is etched away to create a ridge structure. The process is divided in two main steps. The first step is a UV contact lithography to define the etch mask. The second step is the actual etching of the LiNbO$_3$ material. Prior to the lithography, a thin Ti adhesion layer is deposited on the LiNbO$_3$ wafer.

The lithography process is performed using SU8-2007 negative photoresist from MicroChem Corp [22]. A 15 µm thick layer of photoresist is first spun onto the substrate followed by a soft bake at 85°C. Next, the ridge structure on the lithography mask is precisely aligned with the optical waveguides on the wafer before exposing the photoresist to UV light. The non-exposed SU8 photoresist defines the LiNbO$_3$ regions to be etched. It is recommended to use a narrow bandwidth optical filter
during exposure to increase accuracy, repeatability and resolution. After exposure, the photoresist is baked again at 85°C to cross-link the exposed area of the photoresist. Before developing the photoresist, a 24h relaxation period is required to release the stress in the thick photoresist film. After development in SU8-Development solution, a hard bake is performed to further cross-link the SU8 photoresist for etching.

The LiNbO$_3$ etching is performed in an Induced Coupled Plasma (ICP) Reactive Ion Etching (RIE) system in a chlorine environment [23] after mounting the LiNbO$_3$ wafer onto a loading wafer. A uniform layer of vacuum grease is used to bond the two wafers and help maintaining a homogeneous temperature across the LiNbO$_3$ substrate during the process, which is important to obtain uniform etching. A bonding process that releases most of the air trapped between the wafers in the vacuum grease has been developed to prevent the LiNbO$_3$ wafer from shattering during etching. Shattering occurs when too much stress is applied to the LiNbO$_3$ wafer. Already under a considerable amount of mechanical stress caused by RF radiations and temperature change during etching, the LiNbO$_3$ wafer can easily shatter when air bubbles trapped in the grease underneath it burst out due to the low-pressure environment inside the reacting chamber. The bonding process consists in depositing a small amount of vacuum grease at the center of the LiNbO$_3$ wafer on the backside, flipping the wafer onto a loading wafer and placing the assembly under vacuum in an oven at room temperature. By increasing the temperature in the oven to the melting temperature of the vacuum grease, the grease slowly spreads out across the substrate with most of the air bubble in the grease escaping from the sides and bursting out due to the low-pressure environment. Once a uniform and air-free grease layer is achieved, the assembly can be taken out of the oven and placed in an ICP RIE Samco system for
etching. After etching, the grease, the photoresist and the Ti seed layer are removed from the LiNbO$_3$ wafer in piranha solution.

### 3.4 Buffer Layer Deposition and Annealing

A 900 nm thick SiO$_2$ buffer layer is deposited in a Plasma Enhanced Chemical Vapor Deposition (PECVD) system to lower the RF effective index of the modulator down to the optical effective index. The deposition takes place at 300 °C and consequently a heat up phase and a cool down phase are incorporated in the deposition process to limit the stress in the LiNbO$_3$. After deposition, the SiO$_2$ buffer layer is annealed for 5h at 600 °C in a furnace to relieve the stress inside the SiO$_2$ crystal lattice and rearrange the SiO$_2$ crystalline structure into a more uniform layer. Following annealing, the wafer is cleaned once again in piranha solution.

### 3.5 Coplanar Waveguide Structure Fabrication

The CPW structure fabrication process can be divided into five steps: seed layers deposition, lithography, Ti seed layer etching, electroplating, and finally photoresist and seed layers removal.

Three seed layers are necessary to fabricate the CPW structure: a bottom layer of titanium tungsten (TiW) to adhere to the SiO$_2$ buffer layer, a thick intermediate layer of gold (Au) for growing the CPW's electrode, and finally a top layer of Ti for the photoresist adhesion. The layers of TiW, Au and Ti are respectively 60 nm thick, 180 nm thick and 55 nm thick and can be deposited using either sputtering or evaporation process.

SU8-2015 photoresist from MicroChem Corp. has been selected for this lithography because of its high-viscosity property that allows uniform 30 um thick
photoresist films to be deposited. This lithography process is very similar to the one used for ridge structure fabrication except for the photoresist spinning process, which is more complicated in this case because of the high viscosity of SU8-2015. For this lithography, the photoresist spreading and the spinning process need to take place in a solvent-filled environment to facilitate the spreading of the thick SU8. In addition, a rest period must be observed between these two events for optimal film thickness uniformity across the full wafer. The uniformity is critical for good contact between mask and photoresist during exposure.

After lithography, the Au seed layer is revealed by etching away the top Ti seed layer in hydrofluoric acid (HF) solution. During Ti etching, the wafer must be removed from the HF solution and immediately rinsed with DI water as soon as the exposed Ti layer has been completely etched. Keeping the wafer in the HF solution for too long allows HF to etch the Ti layer located underneath the SU8 photoresist, causing the photoresist to peel off and exposing more of the Au seed layer. When that occurs, the CPW electrodes become wider at the base than anticipated because gold grows on any open Au surface during electroplating. Ultimately, this negatively affects the RF properties of the modulator.

The TSG-250 gold solution from Transene is used as the source for growing 25 µm thick gold CPW electrodes. During electroplating, the LiNbO3 wafer immersed in the gold solution is connected through its gold seed layer to the anode of a current source. In the mean time, a conductive plate also immersed in the solution is connected to the cathode. An electrical current flowing from the conductive plate to the LiNbO3 wafer allows the Au+ ions from the solution to be deposited on the LiNbO3 wafer’s exposed gold surface. Once the electrodes have grown to the desired
thickness, the LiNbO$_3$ wafer is removed from the gold solution and immersed into a piranha solution to remove the SU8 photoresist and the top Ti seed layer. Subsequently, the Au seed layer and the TiW seed layer are etched away in gold etchant solution and TiW etchant solution, respectively, to reveal the SiO$_2$ buffer layer. An accurate 3D representation of a processed wafer is shown in Fig. 3.2.

![3D representation of a processed LiNbO$_3$ wafer containing sixty modulators.](image)

**Figure 3.2:** 3D representation of a processed LiNbO$_3$ wafer containing sixty modulators.

SEM pictures of an actual processed LiNbO$_3$ wafer containing sixty modulators before dicing, polishing and optical integration are presented in Fig. 3.3.
The modulator’s cross-section has also been measured under SEM to verify that the critical dimensions of the device in the interaction region are respected (Fig. 3.4). According to this SEM measurement, we are indeed capable of fabricating LiNbO$_3$ modulators that correspond precisely to the device designed.
Figure 3.4: SEM picture of the modulator’s cross-section.

3.6 Dicing and Polishing

To allow light coupling in and out of the modulators, the wafer must be diced into individual modulator chips and the modulators’ end faces must be polished. A high-accuracy automated Disco DAD322 dicing saw is used for creating individual LiNbO$_3$ modulator chip. Before dicing, a photoresist film easily dissolvable in acetone must be deposited on the wafer to protect the CPW structures and prevent unwanted particles to deposit between electrodes during dicing. The wafer is diced (Fig. 3.4) at a 5.5° angle with respect to the x-axis of the LiNbO$_3$ crystal to maximize mode coupling between the optical fibers, cleaved at an 8° angle, and the Ti optical waveguide in LiNbO$_3$ substrate according to Snell’s law. Dicing across the full wafer is possible because the modulators are by design vertically shifted from one another along a line at a 5.5° angle from the x-axis of the LiNbO$_3$ crystal, as seen in Fig. 3.2 and Fig. 3.5.
To polish multiple modulators at the same time, modulators are not diced into individual chips at this point yet but instead are kept together in big groups.

After dicing at a 5.5° angle, the end faces of the modulators are gradually polished down using diamond lapping film discs from Allied High Tech Products Inc. A rough polishing is first performed using a 30 µm rough lapping film to eliminate potential chips in the substrate from dicing. 9 µm, 1 µm and finally 0.5 µm rough diamond lapping films are then used to finish polishing the end faces. Once the substrate is polished, the modulators are diced into single modulator chips (Fig. 3.5) then thinned down to eliminate substrate mode coupling at high frequency. This micromachining process is discussed in details in Chapter 4.

![Figure 3.5: 3D representation of LiNbO₃ modulators after (a) wafer dicing and polishing and (b) individual chip dicing.](image)

3.7 Post Processing Tuning

Post-processing techniques have been developed to accurately match the RF index to optical index and lower the RF losses. Although the modulator is designed for precise RF index matching and low RF propagation losses, those parameters do not usually match exactly with what is expected from the simulation and are not exactly
the same for every modulators. These discrepancies appear not only between modulators from different wafers but also for modulators from the same wafer, although to a lesser extent. This inconsistency is caused by the difficulty to precisely control every parameters of the fabrication process. For instance, the homogeneity and quality of the SiO₂ buffer layer may slightly vary from one deposition to another, so does the quality of the CPW electrodes that are grown in an electroplating solution whose gold concentration always changes. It was found that the RF losses could be easily decreased, to a degree, by annealing the modulator for a few hours after the fabrication is completed. Regarding the RF effective index, it can be lowered by partially etching away the SiO₂ buffer layer in HF solution to replace SiO₂ with air. The amount of SiO₂ that needs to be etched on a specific modulator is directly dictated by the RF effective index measured for that device. A total etching of the SiO₂ between the CPW electrodes might even be required sometimes. Because the etching of SiO₂ in HF is isotropic, the buffer layer underneath the CPW structure gets also etched away during this process, causing an undercut CPW structure (Fig. 3.6). The severity of the undercut is directly related to amount of SiO₂ that is required to be etched. The drawback associated with this index matching technique is the weakening of the CPW mechanical structure. The edges of the CPW central electrode can become suspended in the air instead of being supported by the SiO₂ layer. If too much SiO₂ is etched for index matching, it is likely that the modulator will mechanically fail when RF testing or RF integration are performed.
The RF index tuning technique is a great tool to reach index matching but it has limitations. The technique does not work if the initial RF index is lower than the targeted index. Thus, it is critical to make sure during fabrication that the RF effective index always initially ends up higher than 2.189. This can be achieved by growing CPW electrodes thinner than the simulation results suggest. The mechanical robustness of the CPW structure also needs to be taken into consideration, as the CPW might become excessively fragile if too much SiO₂ is etched.

3.8 Optical Packaging

The last step of the fabrication process is the optical integration of the modulator with optical fibers. Polarization-maintaining (PM) fibers are necessary for this application to ensure that the electric field component of the optical mode coupling into the modulator is aligned with the z-axis of the LiNbO₃ crystal. The fiber pigtailed coupling light in and out of the modulator are v-groove terminated to facilitate optical bonding. They are aligned with the modulator optical waveguide using a semi-automatic Newport alignment system. The input fiber connected to a 1550 nm laser
source is first roughly aligned with the modulator along the $x$, $y$ and $z$ directions by observing the intensity of the optical mode at the modulator’s optical output. When some light is guided through the modulator, the output fiber is then roughly aligned with the modulator by searching the optical mode coming out using an optical power meter. When both input and output fibers are roughly aligned, an automatic alignment routine is launched to search for the positions at the input and output resulting in the highest optical coupling. Once those positions are found, the input fiber is backed up a few millimeters to make room to apply UV curable Norland Optical Adhesive 61 (NOA 61) on the v-groove’s face. The fiber is then brought back to its previous position, realigned one more time and then secured by shining UV light to precure the optical adhesive. To strengthen the assembly, more adhesive is later applied at the v-groove-modulator interface followed by an exposure to a high UV dose. The same process is repeated for the output fiber bonding. The optical integration process is finally completed by aging the optical adhesive at 50°C in an oven for 24h. An accurate 3D representation of an optically integrated LiNbO$_3$ phase modulator is shown in Fig. 3.7.

Figure 3.7: 3D representation of an optically integrated LiNbO$_3$ phase modulator.
Chapter 4
SUBSTRATE MODES

The RF propagation loss generated by the coupling of the CPW mode into RF modes supported by the dielectric substrate can considerably limit the operational modulation bandwidth of LiNbO$_3$ modulators [24]. Substrate mode coupling, also called surface mode coupling, usually becomes a real obstacle at high frequency, like in the millimeter-wave region, because the RF signal’s wavelength in the LiNbO$_3$ substrate is of the same order as or smaller than the substrate thickness [25-27]. The substrate modes supported by a dielectric substrate at a particular frequency are defined by its thickness, its dielectric properties and its boundaries. The impact of substrate modes on LiNbO$_3$ modulator’s performance has been studied in the past through simulations and experimental measurements but only for frequencies up to 50 GHz [28]. A LiNbO$_3$ substrate micromachining technique has been developed to thin the LiNbO$_3$ and prevent substrate mode coupling. The technique allows the operational frequency bandwidth of current LiNbO$_3$ modulators to be considerably extended. It was also proven to be both practical and reliable.

The first section of this chapter provides some background on substrate mode coupling mechanism and establishes the substrate thickness requirement for eliminating substrate mode coupling in LiNbO$_3$ over the full millimeter-wave region. In section 4.2, the impact of substrate modes on RF propagation in the millimeter-wave region is measured through LiNbO$_3$ modulator S-parameter characterization over the 0-220 GHz band. A micromachining technique to thin the LiNbO$_3$ substrates down to thickness required to completely eliminate the substrate modes in the
millimeter-wave region is presented in section 4.3. In section 4.4, the relationship between the substrate mode cutoff frequency and the substrate thickness is demonstrated experimentally through S-parameters measurements of a machined modulator. The impact of the RF launch region and of the surface roughness on substrate mode coupling is also studied.

4.1 Substrate Mode Coupling Mechanism

As the CPW mode propagates along the modulator on the transmission line, some of its energy leaks out into the substrate due to radiation loss. The first general rule for coupling between modes is that the mode coupling to another mode has to have a velocity equal or higher than the other mode. Another rule is that the two modes should not be orthogonal. In the case of the modulator, the second rule is usually satisfied as the CPW mode, which has the form of a quasi-TEM mode, and the modes supported by the substrate, such as the TE and TM modes, are typically not orthogonal. The first rule establishes that the coupling of the CPW mode into a substrate mode can only occur if the CPW mode propagates at the same speed or faster than the substrate mode. In other words, substrate mode coupling can only take place if the effective index of the CPW mode is equal or lower than the effective index of a substrate mode supported by the dielectric. Substrate mode coupling resonance occurs when the effective index of a substrate mode matches with the CPW mode effective index. The strength of the coupling also depends on the overlap between two modes. The strongest resonances are therefore the product of effective index matching and strong mode overlap between the CPW mode and substrate modes.

The electrical properties, the boundaries and the geometry of the dielectric substrate are the three parameters dictating the modes that can be supported by a
dielectric substrate. The velocity of those substrate modes is directly related to those parameters. At low frequencies, this velocity is higher than the CPW mode and therefore no coupling occurs. However, as the frequency increases, the velocity of the substrate modes decreases to the point where it is lower than the CPW mode’s velocity, allowing coupling. The velocity of a substrate mode varies with the frequency. The frequency at which the CPW mode starts coupling into a substrate mode is generally called the cutoff frequency. The relationship between the substrate thickness $d$ (mm) and the substrate mode cutoff frequency $f_c$ (GHz) has been established for LiNbO$_3$ as [29]:

$$f_c \approx \frac{119}{d}.$$  \hspace{1cm} (4.1)

Experiments on substrate modes coupling in fabricated LiNbO$_3$ modulators have been performed in the millimeter-wave region to verify the validity of this expression. The results are reported later in this chapter.

### 4.2 Substrate Mode Coupling Measurement

To understand the importance of eliminating the substrate modes in the operational band of interest, the transmission parameter $S_{21}$ of LiNbO$_3$ modulators has been measured in the 220 GHz. Three different experimental setups were used to characterize the modulator over such a large frequency band. The 220 GHz band was divided into three bands: 0-110 GHz, 110-170 GHz and 170 GHz to which correspond sets of probes, waveguides and Programmable Network Analyzer (PNA) extension modules. Therefore, small discontinuity in $S_{21}$ due to calibration errors are expected at 110 GHz and 170 GHz. The structure of the modulators tested corresponds to the design reported in Chapter 2 with a substrate thickness of 500 µm and a straight CPW
structure. The $S_{21}$ parameter of the modulator presented in Fig. 4.2 is a typical representation of the $S_{21}$ behavior measured for all the modulators fabricated.

![Graph](image)

Figure 4.1: Transmission parameter $S_{21}$ of a LiNbO$_3$ modulator suffering from strong substrate mode coupling in millimeter-wave region.

Based on $S_{21}$ measurement in Fig. 4.1, the CPW mode starts to slightly couple into the substrate at about 60 GHz, but the strong substrate mode resonances occur at 83 GHz and beyond, all the way across the millimeter-wave region. The strong attenuation induced by these substrate modes is a clear obstacle to efficient modulation deep in the millimeter-wave region. 25 dB attenuation dips can be observed at 133 GHz, 172 GHz and 220 GHz, while 10 dB loss is measured at the
resonances in W band (75-110 GHz). The optical upconversion efficiency for passive millimeter-wave imaging would significantly improve if those substrate modes can be eliminated.

4.3 Micromachining Process

A dicing technique has been developed to thin the LiNbO$_3$ substrate in order to prevent substrate mode coupling in the millimeter wave region. The machining process described below allows uniform and accurate thinning of LiNbO$_3$ substrates all the way down to 15 μm. Such thin substrates should theoretically eliminate all substrate modes up to about 800 GHz according to equation (4.1). This machining process is typically performed after a modulator is diced into a single chip and before optical integration with optical fibers. A high precision Disco DAD3220 programmable automatic dicing saw is used to perform the machining. The process is divided into three phases: the mounting of the modulator on a carrier wafer, the dicing of the substrate’s backside and finally the release of the modulator from the carrier wafer.

A proper mounting of the modulator on the carrier wafer is very critical to the success of the dicing process. The mounting determines the uniformity of the thinning and the accuracy of the substrate’s thickness. In addition, the modulator will likely crack during machining if not mounted correctly. The modulator is bonded onto a carrier wafer using wax dissolvable in acetone. Wax is first deposited on the carrier wafer surface located on hotplate set to the wax’s melting temperature on a hotplate. Next, the modulator chip is carefully flipped upside down on the thin layer of melted wax. Pressure is then uniformly applied to the backside of the modulator from atop to flatten the modulator with respect to the carrier wafer. Once a very thin and uniform
layer of wax is obtained between the two parts, the hotplate is turned off to allow the wax to slowly cool down to room temperature and harden. Pressure on the modulator must be maintained during that phase. After accurately measuring the thickness of the modulator-wax-wafer assembly and checking the uniformity of the bonding, the carrier wafer is finally attached to an adhesive tape and mounted onto the dicing saw.

The micromachining of the LiNbO$_3$ modulator consists in dicing the substrate underneath the CPW structure on the backside of device (Fig. 4.2). A 400 µm-wide blade is used for this process, although a narrower blade would also work. The height of the blade relative to the vacuum chuck during dicing sets the depth at which the blade penetrates the backside of the modulator’s substrate. Therefore, it is the blade height during dicing that will determine the substrate modes cutoff frequency. This height is manually entered in the dicing program and can be adjusted for any final thickness desired. The direction of travel of the modulator, and thus of the chuck, relative to the sense of rotation of the blade needs to be set properly to minimize the pressure applied by the blade to the substrate during dicing and prevent eventual cracks in the substrate. To alleviate the downward force exercised by the blade onto the substrate, the material removed by the blade needs to be evacuated upward and not downward between the blade and the substrate. Consequently, the chuck holding the modulator has to travel from the right side of the blade to the left if the blade rotates counter-clockwise, or from the left to the right if the blade rotates clockwise. Once the dicing process is completed, the assembly is immersed in an acetone bath for a few minutes to dissolve the wax and release the modulator from the carrier wafer.
SEM pictures of two modulators are presented in Fig. 4.3. One modulator has been thinned down to 100 µm for operation in the 0-110 GHz band while the other one has been thinned down to 20 µm for potential operation over the full millimeter-wave spectrum. In both cases, the LiNbO$_3$ substrate underneath the interaction region and underneath the RF launch region has been thinned down according to the technique presented in Fig. 4.2.
Overall, the modulators showed good mechanical resistance when handled with tweezers or when probed for electrical characterization. This technique is a big improvement over polishing methods used in the past in terms of performance, simplicity and reliability. The modulators showed no noticeable deformation after dicing even when thinned to 20 µm and the process only takes about 30 min, whereas LiNbO$_3$ substrates typically bend when polished due to the stress and therefore require bonding on a secondary substrate, which is a delicate process.

4.4 Experimental Results

4.4.1 Substrate Thickness Dependence

The modulator whose $S_{21}$ parameter is shown in Fig. 4.2 has been machined to experimentally study the effect of substrate thickness on substrate mode coupling. The
modulator substrate has been thinned down to 65 µm. To verify equation (4.1), which establishes a direct relationship between substrate thickness and substrate mode cutoff frequency, the $S_{21}$ parameter of this 65 µm thick modulator should show no sign of substrate mode coupling from DC all the way to about 183 GHz. Another modulator has also been machined but this time down to 30 µm, which should completely eliminate substrate mode coupling in the 0-220 GHz frequency band. The results of the $S_{21}$ transmission parameter measurements are presented in Fig. 4.4.

Figure 4.4: Effect of substrate thickness on substrate mode coupling cutoff frequency.
This $S_{21}$ transmission parameter measurement of the modulator when the substrate is only 65 µm shows that the substrate modes have been indeed eliminated in the frequency band predicted by equation (4.1). The first substrate mode resonance occurs at 184 GHz and substrate mode coupling begins at about 180 GHz, which is in good agreement with the 183 GHz bandwidth that was predicted. The $S_{21}$ measurement for the 30 µm thick modulator shows no sign of substrate mode coupling over the 0-220 GHz band. Therefore, as equation (4.1) has been experimentally verified, we can be confident that substrate modes will be completely eliminated in the millimeter-wave region if the substrate of the modulator can be thinned down below 39 µm.

### 4.4.2 Role of RF Launch Region in Substrate Mode Coupling

The impact of the launch region on substrate mode coupling has been studied in the past for LiNbO$_3$ modulators in the 0-50 GHz band [26]. It was reported that the launch region is the major contributor to substrate mode coupling and that the attenuation loss due to substrate mode could be greatly reduced by simply thinning the substrate underneath the launch region. The sections of the modulator that are forming the launch region are the wider CPW structure where the probes are landed for testing or where wires are bonded, the tapered CPW section between the wider CPW and the interaction region, and the 90-degree bend if there is one. The advantage of thinning the LiNbO$_3$ substrate only underneath the launch region is a mechanically stronger device and an easier micromachining process.

To verify if those results apply to the modulators we fabricated, a modulator with a straight CPW structure (optimized for modulation over full millimeter-wave spectrum) was thinned down to 100 µm only under the launch region. Theoretically,
The measurements of the transmission parameter $S_{21}$ of the same modulator before and after micromachining are shown in Fig. 4.5. It is obvious by comparing the two $S_{21}$ curves that the launch region has in fact almost no impact on substrate mode coupling. The substrate modes resonance dips when the modulator is micromachined under the launch region are not exactly as deep as when the launch region is still 500 µm. However, this small disparity is most likely caused by the reduction of the length where the CPW mode can couple into the substrate. This experiment shows that it is indeed necessary to thin the LiNbO$_3$ substrate under the entire CPW structure to eliminate the effect of the substrate mode coupling.

Figure 4.5: Impact of launch region on substrate mode coupling for straight CPW structure.
Another experiment on the impact of the launch region thickness on substrate mode coupling loss has been conducted, this time on modulators with 90-degree bends in the launch region for integration into RF photonics modules. The goal of this experiment is to determine whether or not it is necessary to thin the launch region in addition to thinning the interaction region. Ideally, we would like to keep the substrate underneath the launch region as mechanically strong as possible, and therefore thick, for wire bonding. Four micromachining configurations were tested on two different modulators in the 50-110 GHz band, which is where substrate mode coupling typically starts to occur. The two modulators considered here were fabricated on the same wafer and thus have very similar electrical properties and almost identical $S_{21}$ curves. The $S_{21}$ curves of a modulator without any machining, then with a 100 µm thin substrate under the interaction region and finally with a 100 µm thin substrate under the launch region as well are shown in Fig. 4.6. The $S_{21}$ curve of the second modulator thinned down to 100 µm only under the launch region is also presented in Fig. 4.6.

As expected, the best RF propagation is achieved when both the interaction region and the launch region are 100 µm thin. However, the difference made by thinning the launch region is not dramatic. Thinning the launch region helps smoothing out the $S_{21}$ curve overall, in particular in the 100-110 GHz band, but it should not be considered as a requirement for our modulators. The modulator that has only been thinned under the launch shows almost no improvement in $S_{21}$ compared to the modulator whose substrate was kept at 500 µm everywhere.
4.4.3 Surface Roughness

It has been established that the launch region does not impact the coupling of the CPW mode into the substrate modes supported by the LiNbO$_3$ modulator. However, the modulators with a straight CPW structure and the modulators with a 90-degree bend in CPW structure at launch.
degree bend in the launch region have very different $S_{21}$ curves although their design in the interaction region is exactly the same except for the electrode thickness. For those modulators presented in this chapter, the electrodes of the straight CPW structure are 25 µm thick whereas they are only 22 µm thick for the CPW with a 90-degree bend. However, the RF effective index for both types of modulators is matched around 2.19 so the difference in electrode thickness should not impact the substrate mode coupling. It can be observed from Fig. 4.5 and Fig. 4.6 that the coupling resonances of the CPW mode into the substrate modes occur at different frequencies. Moreover, the resonances are a lot stronger for the modulators with the straight CPW structure than for the modulators with the 90-degree bend. One explanation for this discrepancy is the difference in quality between the CPW structures. SEM pictures of the central electrodes of both types of modulators are shown in Fig. 4.7

Figure 4.7: SEM pictures of surface roughness of the CPW structure of (a) modulator with 90-degree bend in launch region and (b) modulator with a straight CPW structure.
In observing the electrodes in Fig. 4.7, it appears that the CPW of the modulator with a 90-degree appears is of much better quality than the CPW of the straight modulator. There are two major differences when looking at the surface of these two CPWs. The CPW surface of the straight modulator is lot rougher and also very uneven when compared with the one of the 90-degree bend modulator. Therefore, more electric field will be scattered by the CPW as the millimeter-wave propagates along the interaction region, leading to more millimeter-wave energy coupling into the substrate. The SEM pictures in Fig. 4.7 prove that smooth electrodes are necessary to avoid strong millimeter-wave scattering. Smooth electrodes can be fabricated by reducing the gold electrode growth rate during electroplating, which is achieved by reducing the intensity of the current flowing through the gold solution.

4.5 Conclusion

The strong RF attenuation generated by the coupling of the CPW mode into the substrate modes supported by the LiNbO$_3$ modulator has been measured in the 0-220 GHz band through the CPW’s transmission parameter $S_{21}$. To eliminate those substrate modes in the entire millimeter-wave spectrum, the thickness of the LiNbO$_3$ substrate must be lowered down to less than 39 $\mu$m. An original micromachining process allowing the LiNbO$_3$ substrate to be thinned down to any desired thickness has been presented and tested on actual devices. Substrate modes suppression was demonstrated through the $S_{21}$ measurement of a 65 $\mu$m thick modulator over the 0-220 GHz band. 20 $\mu$m thick LiNbO$_3$ modulators have been fabricated, which is thin enough to completely eliminate substrate mode coupling in the millimeter-wave region.
By using the micromachining process, it was also experimentally established that the launch region in our modulator design has minimum impact on substrate mode coupling. It is therefore not required to thin the LiNbO$_3$ substrate underneath the launch region if the modulator is to be integrated in an RF photonic module. Moreover, it was shown that surface roughness is a major cause of CPW radiation and should therefore be minimized for strong RF propagation.
Chapter 5

FULL SPECTRUM MILLIMETER WAVE MODULATION

The resolution of the passive millimeter-wave imager improves as the frequency of the millimeter-wave detected by the system increases, while the sensitivity of the system theoretically is enhanced when more of millimeter-wave energy can be detected and modulated. The development of a modulator capable of operating over the entire millimeter-wave spectrum would represent a big first step in the direction of the next generation of high-resolution and high-contrast passive millimeter-wave imaging systems. It would also help deploy faster optical networks for a telecommunication market that is always pushing for more bandwidth. Finally, knowing that an ultra-broadband modulator is available across the full millimeter-wave spectrum could help motivate researchers to keep innovating and developing more components for millimeter-wave applications at higher frequencies.

In this chapter, modulation over the full millimeter-wave spectrum is reported for the LiNbO\(_3\) phase modulator presented in chapter 2. The first section presents the modulator’s electrical characteristics, which consist of the transmission parameter, the RF effective index, the propagation losses, and the characteristic impedance. The modulator’s optical response is reported in the second section, with optical sideband characterization over the entire millimeter-wave band and half-wave voltage (\(V_{\pi}\)) measurement up to 170 GHz.
5.1 Electrical Characterization

5.1.1 S-parameters Measurement

The modulator has been tested over the entire millimeter-wave range through several sets of measurements in different bands of the spectrum. Due to the lack of equipment that would allow covering the entire millimeter-wave bandwidth in one measurement, the DC-300 GHz band has been divided based on the availability of the test equipment. The frequency bands are the following: 0-110 GHz, 110-170 GHz, 170-220 GHz and 220-300 GHz. In each band, a set of equipment consists of two RF probes, two PNA frequency extension modules and some RF waveguides. The electrical properties of the modulators are extracted by measuring in the 0-110 GHz band the S-parameters, namely the reflection coefficients $S_{11}$ at port 1 and $S_{22}$ at port 2 and the transmission coefficients $S_{21}$ and $S_{12}$.

To prevent the CPW mode to couple into the substrate modes at high frequencies, the LiNbO$_3$ substrate has been thinned down following the technique described in Chapter 4. As demonstrated in Chapter 4, the modulator should not support substrate mode in the 0-300 GHz band if the LiNbO$_3$ substrate is thinner than 39 µm. The modulator presented here has been thinned down to 30 µm. Such a thin substrate theoretically should eliminate substrate modes between DC and 396 GHz.

The transmission coefficient $S_{21}$ of this modulator has been measured over the 0-280 GHz range to experimentally demonstrate the suppression of the substrate modes (Fig. 5.1) in the millimeter-wave region. Issues with calibration above 280 GHz prevented an accurate $S_{21}$ measurement in the 280-300 GHz range. Therefore, $S_{21}$ above 280 GHz is not represented in Fig. 5.1. The switch between measurement setups for the different bandwidths can also be observed on the figure through the noticeable
small discontinuities and general allure of S₂₁ at 110 GHz, 170 GHz and 220 GHz, due to the calibration.

![Graph showing transmission parameter S₂₁ over the 280 GHz bandwidth]

Figure 5.1: Measured transmission parameter S₂₁ over the 280 GHz bandwidth. The transmission parameter S₂₁ confirms that the substrate modes have been suppressed for a substrate thickness of 30 µm.

5.1.2 Electrical Parameters Calculation

The electrical properties of the modulator have been extracted from a full S-parameters characterization using a transmission line ABCD matrix curve fitting technique. The ABCD matrix of the lossy CPW transmission line can be derived using the wave propagation theory as:
\[
\begin{bmatrix}
A & B \\
C & D \\
\end{bmatrix} =
\begin{bmatrix}
\cosh(\gamma l) & jZ_m \sinh(\gamma l) \\
jz_m^{-1} \sinh(\gamma l) & \cosh(\gamma l) \\
\end{bmatrix},
\]
where \(\gamma\) is the complex propagation constant, \(Z_m\) is the characteristic impedance of the CPW and \(l\) is the length of the CPW. The complex propagation \(\gamma\) constant is described as:
\[
\gamma = \alpha + j\beta,
\]
where \(\alpha\) is the attenuation constant (in Nepers/m) and \(\beta\) is the phase constant (in radian/m). Those constants are defined as:
\[
\alpha = \alpha_m \sqrt{f} + \alpha_d f,
\]
\[
\beta = 2\pi f \frac{n_{mmw}}{c},
\]
where \(\alpha_m\) and \(\alpha_d\) are the conduction loss and the dielectric loss, respectively, \(f\) is the frequency, \(n_{mmw}\) is the millimeter-wave effective index and \(c\) is the speed of light in vacuum.

Initial values are first attributed to the modulator’s electrical parameters \(\alpha_m, \alpha_d, Z_m\) and \(n_{mmw}\) to generate a first ABCD matrix. The S-parameters for a two-port network can then be backed up from the ABCD matrix through the relations:
\[
S_{11} = \frac{A + BY_0 - CZ_0 - D}{\Delta},
\]
\[
S_{12} = \frac{2(AD - BC)}{\Delta},
\]
\[
S_{21} = \frac{2}{\Delta},
\]
\[
S_{22} = \frac{-A + BY_0 - CZ_0 + D}{\Delta},
\]
\[
\Delta = A + BY_0 + CZ_0 + D,
\]
where \(Z_0\) is the characteristic impedance of the transmission line feeding the modulator (typically 50 Ω) and \(Y_0\) is its corresponding admittance (in siemens). The S-
parameters calculated from the estimated electrical parameters are then compared with
the measured S-parameters and optimized using a curve-fitting algorithm. From the S-
parameters measurement, the electrical parameters $\alpha_m, \alpha_d , Z_{mmw}$ and $n_{RF}$ of the
modulator have been estimated to be 0.28 dB/(cm·GHz$^{1/2}$), 0.01 dB/(cm·GHz), 47 $\Omega$
and 2.19, respectively. Those parameters are in good accordance with the simulation
results presented in Chapter 2 where $\alpha_m, \alpha_d , Z_{mmw}$ and $n_{RF}$ were calculated to be 0.29
dB/(cm·GHz$^{1/2}$), 0.0145 dB/(cm·GHz), 45.6 $\Omega$ and 2.19.

5.2 Modulation Characterization

5.2.1 Optical Modulation Spectrum Analysis

The phase modulator’s efficiency, defined as the ability of the modulating
electrical signal to modify an optical signal, can be determined by characterizing the
optical sidebands formed by the RF signal on each side of the optical carrier [30]. The
half-wave voltage $V_{\pi}$, which represents the modulating voltage necessary to change
the phase of the optical beam by $180^\circ$, is typically used as the figure of merit to
characterize the efficiency of modulators. It can be easily measured at low frequency
by configuring the modulator into a Mach-Zehnder interferometer and observing the
optical interference pattern on an oscilloscope through an optical detector while
varying the modulating voltage. However, detecting the interference pattern becomes
a lot more challenging at very high frequencies due to the lack of photodetectors’
bandwidth and sensibility at such frequencies. Instead, the half-wave voltage can be
measured by analyzing the optical modulation spectrum, specifically the optical
sidebands generated by the RF signal. The optical spectrum analysis leading to the
determination of the half-wave voltage in the millimeter-wave domain is presented below.

The optical power at the output of the modulator, driven by an RF signal of angular frequency \( \omega_m \), can be expressed in the frequency domain from equation (2.23) as:

\[
I(\omega_0 \pm k\omega_m) = I_0 J^2_k(\Gamma_m),
\]

(5.10)

where \( \omega_0, I_0, J_k \) and \( \Gamma_m \) are the angular frequency of the optical carrier, the power of the optical carrier, the Bessel function of first kind of order \( k \) and the modulation index, respectively. If the modulator is driven by a small voltage \( V_m \), the modulation index can be considered very small (\( \Gamma_m \approx \frac{\pi V_m}{V_n} \ll 1 \)), which leads to two approximations for first order Bessel functions: \( J_k(\Gamma_m) \approx 0 \) for \( k>1 \), and \( J_1(\Gamma_m) \approx \Gamma_m/2 \).

The optical power of the first sideband (\( k=1 \)) can be expressed as:

\[
I(\omega_0 \pm \omega_m) = I_0 J^2_1(\Gamma_m),
\]

(5.11)

or, by substituting \( J_1(\Gamma_m) \) and \( \Gamma_m \) as:

\[
I_{FSB} = I_0 \left( \frac{\Gamma_m}{2} \right)^2 = I_0 \left( \frac{\pi^2 V_m^2}{4V_n^2} \right).
\]

(5.12)

By replacing the modulating sinusoidal signal peak voltage \( V_m \) in equation (5.12) by the relation:

\[
I_m = \frac{V_m^2}{2Z_m},
\]

(5.13)

where \( Z_m \) and \( I_m \) are respectively the characteristic impedance of the CPW, which can be measured through the S-parameters characterization, and the power of the signal, the optical power of the first sideband \( I_{FSB} \) can be expressed as:

\[
I_{FSB} = I_0 I_m \left( \frac{\pi^2 Z_m}{2V_n^2} \right) = I_0 I_m \eta_{mod} \eta_{mod},
\]

(5.14)
where $\eta_{\text{mod}}$ (in $\text{W}^{-1}$) is defined as the modulation efficiency of the modulator.

Therefore, the modulation efficiency can be determined from equation (5.14) by either measuring $I_{\text{FSB}}$, $I_m$ and $I_0$ and using the relation:

$$\eta_{\text{mod}} = \frac{I_{\text{FSB}}}{I_0 I_m}, \quad (5.15)$$

or by measuring $V_\pi$ and $Z_m$ and using the relation:

$$\eta_{\text{mod}} = \frac{\pi^2 Z_m}{2 V_\pi^2}. \quad (5.16)$$

If $\eta_{\text{mod}}$ is determined using the first technique and equation (5.15), then $V_\pi$ can be deducted from $\eta_{\text{mod}}$ using equation (5.16) such as:

$$V_\pi = \sqrt{\frac{\pi^2 Z_m}{2\eta_{\text{mod}}}}, \quad (5.17)$$

The modulation efficiency can be easily read from the optical modulation spectrum if it is expressed in the logarithm domain (Fig. 5.2), where its expression in equation (5.15) becomes:

$$\eta_{\text{mod (dBW}^{-1})} = 10 \times \log \left( \frac{I_{\text{FSB}}}{I_0 I_m} \right) = I_{\text{FSB (dBW)}} - I_0 (\text{dBW}) - I_m (\text{dBW}). \quad (5.18)$$

The modulation efficiency can be simply obtained by subtracting the RF power to the difference in optical power between the optical carrier and the first order sidebands.

Very commonly are the powers expressed in dBm instead of dBW. In that case, a -30 term should be added to the expression to account for the transformation from dBm to dBW ($P_{\text{(dBW)}} = P_{\text{(dBm)}} - 30$). By normalizing the sidebands to the optical input power, the modulation efficiency can be further reduced to:

$$\eta_{\text{mod (dBW}^{-1})} = I_{\text{FSB (dBm)}} - I_m (\text{dBm}) - 30, \quad (5.19)$$

where $I_{\text{FSB}}$ corresponds to the optical power in the first order sideband normalized to the optical power of the carrier. In addition, if the sidebands are normalized to the RF

81
input power as well, the modulation efficiency in the logarithm domain can be directly interpreted as the difference in power between the optical carrier and the sidebands. The conversion efficiency can be expressed as:

$$\eta_{\text{mod}}(\text{dBW}^{-1}) = I_{\text{FNSB}}(\text{dBm}) - 30,$$

(5.20)

where \(I_{\text{FNSB}}\) is the optical power of the first order sideband fully normalized to the RF power and optical carrier power.

Figure 5.2: Sidebands normalization process. The optical power at the modulator output is measured (a), then normalized to the optical carrier power (b), and finally normalized to the RF power (c) where the modulation efficiency can be finally read in the logarithm domain.

5.2.2 Optical Sidebands Measurement

As a result of good index and input impedance matching, low propagation losses and very thin substrate, first order optical sidebands have been observed up to 300 GHz on the OSA. A picture of the modulator under test in the 220-300 GHz band is shown in Fig. 5.3. Optical insertion loss were measured at 3.7 dB for this device.
The modulator's response has been measured in 1 GHz increments. However, to better illustrate the sidebands, a modulation spectrum with a 5 GHz spacing between each RF frequency is represented in Fig. 5.4. In the figure, each pair of sidebands, upper and lower, corresponds to the modulator's optical response normalized to the optical carrier power and to the RF input power feeding the modulator, as shown in Fig. 5.2(c), at the corresponding RF frequency.
The probe and feed losses, harmonic generation in the RF source, and power meter limitations are all the factors accounted for in the optical sidebands normalization process. The vendor provided the insertion losses of the different probes employed, whereas the feed losses were determined by backing out the return losses from $S_{11}$ parameter measurements. The harmonic generation in the RF sources was determined directly by measuring the corresponding optical sidebands. However, limits in the available power of the RF source above 170 GHz and the measurement capabilities of power meters at these frequencies inhibited the accurate characterization of the modulator beyond 170 GHz. The RF power measured at the probes' inputs in each frequency bandwidth is represented in Fig. 5.5. The power measurements at the bandwidth transitions were adjusted, as necessary, to obtain a consistent and continuous modulation characterization.
5.2.3 Modulation Efficiency

The $V_z$ of the modulator can be extracted from the sidebands measurements using eq. 5.17. The $V_z$ measured using the normalized optical sidebands and the $V_z$ calculated using the S-parameters measurements are represented in Fig. 5.6 in the 0-170 GHz band, which corresponds to the bandwidth where the equipment allows accurate sidebands normalization.
Figure 5.6: Measured and calculated modulator half-wave voltage $V_\pi$. The measured $V_\pi$ extracted from the sidebands measurements is in agreement with the $V_\pi$ calculated using the S-parameters.

The good agreement in Fig. 5.6 between the $V_\pi$ measured from the normalized sidebands and the $V_\pi$ calculated from the S-parameters shows that the normalization process used to characterize the modulator is correct and that the modulation spectrum shown in Fig. 5.4 is accurate in the 0-170 GHz range. A $DC-V_\pi$ on the order of 8.6 V is obtained from extrapolating down to DC the measured $V_\pi$ represented in Fig. 5.6.

To experimentally measure $DC-V_\pi$, and verify that the previous extrapolation of DC- using high frequency measurements is correct, a Mach-Zehnder configuration has been setup where the optical input signal is split into two different fibers, with one fiber coupled to the modulator and while the other is not. The two optical signals, the
modulated signal propagating through the modulator and the non-modulated optical signal traveling around the modulator, are then recombined at the output of the modulator to interfere with each other. The intensity of the interference pattern, detected through a D400FC InGaAs photodetector, is observed on an oscilloscope. A function generator is used to apply a triangular electrical signal to the modulator’s CPW. The half-wave voltage can be directly observed on the oscilloscope by measuring the voltage difference necessary to shift the interference pattern by 180 degrees. On Fig 5.7, the voltage for a 360 degrees phase shift instead of 180 degrees is measured for a more accurate $DC-V_\pi$. The voltage measured on Fig. 5.7 is equal to 17.19 V. By taking half of that voltage, we obtain a $DC-V_\pi$ of about 8.6 V, which confirms the results obtained through the optical modulation spectrum analysis.

![Figure 5.7: Half-wave voltage DC-\(V_\pi\) measurement on oscilloscope.](image)
5.3 Conclusion

A modulator capable of modulation over the full millimeter-wave region is reported for the first time. By thinning the LiNbO$_3$ substrate to few tens of microns, the millimeter-wave radiation is able to propagate along the entire length of the CPW structure without coupling into the substrate. A smooth transmission coefficient has been measured over the entire millimeter-wave spectrum. The modulator has shown excellent index matching at 2.189, low propagation loss and good impedance match. As a result, the millimeter-waves are able to strongly interact with the optical mode over the entire length of the device. It was experimentally demonstrated that the modulator is capable of responding to modulation signals with frequencies as high as 300 GHz. The optical sidebands observed on an OSA are used to calculate the half-wave voltage $V_\pi$. A $DC-V_\pi$ of 8.6 V was measured by configuring the phase modulator in a Mach-Zehnder interferometer. The modulator’s optical insertion loss was measured at 3.7 dB. The 3 dB electrical bandwidth and the 3 dB optical bandwidth of the modulator are 47 GHz and 168 GHz, respectively, which is well above commercially available LiNbO$_3$ modulators or other LiNbO$_3$ modulators reported in the literature.
Chapter 6

ULTRA BROADBAND MODULATOR RF PACKAGING

Recent progresses in design and fabrication techniques have allowed LiNbO$_3$ phase modulators to operate well into the millimeter-wave region. Modulation up to 300 GHz, therefore covering the entire millimeter-wave spectrum, has been reported in this dissertation for the first time. However, this latest technology cannot be packaged yet into an integrated module for full millimeter-wave bandwidth operation due to the lack of available components. Flip-chip technology has previously been investigated for LiNbO$_3$ modulator packaging [31-33], and the packaging of LiNbO$_3$ modulators with K connectors has been reported [34]. However, those techniques were only reported for frequencies up to about 40 GHz. This is precisely the bandwidth to which commercially available packaged LiNbO$_3$ modulators are currently limited. Coplanar waveguide to rectangular waveguide transition on LiNbO$_3$ substrate have been reported, but this method limits the frequency to a particular band, here W band (75-110 GHz) [35]. A solution to increase the RF feed bandwidth of packaged modules is the 1.0 mm RF connector [36], which allows operations from DC up to 110 GHz. In order to extend the capacity of optical networks and provide more bandwidth for applications such as millimeter-wave imaging, we have developed a fully packaged LiNbO$_3$ modulator module that offers a 110 GHz bandwidth by integrating a 1.0 mm connector with a high-speed LiNbO$_3$ phase modulator.

In this chapter, the design of the module is first introduced, followed by the fabrication process of the modulator and the integration process of the different parts.
into the module. Finally, the RF characterization results of the module are presented. The RF insertion loss introduced by the packaging has been determined through the S-parameters measurement of the transition from the 1.0 mm connector to the modulator as well as through the comparison between the S-parameters of the modulator chip and of the packaged modulator. The module's optical modulation spectrum, normalized to the RF input power, has been measured over the entire 0-110 GHz band.

6.1 Module Design

The module consists of four different elements: a LiNbO$_3$ electro-optic phase modulator, a 1.0 mm input coaxial RF connector, a transition chip formed by a
coplanar waveguide on an alumina (Al₂O₃) substrate, and finally a housing into which the previous three elements are packaged (Fig. 6.1). In the module's design, the modulating RF signal is fed to the module through the 1.0 mm connector that is soldered to the Al₂O₃ transition chip connected to the LiNbO₃ modulator through ribbon bonds transition (Fig. 6.2).

Figure 6.2: Broadband module's integration design.

The design of the modulator used in the packaged module is almost identical to the modulator operating over the full millimeter-wave spectrum and presented in the previous chapter. One major difference between the two designs is the thickness of the
LiNbO$_3$ substrate. For the packaged modulator operating over 0-110 GHz band, the substrate only needs to be thinned down below 108 µm to prevent the coupling of the CPW modes into LiNbO$_3$ substrate modes in the intended operational bandwidth, whereas the substrate's thickness needs to be below 39 µm for operation over the full millimeter-wave spectrum. To maintain the modulator as mechanically strong as possible, the modulator's substrate needs to be kept as thick as possible. A LiNbO$_3$ substrate thickness in the order of 100 µm should be targeted to prevent substrate mode coupling and ensure strong mechanical properties. Another notable difference between the designs of the two modulators is the RF launch region. For the modulator to be packaged, a ribbon bonding region including a 90-degree bend has been designed to launch the RF signal from the source onto the modulator, whereas the CPW structure is straight with no bend for the full spectrum millimeter-wave modulator.

In order to achieve efficient modulation within 110 GHz bandwidth, it is critical to not only prevent the coupling of the CPW modes into LiNbO$_3$ substrate modes but to minimize the overall RF losses. Besides substrate modes coupling, the RF losses consist of the RF reflection loss, the dielectric loss and the conduction loss. The RF reflection loss is minimized by matching the modulator impedance to 50 Ω. An RF launch region that includes a ribbon bonding section, a tapered section and a 90-degree bend section has been designed to guide with minimum reflection the RF signal from the ribbons to the interaction region of the modulator (Fig. 6.3). In the ribbon bonding section, the signal electrode width $S_m$ and the gap $G_m$ between the signal electrode and the ground electrodes are both 50 µm. The length $L_m$ of the tapered section between the ribbon bonding section and the 90-degree bend is 350 µm.
The electrode signal width $S$ and the gap $G$ at the end of the tapered section, respectively 8 µm and 25 µm, correspond to the dimensions of the CPW in the interaction region of the modulator, where modulation occurs. The radius of the 90-degree bend is 100 µm. The ridged CPW structure designed for index matching in the interaction region has been extended into the RF launch region to reduce the dielectric loss and lower the RF index in that region. This low RF index helps reducing substrate mode coupling in the launch region where the LiNbO$_3$ substrate has been kept to its original thickness for packaging reasons. In addition, the ridge also improves the CPW mode transition from the launch region to the interaction region by keeping a similar CPW shape in those two regions.

Figure 6.3: Ribbon bonding transition design.
A gold CPW structure on a 150 \( \mu m \) thick \( Al_2O_3 \) substrate of purity 99.6 % has been designed and simulated on HFSS to achieve RF signal transition from the 1.0 mm RF connector to the \( LiNbO_3 \) modulator. In order to cover 110 GHz bandwidth, the Anritsu W1-103F 1.0 mm RF connector is selected for this application. The RF connector is specified with a 0.7 dB insertion loss and a 50 \( \Omega \) impedance. Therefore, a CPW characteristic impedance of 50 \( \Omega \) is required at both ends of the \( Al_2O_3 \) chip to minimize the RF return loss at the transitions from the 1.0 mm connector to the CPW on \( Al_2O_3 \) and from the CPW on \( Al_2O_3 \) to the modulator. However, the diameter of the 1.0 mm connector pin is 130 \( \mu m \), whereas the width \( S_m \) of the CPW signal electrode of the modulator in the launch section is 50 \( \mu m \). Consequently, the geometry of the CPW on the \( Al_2O_3 \) chip must contain a taper section to ensure a continuous mode matching at both transitions (Fig. 6.3). On the 1.0 mm connector end of the CPW, a signal electrode of width \( S_a \) equal to 150 \( \mu m \) and a gap \( G_a \) between the signal electrode and the ground electrodes of 60 \( \mu m \) have been designed, whereas \( S_t \) is 65 \( \mu m \) and \( G_t \) is 35 \( \mu m \) on the modulator end. The length \( L_t \) of the taper section is 150 \( \mu m \). The thickness of the CPW has been set to 10 \( \mu m \) to comply with the impedance requirement. The total width and length of the \( Al_2O_3 \) chip are 1.4 mm and 2.19 mm, respectively. The HFSS simulation results of the RF transition from the 1.0 mm connector to the modulator end of the \( Al_2O_3 \) chip are shown in Fig. 6.8.

The ribbon bond transition between the \( Al_2O_3 \) chip and the \( LiNbO_3 \) modulator has been designed to operate over the 110 GHz bandwidth based on published work [37]. Ribbon bond is preferable to wire bond for wideband operation because of the lower impedance provided by its wide cross-section. In order to keep the impedance low, the length \( L_w \), the height \( H_w \) and the pitch \( P_w \) of the ribbon bond must be
The quality of the CPW deposition, the chips placement and the wire bonding equipment are limiting factors to the ribbon bond transition design. A ribbon bond transition with $L_w$, $H_w$ and $P_w$ equal to 250 µm, 30 µm and 50 µm, respectively, is the best configuration achieved consistently for this module. RF losses under 0.6 dB in the 1-50 GHz range and under 2 dB in the 50-110 GHz range are expected for this configuration [37].

A compact housing (13.3x9x50 mm) has been designed to integrate the three components previously introduced. A top housing part, represented in Fig. 6.4, adds 2 mm to the overall height of the module presented in Fig. 6.2 when the housing is fully assembled. The typical housing material of choice for applications in the millimeter-wave domain is Kovar because of the coefficient of thermal expansion (CTE) compatibility between Kovar and $\text{Al}_2\text{O}_3$, which is about $6 \times 10^{-6}$ ppm/°C for both
materials. Kovar also offers good matching with LiNbO$_3$ who has a CTE of about 13×10$^{-6}$ ppm/°C. Given the thickness of the Al$_2$O$_3$ substrate, a 700 µm wide and 500 µm deep groove has been machined in the housing where the Al$_2$O$_3$ substrate fits to prevent the CPW modes on the Al$_2$O$_3$ chip from coupling into conductor-backed CPW (CBCPW) modes. To reduce the loss at the wire bonding transition between the Al$_2$O$_3$ chip and the LiNbO$_3$ modulator, the housing was designed to ensure that both components are precisely aligned and leveled when mounted. Alignment tolerance analysis has been reported for CPW transition between LiNbO$_3$ substrate and Al$_2$O$_3$ substrate using ribbon bonding [37]. Results from the theoretical analysis show that a 20 µm lateral misalignment between the CPWs has no significant effect on the RF transmission performance with only a 0.1 dB drop-off at 100 GHz. They also show that the RF transmission difference between a 25 µm air gap between the two chips and a 75 µm air gap is only 0.3 dB at 100 GHz.

6.2 Integration Process

Before integrating each element into the housing, a 30 nm thick titanium layer followed by a 200 nm-thick gold layer are evaporated on the kovar housing for adhesion concerns when soldering the CPW ground electrodes of the Al$_2$O$_3$ chip to the housing ground. The Al$_2$O$_3$ chip is the first element to be integrated. Once the CPW structure on the Al$_2$O$_3$ substrate is fabricated following conventional UV lithography and electroplating techniques, an Al$_2$O$_3$ chip is precisely diced to fit into the housing and attached to it using thermally curable silver epoxy. The 1.0 mm connector is then carefully inserted into the housing and tightly screwed to prevent any looseness. Subsequently, the CPW signal electrode and the ground electrodes are soldered to the 1.0 mm connector pin and to the housing ground, respectively, as shown in Fig. 6.2. A
no-clean low temperature indium-based solder paste Indalloy 2 from Indium Co. is employed for this operation. The length of the solder transition from the 1.0 mm connector pin to the CPW is about 400 µm (Fig. 6.5). Finally, the modulator is integrated into the housing and attached using UV curable epoxy after being precisely aligned with the Al$_2$O$_3$ transition chip. The air gap between the Al$_2$O$_3$ chip and the LiNbO$_3$ modulator and the misalignment between the CPWs are typically under 50 µm and 5 µm, respectively. Both CPW structures on the modulator and the Al$_2$O$_3$ substrate are then ribbon bonded on a F&K Delvotec Desktop Micro Factory Wire Bonder using a 50 µm wide and 25 µm thick gold ribbon. The length $L_w$ and the height $H_w$ of the ribbon bonds are typically around 250 µm and 30 µm, respectively. A picture of the fully integrated module connected to the RF source through a 1.0 mm cable is shown in Fig. 6.5.

Figure 6.5: Packaged module connected to the RF source through a 1.0 mm cable.
6.3 Module Characterization

6.3.1 Modulator RF Characterization

Before integrating all the different elements in the housing and measuring the performance of the module, the LiNbO$_3$ modulator and the Al$_2$O$_3$ transition chip have been characterized separately over the full 0-110 GHz bandwidth. An Agilent E8361C programmable network analyzer (PNA) connected to Agilent N5260 T/R waveguide modules with 1.0 mm test ports were used to measure the S-parameters. A full 2-port calibration of the test environment was performed using a short-open-load-through (SOLT) scheme at the tips of ground-signal-ground (GSG) 110 GHz GGB Industries probes. The probes were connected to the 1.0 mm test ports through 1.0 mm cables. The measured reflection parameter $S_{11}$ and transmission parameter $S_{21}$ of the modulator to be integrated are shown in Fig. 6.6.

The RF effective index $n_{RF}$, the conduction loss $\alpha_c$, the dielectric loss $\alpha_d$, the characteristic impedance $Z$ and the optical insertion loss $L_{opt}$ of the modulator were respectively measured at 2.19, 0.25 dB/(cm⋅GHz$^{1/2}$), 0.027 dB/(cm⋅GHz), 39.4 $\Omega$ and 2.8 dB. Results from the S-parameters measurements of the 2.05 mm long CPW structure on the Al$_2$O$_3$ transition chip showed a RF transmission loss of about 0.3 dB over the 110 GHz bandwidth.
6.3.2 Insertion Loss Measurement

After integrating the Al₂O₃ chip with the 1.0 mm connector into the housing, the S-parameters of the 1.0 mm connector-to-modulator transition have been measured. However, this measurement could not be performed over the full 0-110 GHz band because of a lack of 1.0 mm connector RF calibration kit. Instead of connecting the PNA to the RF input of the module through a simple 1.0 mm cable, we used WR-10 waveguides with a 1 mm WR10-W1 waveguide adapter from Anritsu (Fig. 6.7). Therefore, the S-parameters were measured only in the 70-110 GHz band. A 110 GHz GGB Industries probe was connected to the Al₂O₃ chip on the other end of the module. A short-offset/short/load-offset/load calibration at the WR-10 waveguide on the 1.0 mm connector end of the module and a short/open/load calibration at the tip...
of the GSG 110 GHz probe were performed. A through calibration of the transition, by connecting the WR-10 waveguide to the 1.0 mm connector through the 1 mm WR10-W1 waveguide adapter and by landing the probe on the Al₂O₃ chip end, completed the calibration process.

![Image](image.png)

**Figure 6.7:** Setup for module's RF insertion loss measurement.

The insertion loss of the 1 mm WR10-W1 waveguide adapter, equal to about 0.5 dB over the 70-110 GHz band, has been measured independently and was subsequently subtracted from the S₂₁ measurements performed for the transition and for the full module. The measured and simulated S₁₁, S₂₂ and S₂₁ parameters of the transition are represented in Fig. 6.8. The S-parameters for the transition were
simulated using HFSS. The measured $S_{21}$ shows an RF insertion loss for the module varying between 1 dB and 2.5 dB, which is in good agreement with the simulation. An impedance mismatch in the 95-100 GHz range can be observed through the strong increase of the RF reflection coefficients $S_{11}$ and $S_{21}$ and the decrease of the transmission coefficient $S_{21}$. This mismatch is also in accordance with the simulation, although the simulation predicted a smaller mismatch over a narrower frequency range centered at 100 GHz.

Figure 6.8: Simulated and measured S-parameter characteristics of the 1.0 mm connector-$\text{Al}_2\text{O}_3$ chip transition.
6.3.3 Module RF Characterization

A fully packaged module has been tested after integrating the modulator reported in Fig. 6.6 in the housing. The S-parameters of the module have been measured in the 70-110 GHz band by following the same technique used for the transition. The S-parameters $S_{11}$ and $S_{21}$ of the module, along with the $S_{11}$ and $S_{21}$ previously reported in Fig. 6.8 for the modulator, are represented in Fig. 6.9.

Figure 6.9: S-parameter characteristics of modulator and module.

The actual RF insertion loss of the module can be extracted from Fig. 6.9 by simply comparing the two $S_{21}$ curves, as the difference between the $S_{21}$ transmission coefficients corresponds to the module's insertion loss. According to these
measurements, the insertion loss varies between 1 dB and 2.5 dB in the 70-110 GHz bandwidth except in the 95-100 GHz band where it is comprised between 3 dB and 5 dB. There are therefore some disparities between the RF insertion loss measured in Fig. 6.8 and the one extracted from Fig. 6.9. However, they were expected. The total RF insertion loss measured in Fig. 6.9 includes the RF insertion loss from the 1.0 mm connector to the Al₂O₃ chip measured in Fig. 6.8 and also the losses introduced by the ribbon bond transition from the Al₂O₃ chip to the modulator. Therefore, the difference between the RF insertion losses measured in Fig. 6.8 and in Fig. 6.9 corresponds to the loss introduced by the ribbon bond transition. The ribbon bond transition loss can be estimated from Fig. 6.9 between 0.5 dB and 1.5 dB, which is in good agreement with the results reported in [37]. The strong insertion loss in the 95-100 GHz range can be explained by the large impedance mismatch at the RF transition from 1.0 mm connector to the CPW on the Al₂O₃ chip observed in Fig. 6.8 in this same frequency range. The ribbon bond transition only accentuates that mismatch. This large impedance mismatch can also be observed in Fig. 6.9 through the $S_{11}$ reflection coefficient measured for the module. Overall, the module performs in agreement with the modulator's measurements reported in Fig. 6.6 and the RF insertion loss measured in Fig. 6.8.

6.3.4 Modulation Efficiency of Packaged Modulator

The optical modulation spectrum has been measured over the 0-110 GHz band by connecting the PNA to the module using a 1.0 mm cable and launching a 15 dBm optical signal from a 1550 nm laser source into the input fiber of the module (Fig. 6.10).
The resulting modulated light was observed on a Yokogawa AQ6319 optical spectrum analyzer (OSA) at the optical output of the module. Each pair of first order optical sidebands represented in Fig. 6.11, namely upper sideband (USB) and lower sideband (LSB), corresponds to the conversion of the RF energy into the optical domain. The optical sidebands in Fig. 6.11 have been normalized to the module's RF input power, which has been measured for each frequency on an Agilent E4418-B power meter at the 1.0 mm cable output. A 3 GHz RF input increment is used to represent distinctively in Fig. 6.11 each optical modulation sidebands and avoid overlaps.
Figure 6.11: Optical modulation spectrum.

The half-wave voltage $V_\pi$, a widespread figure of merit for modulators, has been indirectly measured over the 0-110 GHz band. This $V_\pi$ measurement is compared with the theoretical $V_\pi$ calculated based on the geometry of the fully packaged module (Fig. 6.12). The measured $V_\pi$ is backed out at high frequencies from the $V_\pi$ measurement at DC and the normalized optical modulation sidebands using equation (5.18) and equation (5.17) presented in the optical modulation spectrum analysis in chapter 5. The theoretical $V_\pi$ is obtained based on three different simulations. The theoretical $V_\pi$ of the modulator alone is first simulated using the finite element method (FEM). Then the total module RF insertion loss is calculated by adding together the simulated RF insertion loss $S_{21}$ of Fig. 6.8 and the simulated ribbon bond RF insertion loss from [37]. The module's theoretical $V_\pi$ can then be determined by replacing the power delivered to the modulator by the RF power delivered to the full module in equation (5.18).
A $V_\pi$ between 15 V and 20 V has been measured in W-band (Fig. 6.8). Overall, a $V_\pi$ of about 10 V can be measured at 10 GHz whereas it is in the order of 20 V at 110 GHz. A DC-$V_\pi$ of 8.5 V has been directly measured using a Mach-Zehnder interferometer configuration. The 3 dB electrical bandwidth and the 3 dB optical bandwidth of the module are about 45 GHz and 90 GHz, respectively. Those results are not in total agreement with what the theoretical data predicted. A difference of about 2.5 V can be observed across the 0-110 GHz band. According to the FEM simulation of the modulator without packaging, a 6.2 V DC-$V_\pi$ should be obtained. Instead, the DC-$V_\pi$ measured was 8.5 V, which represents a 2.3 V difference with the simulation. This $V_\pi$ difference in Fig. 6.12 is due to the modulator's higher dielectric loss and lower impedance than expected and is not the result of the packaging, except
in the 95-102 GHz range. The high $V_\pi$ measured in this bandwidth reflects the low transmission coefficient $S_{21}$ and the impedance mismatch reported in Fig. 6.8 in that same frequency range. The spike in $V_\pi$ in the 68-74 GHz range represents the combination of poor RF calibration and strong RF harmonics generation from the RF source in that range. The power sensors used for the RF source power measurements only had calibration available in the 7-67 GHz and 75-110 GHz bands.

6.4 Conclusion

A broadband LiNbO$_3$ modulator has been packaged for the first time with a 1 mm coaxial connector for operation over the 0-110 GHz band. An RF transition between the 1 mm connector and the modulator has been designed and characterized. Low RF insertion loss in the 1-2.5 dB range has been measured in the 70-110 GHz band, except in the 95-100 GHz where higher RF reflection loss is experienced. Overall, the modulator integrated into the module offers a DC-V of 8.5 V for a 45 GHz 3 dB electrical bandwidth and a 90 GHz 3 dB optical bandwidth.
Chapter 7

CONCLUSION

In this chapter, I present a summary of the work discussed in this dissertation and emphasize the major accomplishments. Following that, I introduce new designs being developed to further improve the efficiency of LiNbO$_3$ modulators.

7.1 Summary

In this dissertation, I have demonstrated for the first time optical modulation over the full millimeter-wave spectrum using a LiNbO$_3$ phase modulator. The modulator was developed as part of our effort to build at the University of Delaware passive millimeter-wave imaging systems. The architecture of the system is based on optical upconversion of native blackbody radiations detected by a distributed aperture antenna array and emitted by the scene in the millimeter-wave domain.

LiNbO$_3$ electro-optic phase modulation principle based on Pockels effect was presented at millimeter-wave frequency. LiNbO$_3$ material is very attractive for broadband modulation because it is transparent at 1550 nm, it offers good optical coupling, it can handle very high optical power, it has a strong electro-optic coefficient along its z-axis at $30.9 \times 10^{-12}$ V/m, it is very stable over a wide temperature range and extremely reliable over time. However, its RF effective index and optical effective index are far apart and required to be matched for high modulation efficiency. A design was presented to match those indices based on CPW ridge structure. A post-processing tuning technique was developed to precisely match the RF effective index
to the optical effective index at 2.189. In addition, a LiNbO$_3$ micromachining technique was also developed to eliminate substrate mode coupling and extend the modulation bandwidth of LiNbO$_3$-based modulators. Through this technique, it was experimentally demonstrated that most substrate mode coupling occurs along the interaction region of the CPW. As a result, it is not required the thin the modulator’s substrate under the launch region for devices to efficiently operate deep into the millimeter-wave region, although thinning the launch region helps improving the performances.

The modulator has been fully characterized accurately over the 0-170 GHz and optical modulation sidebands were observed over the entire millimeter-wave region. The modulator shows excellent modulation efficiency with a 3 dB electrical bandwidth of 47 GHz and a 3 dB optical bandwidth of 168 GHz. The half wave voltage $V_\pi$ at DC was measured at 8.6 V.

Finally, the modulator has been fully packaged with a 1 mm coaxial connector into a module to provide a 110 GHz operational bandwidth. The module offers RF insertion loss that was measured between 1 dB and 2.5 dB over most of the 70-110 GHz, with a 3 dB electrical bandwidth of 45 GHz and a 3 dB optical bandwidth of 90 GHz for a DC- $V_\pi$ of 8.5 V.

Overall, the modulator presented in this dissertation is the fastest LiNbO$_3$ modulator ever reported. The integrated module is also the fastest LiNbO$_3$ ever packaged.

### 7.2 Millimeter-Wave Imaging

The modulator presented in this dissertation has been integrated into two of the passive millimeter-wave imaging systems built so far at the University of Delaware in
W band. The modulator was first implemented into a single-pixel scanning imaging system at 94 GHz that was taken out for on-field testing (Fig. 7.1). In Fig. 7.1, we can see the millimeter-wave image at 94 GHz (c) taken by the imager on the cart (a) of a boat in the sea with passengers on board seating inside (b). The water reflecting the native blackbody radiations of the cold sky appears dark on the image, whereas blackbody radiations emitted around room temperature by the boat and the passengers appear lighter.

Figure 7.1: Sample imagery from passive millimeter-wave perception study of small watercraft.
The modulator was also integrated in a real-time, video rate passive millimeter-wave imaging system at 77 GHz [39]. The first version of the optical upconversion multi-chip module for distributed aperture detection is presented in Fig. 7.2 [40].

Figure 7.2: LiNbO$_3$ modulator integrated into Optical Upconversion Module.

In this module, millimeter-waves radiations are captured by a horn antenna (not represented) and coupled into CPW mode through a Waveguide-to-CPW transition [41]. The CPW mode carrying the millimeter-wave radiation is then guided to a commercially available 77 GHz PIN switch to provide a Dicke switch configuration with a known resistive load (typically 50 Ω), which is a well-established...
technique to reduce or eliminate the errors introduced by the normal fluctuations or
time-dependent variations in the gain of the receiver. The millimeter-wave signal is
then guided to a three-stage low noise amplification (LNA) section to improve the
sensitivity of the system. According to Friis formula, the overall noise figure of the
amplification is dominated by the first LNA. Therefore, an expensive high-end LNA
with low noise figure is used for the first stage, whereas cheaper LNAs are integrated
for the following amplification stages. Finally, the amplified millimeter-wave signal is
guided through a bandpass filter and then coupled to the modulator for optical
upconversion.

7.3 LiNbO₃ Modulator Future Improvements

New designs and fabrication techniques are being developed to further
improve the efficiency of LiNbO₃-based modulators. Thin-film LiNbO₃ (TFLN)
modulators are showing good promises and have the potential to provide very low $V_{π}$
and ultra broadband bandwidth [42]. In this design, a few microns thin layer of
LiNbO₃ prepared crystal ion slicing is bonded onto a RF-friendly carrier wafer with
low dielectric constant, such as quartz. The benefits from this technique is a dramatic
reduction of the dielectric loss, which is critical at high frequency, and the lowering of
the RF effective index due to the low dielectric constant of the carrier wafer, which
would therefore the etching of LiNbO₃ unnecessary. If effective index match and low
optical insertion loss and strong mode overlapping can be obtained, this technology
should be able in the future to outperform the modulator presented in this dissertation.
REFERENCES


Appendix A

COPYRIGHT TRANSFER AGREEMENTS

The paper “Development of Electro-Optic Phase Modulator for 94 GHz Imaging System” was published in the Journal of Lightwave Technology and is made available as an electronic reprint with the permission of OSA. The paper can be found at the following URL on the OSA website:
http://ieeexplore.ieee.org/stamp/stamp.jsp?tp=&arnumber=5325902. Systematic or multiple reproduction or distribution to multiple locations via electronic or other means is prohibited and is subject to penalties under law.

The paper “Full Spectrum Millimeter-Wave Modulation” was published in Optics Express and is made available as an electronic reprint with the permission of OSA. The paper can be found at the following URL on the OSA website:
http://www.opticsinfobase.org/oe/abstract.cfm?uri=oe-20-21-23623. Systematic or multiple reproduction or distribution to multiple locations via electronic or other means is prohibited and is subject to penalties under law.

The paper “Ultra-Broadband Modulator Packaging for Millimeter-Wave Applications” was published in IEEE Transactions on Microwave Theory and Techniques and is made available as an electronic reprint with the permission of IEEE. The paper can be found at the following URL on the IEEE website:
http://ieeexplore.ieee.org/xpl/articleDetails.jsp?arnumber=6702509. Systematic or multiple reproduction or distribution to multiple locations via electronic or other means is prohibited and is subject to penalties under law.