Study on electroabsorption modulators and grating couplers for optical interconnects

YONGBO TANG

Doctoral Thesis in Microelectronics and Applied Physics
Stockholm, Sweden 2010
Akademisk avhandling som med tillstånd av Kungliga Tekniska Högskolan framlägges till offentlig granskning för avläggande av teknologie doktorsexamen Måndag 27 September 2010 klockan 10.00 i sal 438, Forum, Kungliga Tekniska Högskolan, Isafjordsgatan 39, Kista, Stockholm.

© Yongbo Tang, August 2010

Tryck: Universitetservice US AB
Abstract

Decades of efforts have pushed the replacement of electrical interconnects by optical links to the interconnects between computers, racks and circuit boards. It may be expected that optical solutions will further be used for interchip and intra-chip interconnects with potential benefits in bandwidth, capacity, delay, power consumption and crosstalk. Silicon integration is emerging to be the best candidate nowadays due to not only the dominant status of silicon in microelectronics but also the great advantages brought to the photonic integrated circuits (PICs). Regarding the recent breakthroughs concerning active devices on silicon substrate, the question left is no longer the feasibility of the optical interconnects based on silicon but the competitiveness of the silicon device compared with other alternatives.

This thesis focuses on the study of two key components for the optical interconnects, both especially designed and fabricated for silicon platform. One is a high speed electroabsorption modulator (EAM), realized by transferring an InP-based segmented design to the hybrid silicon evanescent platform. The purpose here is to increase the speed of the silicon PICs to over 50 Gb/s or more. The other one is a high performance grating coupler, with the purpose to improve the optical interface between the silicon PICs and the outside fiber-based communication system.

An general approach based on the transmission line analysis has been developed to evaluate the modulation response of an EAM with a lumped, traveling-wave, segmented or capacitively-loaded configuration. A genetic algorithm is used to optimize its configuration. This method has been applied to the design of the EAMs on hybrid silicon evanescent platform. Based on the comparison of various electrode design, segmented configuration is adopted for the target of a bandwidth over 40 GHz with as low as possible voltage and high extinction ratio.

In addition to the common periodic analysis, the grating coupler is analyzed by the antenna theory assisted with an improved volume-current method, where the directionality of a grating coupler can be obtained analytically. In order to improve the performance of the grating coupler, a direct way is to address its shortcoming by e.g. increasing the coupling efficiency. For this reason, a nonuniform grating coupler with apodized grooves has been developed with a coupling efficiency of 64%, nearly a double of a standard one. Another way is to add more functionalities to the grating coupler. To do this, a polarization beam splitter (PBS) based on a bidirectional grating coupler has been proposed and experimentally demonstrated. An extinction ratio of around $-20$ dB, as well as a maximum coupling efficiency of over 50% for both polarizations, is achieved by such a PBS with a Bragg reflector underneath.

Keywords: Photonic integrated circuit, optical interconnect, transmission line, traveling-wave, electro-absorption modulator, hybrid silicon evanescent platform, grating coupler, vertical coupling scheme, optical antenna, lag effect and polarization beam splitter.
Acknowledgements

First and foremost, I want to thank Dr. Urban Westergren, my main supervisor in Sweden, for his support and advice during my study in KTH. I appreciate his hand-in-hand guidance on the characterization of the high speed device, valuable discussion on the modulator design and the help on the trivial paper work for the defense.

I would like to express my greatest gratitude to Prof. Sailing He, my supervisor in China, for his ongoing support, encouragement and help through my entire Ph.D study, for his coordination of the study in Sweden, and for his concern and suggestion on my life.

Grateful thanks to Dr. Lech Wosinski, one of my supervisors, for his full support on the research on grating couplers, the coordination of the visiting to UCSB and the help on the preparation of this thesis.

I would like to thank Prof. John Bowers in University of California Santa Barbara, for giving me the opportunity to do the exiting research on hybrid silicon modulators. I really enjoy the work in UCSB, with the nice atmosphere he and his group members have created.

Grateful thanks to Dr. Daoxin Dai for his continuing help and valuable advices to my study and my life. My Ph.D life must be much tougher without him.

I want to thank Prof. Lars Thylén, giving me the opportunity to be a member of FMI and the inspiring discuss of the research. I would also like to thank China Scholarship Council for partial financial support of abroad study.

Special thanks to Zhechao Wang, for the collaboration on the research of grating couplers. Special thanks to Hui-Wen Chen for her help on the mask design and the high speed characterization and for her hard work on the fabrication of the hybrid modulators. Thanks to Dr. Yichuan Yu, who helps me start the work on the modulator design.
I would like to thank the Photonics Research Group of Ghent University. Their open resources including the papers, dissertations, and codes shared on the website help me a lot.

I want to thank my colleagues in Electrum, Eva Andersson, Marek Chaciński, Prof. Min Qiu, Dr. Lena Wosinska, Dr. Min Yan, Dr. Richard Schatz, Dr. Jie Li, Dr. Qin Wang and many others for their kindness and help.

Thanks to my roommates in Sweden, Lingquan Deng first and Ke Wang then. Thanks to my friends in Kista mentioned above and below: Lin Dong, Dr. Mingyu Li, Ning Zhu, Dr. Jun Hu, Haiyan Qin, Pu Zhang, Tianhua Xu, Jin Lv, Zhangwei Yu, Geng Yang, Jiajia Chen, Weiquan Mei, Dr. Shan Qin, Dr. Zhili Lin, Xin Hu, Dr. Biao Chen, Qiang Li, Yi Song, Jing Wang, Dr. Jiaming Hao, Bin Tian, Wei Yan, Samul, Dr. Jie Tian, Yiting Chen, Zhenzhong Zhang, Dr. Zhibin Zhang, Jiantong Li, Zhiying Liu, Botao Shao, Ye Xu, Sha Tao, Yi Wang, Naeem Shahid, Xiaoyue Wu... I can't imagine the life in Sweden without them.

Thanks to my other colleagues and friends in UCSB: Dr. Zhi Wang, Dr. Di Liang, Dr. Gehong Zeng, Dr. Jason Tien, Dr. Martijn Heck, Jon Magnani, Christine Dillard-DeHerrera, Jon Peter, Jock Bovington, Jared Bauters, Anand Ramaswamy, Molly Piels, Géza Kurczveil, Sudharsanan Srinivasan, Andy Chang, Siddarth Jain, Michael Davenport, Shane Todd, Maaike van’t Westeinde, Yuji Zhao, Yuli Yan, Wenbo Xu, Fang Guo, Yichi Zhang, Zhisheng Li, Dr. Garay Menicucci, Dr. Min-feng Gu, Xiang Qiu, Haosheng Huang, Dr. Changwan Son, Yuanbo Mao... for the memorable living in Santa Barbara.

And, thanks to the folks in COER!

I would like to express my sincere appreciation to my parents, young brother, cousins and their families for their continuous support and care. And special thanks to dear Ying Gao for her love, encouragement and understanding.

Yongbo Tang
July 2010.
Contents

List of Papers ix

Acronyms xi

1 Introduction 1

1.1 Background .................................................. 1
1.2 Scope .......................................................... 3
1.3 Thesis Outline .................................................. 4

2 Hybrid Silicon Electroabsorption Modulator 5

2.1 Introduction .................................................. 5
2.1.1 Silicon-based Modulators .................................. 6
2.1.2 III/V-based Modulators .................................. 7
2.1.3 Hybrid Silicon Evanescent Modulators ................. 9
2.2 High Speed Electrode Design ................................. 10
2.2.1 Circuit Model ............................................. 10
2.2.2 Transmission Line Analysis and Optimization ........ 13
2.2.3 Discussion of Electrode Design ......................... 18
2.3 Segmented Hybrid EAM ...................................... 21
2.3.1 Structure Description ................................... 22
2.3.2 Device Design ........................................... 25

3 Silicon Grating Coupler 29

3.1 Introduction .................................................. 29
3.2 Analysis and Simulation ..................................... 31
3.2.1 Periodic Analysis ....................................... 31
3.2.2 Array Antenna Theory ................................... 34
3.2.3 Full-wave Simulation .................................... 38
3.3 Fabrication and Characterization ......................... 40
3.3.1 Process Flow ............................................. 40
3.3.2 Characterization ......................................... 43
### CONTENTS

3.4 Improving the Coupling Efficiency ........................................... 45  
3.4.1 Introduction ........................................................................... 45  
3.4.2 Directionality and Improved VCM Method ............................... 47  
3.4.3 Mode Matching and Nonuniform Design ................................. 51  
3.5 Coupling for Different Polarizations ......................................... 54  
3.5.1 Polarization Issues ................................................................. 54  
3.5.2 Polarization Beam Splitter Using a Bidirectional Grating Coupler ................................................................. 55

4 Summary, Conclusions and Future Work ...................................... 63

5 Summary of the Original Work ....................................................... 67

Bibliography ................................................................................. 71
List of Papers

List of papers included in the thesis:


List of papers not included in the thesis, but related:


## Acronyms

<table>
<thead>
<tr>
<th>Acronym</th>
<th>Definition</th>
</tr>
</thead>
<tbody>
<tr>
<td>1D, 2D, 3D</td>
<td>One-Dimensional, Two-Dimensional, Three-Dimensional</td>
</tr>
<tr>
<td>ASE</td>
<td>Amplified Spontaneous Emission</td>
</tr>
<tr>
<td>CLTWE</td>
<td>Capacitively-Loaded Traveling-Wave Electrode</td>
</tr>
<tr>
<td>CPW</td>
<td>CoPlanar Waveguide</td>
</tr>
<tr>
<td>DQPSK</td>
<td>Differential Quadrature Phase Shift Keying</td>
</tr>
<tr>
<td>DUV</td>
<td>Deep UltraViolet</td>
</tr>
<tr>
<td>DUT</td>
<td>Device Under Testing</td>
</tr>
<tr>
<td>EAM</td>
<td>ElectroAbsorption Modulator</td>
</tr>
<tr>
<td>EBL</td>
<td>Electron Beam Lithography</td>
</tr>
<tr>
<td>EME</td>
<td>EigenMode Expansion</td>
</tr>
<tr>
<td>EML</td>
<td>Electroabsorption Modulated Laser</td>
</tr>
<tr>
<td>E/O</td>
<td>Electro/Optical</td>
</tr>
<tr>
<td>EPIC</td>
<td>Electronic Photonic Integrated Circuit</td>
</tr>
<tr>
<td>FDTD</td>
<td>Finite Difference Time Domain</td>
</tr>
<tr>
<td>FEM</td>
<td>Finite Element Method</td>
</tr>
<tr>
<td>FOM</td>
<td>Figure Of Merit</td>
</tr>
<tr>
<td>FTTH</td>
<td>Fiber-To-The-Home</td>
</tr>
<tr>
<td>HFSS</td>
<td>High Frequency Structure Simulator</td>
</tr>
<tr>
<td>HSEP</td>
<td>Hybrid Silicon Evanescent Platform</td>
</tr>
<tr>
<td>Acronym</td>
<td>Description</td>
</tr>
<tr>
<td>---------</td>
<td>-------------</td>
</tr>
<tr>
<td>ICP-RIE</td>
<td>Inductively Coupled Plasma Reactive Ion Etching</td>
</tr>
<tr>
<td>MMI</td>
<td>MultiMode Interferometer</td>
</tr>
<tr>
<td>MoM</td>
<td>Method of Moments</td>
</tr>
<tr>
<td>MOS</td>
<td>Metal-Oxide-Semiconductor</td>
</tr>
<tr>
<td>MQW</td>
<td>Multiple Quantum Well</td>
</tr>
<tr>
<td>MZM</td>
<td>Mach-Zehnder Modulator</td>
</tr>
<tr>
<td>NRZ</td>
<td>Non-Return-to-Zero</td>
</tr>
<tr>
<td>PBC</td>
<td>Periodic Boundary Condition</td>
</tr>
<tr>
<td>PBS</td>
<td>Polarization Beam Splitter</td>
</tr>
<tr>
<td>PC</td>
<td>Polarization Controller</td>
</tr>
<tr>
<td>PDL</td>
<td>Polarization-Dependent Loss</td>
</tr>
<tr>
<td>PDM</td>
<td>Polarization-Division Multiplexing</td>
</tr>
<tr>
<td>PIC</td>
<td>Photonic Integrated Circuit</td>
</tr>
<tr>
<td>PhC</td>
<td>Photonic Crystal</td>
</tr>
<tr>
<td>PML</td>
<td>Perfectly Matched Layer</td>
</tr>
<tr>
<td>QCSE</td>
<td>Quantum Confined Stark Effect</td>
</tr>
<tr>
<td>SCH</td>
<td>Separate Confinement Heterostructure</td>
</tr>
<tr>
<td>SEM</td>
<td>Scanning Electron Microscope</td>
</tr>
<tr>
<td>SIMOX</td>
<td>Separation by IMplantation of OXygen</td>
</tr>
<tr>
<td>SOA</td>
<td>Semiconductor Optical Amplifier</td>
</tr>
<tr>
<td>SOI</td>
<td>Silicon-On-Insulator</td>
</tr>
<tr>
<td>TE, TM</td>
<td>Transverse Electric, Transverse Magnetic</td>
</tr>
<tr>
<td>TML</td>
<td>TransMission Line</td>
</tr>
<tr>
<td>TW</td>
<td>Traveling-Wave</td>
</tr>
<tr>
<td>VCM</td>
<td>Volume Current Method</td>
</tr>
</tbody>
</table>
Chapter 1

Introduction

1.1 Background

On Tuesday, October 6, the royal Swedish academy of sciences announced the 2009 Nobel prize in physics. Half of the prize was awarded to Charles K. Kao, the father of the fiber optics, "for groundbreaking achievements concerning the transmission of light in fibers for optical communication" [1]. The discovery of Prof. Kao actually sounded the horn of the evolution of using optical interconnects to replace the electrical interconnects.

Optical interconnects possess the advantages of wide bandwidth, huge data capacity, small interconnect delay, low power consumption, minimum crosstalk and high tolerance of electromagnetic interference over traditional metallic (Cu or Al) electrical interconnects [2, 3]. The success of the optical interconnects has completely revolutionized the telecommunication and promoted the prosperity of the Internet era. In the last several decades, the replacement of electrical interconnects by optical links has spread from the long-haul backbones to the metropolitan area networks and local area networks, and currently to the last mile to the customer with the speed-up progress of the fiber-to-the-home (FTTH) deployment. Similar evolution has also appeared in the field of the computer and microelectronic industry e.g. the optical ethernet adaptor for clusters and the light peak technology for personal computers [4]. It is no doubt that the optical technologies will further reshape the computer and microelectronic industry with the light entering the field of the chip-to-chip and core-to-core interconnections. The research on the inter-chip and intra-chip optical interconnects, driven by government, academia and industry, is working towards the seamless interfaces from the electronics to photonics and from the chip side to the outside world, and vice versa. The future chip would probably be a hybrid electronic photonic integrated circuit (EPIC) consisting of the electronic logic part and the optical transmission part.

However, optical technology has always suffered from arguments of economic concerns. Unlike the electronic integrated circuits, mainly based on one mate-
Chapter 1. Introduction

Material (Silicon) and one basic unit (transistor), optical interconnects need a lot of components including light source, photo detector, modulator, filter, waveguide, coupler and splitter etc. and those components, depending on the function and the structure, usually have their own favorite materials and individual optimized process methods. Such a condition of diversity spreads the limited devotion into various small directions and makes it hard for the optical society to develop the standardizations for the design, fabrication, testing and packaging, let alone the forward-looking roadmap. Even worse, additional efforts have to be paid on the techniques of e.g. coupling and polarization controlling to improve the compatibility for different platforms, challenged by the critical optical, thermal and mechanical requirements. Taking into account the specific material, the immature process, and the expensive assembly and packaging, it is clear why the optical interconnects are still only suitable for the cost-insensitive applications.

For the solution to reduce the cost, microelectronics has set a successful example in the past half century. The key is to achieve a high volume, large scale, high dense monolithic integration. This strategy can bring significant benefits at least in the following aspects: the expensive processing cost can be shared by every sample on the same wafer; the packaging cost can be greatly reduced when most interfaces between components are eliminated; and the system can be more complex but with reliable and stable performance. Those benefits may be more meaningful for photonics since it can save more in packaging (which takes a large fraction of the total expenditure) and the cost of high-accuracy controlling units used to keep the fragile optical system stable.

With the recent progress, silicon integration is emerging to be the best candidate nowadays for the photonic integrated circuits (PICs) and future EPICs [5]. Firstly, it is no doubt that we should choose silicon as the common substrate for the future EPICs, regarding the dominant status of silicon in the microelectronics. Secondly, silicon photonics can benefit a lot from the legacy of the well-developed microelectronic industry like the mature CMOS processing technologies, the large scale and high quality substrates, the existing production lines and foundries or even current industry infrastructure and business models. Most of them are actually more than sufficient for the development of PICs, even in the next five years. Thirdly, silicon is indeed an excellent material to guide light for integration due to the transparent attribute in the wavelength range for optical communication and the high index contrast between its oxide and itself. Silicon waveguide with propagation loss as low as 0.2 dB/cm and negligible bending loss even for a radius of 2 µm are available [6, 7]. Furthermore, the "short board" of the silicon for active functions has been greatly improved by recent breakthroughs on silicon-compatible techniques [8–10]. Among them, the hybrid silicon evanescent platform (HSEP), with the active III-V materials transplanted to the silicon-over-insulator (SOI) wafers by low-temperature wafer bonding, seems the most promising approach to "active" silicon so far with the success in the high performance lasers [11, 12], amplifiers [13], modulators [14, 15], tunable filters [16] and photodetectors [17].

The question left is no longer the feasibility for the optical interconnects based
1.2. Scope

on silicon PICs but the competitiveness of silicon device compared with its counterparts. Taking the modulator as an example, although the fastest silicon modulators based on carrier depletion can support a data rate of 40 Gb/s right now [18], the weak index modulation usually requires a long interaction length of mm-scale and very high driving voltage to achieve a sufficient extinction ratio. Resonator structure like the microring [19] and Fabry-Perot cavity [20], as well as the photonic crystal (PhC) structure [21] can significantly enhance the modulation efficiency but at the expense of a narrow optical bandwidth and a critical fabrication tolerance. Introducing other materials to overcome the drawback inherent to the pure silicon has revealed the potential for a better modulation performance but still need efforts in engineering [14, 15, 22–24]. Hence, there is still a long way to go for a commercially-available high performance silicon modulator.

Except the challenge from the silicon integration, special concern should be paid on the compatibility of the silicon interconnect to the outside fiber-based interconnect network. Although monolithic integration can reduce the troublesome chip-to-fiber coupling to the most, at least one chip-to-fiber interface is unavoidable once the chip needs interaction with outside world (mainly via the fiber link). However, it is nontrivial to get an efficient and cost-effective chip-to-fiber coupler, especially for the silicon circuits due to the large mode difference and the polarization mismatch between the high index-contrast silicon waveguide and the single mode fiber. Mode converters and polarization control schemes should be adopted to avoid a large coupling loss. Besides, such coupler should have a large alignment tolerance for cost-effective packaging and ease of mass-production.

1.2 Scope

In this thesis, part of the effort is devoted to developing a high speed electroabsorption modulator (EAM) based on hybrid silicon evanescent platform (HSEP). Through HSEP, the well-developed techniques for the InP substrate can be transferred to the silicon platform. The benefits are not only an ability of lasing but also an approach of efficient and fast modulation. Currently, the reported EAM based on HSEP has a bandwidth of around 16 GHz due to the limitation from the lumped design [14]. The bandwidth can be greatly improved if adopting traveling-wave electrode, not to speak of the advanced electrode design like the capacitively-loaded structure [15] or the segmented configuration[25]. We will focus on the segmented implementation for a bandwidth beyond 40 GHz.

The other part of this work is focusing on the grating coupler, an attractive vertical coupling scheme for the interface of the silicon PICs and the external fiber due to the advantages like the capability of on-wafer testing and potential low-cost packaging. Two problems will be addressed in order to improve the performance of the grating coupler. The coupling efficiency, typically, with a maximum of 20%-35% for a standard SOI grating coupler, is still not competitive for practical applications. Deeper understanding of the diffraction mechanics is required to guide the design
and advanced structures like the nonuniform grating [26] and the Bragg reflector can be attempted to improve the coupling efficiency with a moderate complexity of fabrication. Another way to improve the performance is to add more functionality to a grating coupler. The benefit from the new functionality can compensate the shortages to some extent. For instance, a bidirectional grating coupler simultaneously acting as a polarization beam splitter (PBS) [27] could be more preferable than a loose combination of a coupler and a PBS.

1.3 Thesis Outline

After the introduction of the thesis in Chapter 1, the study of the EAMs will be discussed in Chapter 2. The recent progress of silicon-based modulators and III/V-based modulators will be reviewed first. A robust small-signal model based on transmission line analysis will be presented. This method is comprehensive and can be applied to the EAMs with lumped, traveling-wave, capacitively-loaded or segmented configuration. The genetic algorithm is used for the optimization for a complex configuration. The pros and cons of the lumped, traveling-wave, capacitively-loaded and segmented configuration are carefully discussed. Finally, a design of a high speed hybrid EAM modulator is presented. Over 40 GHz bandwidth is expected from our simulation results.

In Chapter 3, theoretical analysis, standing on the viewpoint from the grating/photonic crystal and antenna theory, is depicted to offer an intuitive understanding of the principle of the grating coupler. Two numerical methods (i.e., the eigenmode expansion method and the finite difference time domain method) usually employed for the grating coupler simulation are introduced. We will then discuss the method to improve the coupling efficiency of the grating coupler. By taking the interface reflection into consideration, an improved volume current method is developed to analyze the grating coupler on an SOI wafer. Significant analytic formula is obtained to guide the grating coupler design, especially the choice of the initial structure parameters like the groove depth, waveguide thickness and the buffer layer thickness. Nonuniform design is experimentally demonstrated to improve the mode matching. We have also discussed the polarization issues related to the grating coupler. A novel PBS based on a bidirectional grating coupler is designed, fabricated and characterized.

Chapter 4 summarizes the research work included in this thesis and gives the prospect for the future work. At last, a brief summary of each published paper and the description of the author’s contributions are listed in Chapter 5.
Chapter 2

Hybrid Silicon Electroabsorption Modulator

In this chapter, the recent progress in both silicon-compatible modulators and III/V modulators will be reviewed and the hybrid silicon evanescent platform (HSEP), an promising approach to "active" silicon, will be introduced. Then we will present a comprehensive design and optimization method for the electrode design. It will be applied to the design of a segmented hybrid silicon electroabsorption modulator.

2.1 Introduction

Optical modulator is one of the key components for the optical interconnects. It converts the electrical data into the optical signal and mainly determines the data transmission speed of the optical link. The modulation of the light can be achieved by using many different physical effects, but in the final analysis, it can be concluded to be the change of the complex refractive index of the medium, i.e.

\[ \tilde{n} = n + j\kappa = n + j\frac{\lambda\alpha}{4\pi} \]

Here, \( n \) is the refractive index. \( \kappa \) is the extinction coefficient (proportional to the power absorption coefficient \( \alpha \) at certain wavelength \( \lambda \)). The real part and the imaginary part are not independent with each other but related by the Kramers-Kronig relations.

The type based on the control of \( n \) is referred to the refractive modulator. The phase modulation can be directly generated and it can be converted into the intensity modulation with the help of an interferometer e.g. a Mach-Zehnder structure or an resonator e.g. a microring. The type related to the manipulation of \( \kappa \) (or \( \alpha \)) is referred to the absorptive modulator. For example, the EAMs studied in this thesis are based on the field-dependent absorption, stemming from the Quantum Confined Stark Effect (QCSE) in a Multiple Quantum Well (MQW) structure.
Chapter 2. Hybrid Silicon Electroabsorption Modulator

For a high performance modulator design, the goals include: (a) wider bandwidth for higher transmission speed; (b) lower driving voltage for smaller power consumption and simpler driving circuits (without the amplifiers); (c) more compact design to raise the volume; (d) low cost for the market requirement. Typically, there are tradeoffs among the bandwidth, the extinction ratio (device length) and the driving voltage. A frequently-used figure of merit (FOM) for the optical modulators can be written as [28]:

$$FOM = \frac{2Z_{in}}{50 + Z_{in}} \frac{f_{3dBc}}{V_{pp}} \frac{\lambda}{1.55\mu m}$$  \hspace{1cm} (2.2)

where $V_{pp}$ is the swing of the driving voltage. For the absorptive modulator, $V_{pp}$ is chosen for a sufficient extinction ratio (ER), e.g. 10 dB; For the refractive modulator, it is chosen to achieve a phase shift of $\pi$ and usually denoted by $V_{\pi}$. $f_{3dBc}$ is the electro/optical (E/O) 3 dB bandwidth, corresponding to the frequency where the responsivity of the modulator ($\Delta P_0/\Delta I_e$) has dropped to its $1/\sqrt{2}$. $Z_{in}$ is the input impedance looking towards the modulator and it reflects the power consumption.

2.1.1 Silicon-based Modulators

It is well-known that silicon is not a good material to achieve fast optical modulation due to the lack of the linear electro-optic (Pockels) effect and the weak Franz-Keldysh effect. So far, gigascale modulator based on pure silicon is only achieved by using the free carrier dispersion effect related to the carrier concentration. The variation of the carrier concentration, achieved by using either the carrier injection (e.g. in a forward biased PIN diode) or the carrier depletion (e.g. in a reverse biased PN diode), can bring a small change of the refractive index on the order of $10^{-4} \sim 10^{-3}$ at the expense of certain loss.

Table. 2.1 summarizes the recent progress in pure silicon modulators. Only the one that can support over 10 Gb/s operation are included. Carrier injection was first demonstrated to achieve fast silicon modulators. Two papers about this topic published in Nature [19, 40] greatly encouraged the researchers in this field and attracted more people to devote into the study of silicon modulator. However, carrier injection into a PN junction is slow, typically more than 1 ns. Even with the pre-emphasis driving circuit, the fastest reported transmission rate is only 12.5 Gb/s [32]. The injection into a MOS capacitor is a bit faster but less efficient. The big capacitance limits its high speed performance[29]. Projection MOS design shows a significant improvement with fine structure optimization [30].

The carrier depletion is fast and no meaningful bandwidth limitation has been found. However, suffering from the tradeoff between the index change and the effective depletion area, the modulation efficiency is small, even less than the case of the carrier injection. Although the fastest silicon modulator based on this effect has reached a bandwidth of over 30 GHz, it needs very long interaction length and pretty high voltage to make the eye diagram open. Even so, the extinction ratio of 1 dB at 40 Gb/s in [29] is still very low compared with the commercial modulators.
Tab. 2.1. Reported pure silicon modulators for over 10 Gb/s applications. For the same design from one group, only the best results are listed. MZI: Mach-Zehnder interferometer; MOS: metal oxide semiconductor; TW: traveling-wave; $E_{R_{\text{dyn}}}$: dynamic extinction ratio.

| Group                  | Structure | $f_{3\text{dB}_e}$ | Len | $V_e \cdot L$ | $E_{R_{\text{dyn}}} | V_{\text{pp}} | \text{Gb/s}$ | Note                  |
|------------------------|-----------|---------------------|-----|---------------|------------------|-----------------------|
| carrier injection      | Basak, Intel, 2008 [29] | MZI|MOS | 10 | 3450 | 33 | 3.8 | 1.1 | 10 | cascaded elec. |
|                        | Fujikata, NEC, 2010 [30] | MZI|MOS | 25 | 120 | 5 | 3 | 3.5 | 12.5 | projection MOS |
|                        | Green, IBM, 2006 [31] | MZI|pin | 200 | 0.36 | ? | 1.2 | 10 | pre-emphasis, 7 V |
|                        | Xu, Cornell, 2007 [32] | Ring|pin | 10 | 9 | 8 | 12.5 | 16 V |
| carrier depletion      | Basak, Intel, 2008 [29] | MZI|pn | 33 | 1000 | 40 | 1.1 | 6.2 | 40 | TW, 14 Ω |
|                        | You, ETRI, 2008 [33] | Ring|pn | 8 | 100 | 1.2 | 4 | 12.5 |
|                        | Park, ETRI, 2009 [34] | MZI|pn | 7 | 1500 | 18 | 3 | 4 | 12.5 |
|                        | Feng, Kotura, 2010 [35] | MZI|pn | 12 | 1000 | 14 | 7 | 8 | 12.5 |
|                        | Dong, Kotura, 2009 [36] | Ring|pn | 11 | 30 | 6.5 | 2 | 10 |
|                        | Zheng, Sun, 2010 [37] | Ring|pn | 15 | 30 | 3.1 | 2 | 5 | drive-integrated |
|                        | Marris, PSUD, 2008 [38] | MZI|pip | 10 | 4000 | 50 | 3.1 | 2 | 5 | drive-integrated |

Based on III/V material or LiNiO3. The extinction ratio can be improved in a microring modulator in spite of a weak modulation efficiency [35]. But the speed will now be strictly limited by the optical bandwidth of the microring.

In order to overcome the disadvantage inherent to the silicon material, some other materials good for fast modulation are introduced to the silicon platform. Germanium (Ge) is another group IV element and can be mixed with silicon over the complete range of compositions. The discovery of the enhanced Franz-Keldysh effect in the bulk tensile-strained Ge/Si [22] and the strong QCSE effect in the strained Ge/GeSi MQW [41] laid the foundation for the Ge/Si electroabsorption modulators although there is still room to improve the performance of the reported devices [22, 23]. Another potential candidate is to use a nonlinear polymer mixture with a high electro-optic coefficient. Embedding the polymer into the silicon slot or cladding the polymer over silicon waveguide has been studied to enable effective modulation for silicon waveguide [24, 42]. Those devices are very promising but obviously still in their very early stage. Table 2.2 lists the performance of the reported Ge/Si modulators and polymer-based modulators.

2.1.2 III/V-based Modulators

III/V-based modulators have been studied for several decades and the commercial products can be found in the form of an individual device or a part of an electroabsorption modulated laser (EML). The mechanics behind an EAM is the
field-dependent absorption, stemming from the Franz-Keldysh effect in bulk material or quantum confined Stark effect (QCSE) in a MQW structure. In both cases, the bandgap can be manipulated by changing the applied voltage to make the absorption spectrum shift towards the longer wavelength. Thanks to the exciton resonance, QCSE effect will lead to a sharper absorption edge and a more drastic absorption shift. Those mean a better extinction ratio and a smaller driving voltage and make MQW preferable for the EAM design. Mach-Zehnder modulator (MZM) can also be achieved based on the effective refractive index change in the III/V MQW region as a result of the QCSE absorption, carrier density change, and band filling. The operation wavelength for the MZM should be chosen far away from the strong absorption region in order to keep a low loss.

The recent progress in III/V-based EAMs suitable for over 40 Gb/s transmission are summarized in Table 2.3. Only the publications after the year 2004 are taken into account. The summary for the earlier publications can be found in [63, 64].

Thanks to the presence of QCSE in the III/V MQW, III/V material renders much better performance for a fast modulation compared with silicon. Decades of efforts have promoted the performance of III/V-based EAMs to a very high level. The FOMs in most cases have gone beyond 20. The bandwidth of the EAM with segmented electrode has approached 100 GHz [58, 59] and the field trial at 112 Gb/s has been successfully carried out [65]. It seems that the main obstacle of the exploration for a higher modulation speed is no longer the design of the optical modulator itself but the lack of the high speed electrical driving circuits and corresponding testing instruments. Many researchers have turned to pursue some other goals, e.g. a sufficient extinction ratio with a driving voltage close to 1 V or even less so that it can be compatible with the output of the CMOS circuit without any amplifier. The choice of the MQW is very important to release the tradeoff between the extinction ratio and the driving voltage. Nowadays, InGaAlAs MQW is becoming more and more popular, due to the benefits from a larger band offset [44]. Another trend is to integrate more functional devices on one chip such like the combination of the semiconductor optical amplifier with EAM [52, 56] and the electroabsorption modulated laser (EML) [45, 47, 53, 59], which is believed to be the most successful PIC product so far. Although not listed in Table 2.3, developing an EAM-based advanced modulation format transmitter is also attractive for the ability to support multiple transmission rates with a limited electrical bandwidth.

### Tab. 2.2. Selected promising silicon-compatible modulators.

<table>
<thead>
<tr>
<th>Group</th>
<th>Material</th>
<th>$f_{3dB}$</th>
<th>Len</th>
<th>$V_x \cdot L$</th>
<th>$E_{R_{dyn}}$</th>
<th>Note</th>
</tr>
</thead>
<tbody>
<tr>
<td>Liu,MIT,2008[22]</td>
<td>GeSi</td>
<td>1.2</td>
<td>50</td>
<td>-</td>
<td>-</td>
<td>EAM</td>
</tr>
<tr>
<td>Rong,Stanford,2010[23]</td>
<td>GeSi</td>
<td>13</td>
<td>30</td>
<td>0.53</td>
<td>2.5</td>
<td>3.125</td>
</tr>
<tr>
<td>Baehr-Jones,UW,2008[24]</td>
<td>polymer</td>
<td>1kHz</td>
<td>2000</td>
<td>5</td>
<td>-</td>
<td></td>
</tr>
</tbody>
</table>
An compact DQPSK modulator was realized by embedding two EAMs into a three-arm interferometer, and 107 Gb/s transmission rate has been demonstrated for that device [66]. Mach-Zehnder modulator is more straightforward to achieve an advanced modulation format transmitter. Table 2.3 also lists the progress of III/V-based MZMs.

### 2.1.3 Hybrid Silicon Evanescent Modulators

The advantages of III/V materials in active function, e.g., the state-of-the-art modulator shown above, is exactly what the silicon platform is lacking. Many researchers are looking for a way to import III/V materials to the silicon platform. But because of the large mismatch in lattice constant and thermal expansion coefficient, it is very difficult to grow III/V materials directly on silicon. An indirect way is to use the wafer bonding to transplant the III/V material to the silicon substrate. A technique called "O$_2$ plasma-assisted low temperature wafer bonding" has been developed for this purpose and hybrid wafers consisting of high quality
Chapter 2. Hybrid Silicon Electroabsorption Modulator

III/V materials and SOI substrates can be obtained even for a wafer diameter as large as 150 mm [8]. Fig. 2.1 shows the bonding process flow.

![Flow diagram showing the bonding process flow](image)

**Fig. 2.1.** Process flow of the O$_2$ plasma-assisted low temperature wafer bonding. VC: vertical channel for outgassing

Based on the hybrid wafer, silicon evanescent waveguide is developed for practical optical functionality. In this structure, the optical mode is confined partly in the silicon waveguide and partly in the active core (typically the MQW region). This integrated PIC configuration is called *hybrid silicon evanescent platform* (HSEP). It has been proven to be feasible and potentially cost-effective with the success in high performance lasers [11, 12], amplifiers [13], modulators [14, 15], tunable filters [16] and photodetectors [17].

Table 2.4 summarizes the performance of the reported hybrid modulators. Compared with the other silicon modulators, significant progress in bandwidth and extinction ratio can be seen in those prototypes. The performance can be further improved with advanced electrode design as shown in this thesis.

| Group                  | $f_{3dBc}$ [GHz] | $L$ [µm] | $V_n \cdot L$ [V·mm] | $E/R_{dyn}$ [dB|V$_{pp}$|Gb/s] | Electr. load [Ω] | Note     |
|------------------------|------------------|-----------|----------------------|---------------------------------|------------------|----------|
| Chen, UCSB, 2008[67]   | 8                | 500       | 2                    | 6.3|1.5|10                                 | TW, 25            | MZM      |
| Kuo, UCSB, 2008[14]    | 16               | 100       | -                    | 6|3.2|10                                 | Lumped            | EAM      |

2.2 High Speed Electrode Design

2.2.1 Circuit Model

Modeling an EAM usually involves the propagations of both the microwave and the lightwave, which are coupled with each other. For an EAM, controllable optical absorption is the basis of the optical modulation. The absorption is strongly dependent on the electric field inside the MQW region and therefore related by the
2.2. High Speed Electrode Design

Fig. 2.2. Equivalent circuit presentation of an EAM cell with a length of $\Delta l$. Dashed box models the interaction of microwave and the lightwave; Dotted box shows the equivalent circuit for the transmission line, usually employed in a small signal model for an EAM.

Voltage over the intrinsic layer. As a result of the optical absorption, the photocurrent will be generated and fed back into the driving circuit. It will in turn affect the microwave propagation to certain extent. Besides, coming with the change of the absorption, there will be also a change of the refractive index described by the Kramers-Kronig relations. That means a frequency chirp, which should be carefully treated in the optical communication system.

An effective way describing the aforementioned phenomena is an equivalent circuit model as shown in Fig. 2.2. The dotted box presents a quasi-static circuit model for the transmission line (the electrode), where $R_c$ is the resistance of the electrode, proportional to $\sqrt{f}$ at high frequency range due to the skin effect, $L_m$ is the inductance, $R_{sm}$ is the series resistance mainly contributed by the contact layers, $C_{im}$ is the intrinsic capacitance and $C_e$ is the parasitic capacitance. The interaction between the microwave and the lightwave is included in the dashed box, where corresponding parameters are given as follows:

\[
\tau = \frac{\Delta l}{c}(n_o - \Delta n) \quad (2.3a)
\]

\[
T = e^{-\Gamma\alpha(V_i, P_{in})\Delta l} \quad (2.3b)
\]

\[
I_p = R_\lambda(P_1 - P_2) = R_\lambda P_1(1 - T); \quad (2.3c)
\]

where $\tau$ is the time delay for the light traveling along the short section, $\Delta n$ is used to take into account the chirp effect, $n_o$ is the group index of the light and $c$ is the light speed in vacuum. $T$ is the transmission function of the light power, $\alpha(V_i, P_{in})$ denotes the relation among the absorption, the voltage and the optical power, typically obtained by measurement and $\Gamma$ is the confinement factor. $I_p$ is the photocurrent directly related to the absorbed power and $R_\lambda$ is the responsivity.
Δl is the modulator section’s length and should be kept small enough for a high accuracy.

Generally, the whole cascaded circuit can be input to a circuit simulator (e.g. Pspice, LTspice) to get both the optical and electrical response in frequency domain or time domain. The premise of such a large signal model is a full description of α(Vi, Pin), which needs a lot of efforts in measurement. The measurement-based model would be accurate for the device under investigation and useful for the same type, but for different structures new measurement is usually required.

For the purpose of performance evaluation during the design period, more meaningful way is to use a small-signal model, where only the linear item of Vi in the complex expression α(Vi, Pin) is taken into consideration and the optical response is directly related to the integral of the voltage the lightwave experiences during the propagation [68]. It is found that the increase of the photocurrent will actually flatten the frequency response and increases the bandwidth to some extent [68]. Hence, the feedback of the photocurrent can be ignored for a conservative evaluation. Then the EAM model can be simplified to a microwave problem of the transmission line analysis.

According to the circuit parameters shown in Fig. 2.2, the characteristic impedance and the propagation constant of an EAM can be given by

\[ Z_C(\omega) = \sqrt{\frac{Z_S(\omega)}{Y_p(\omega)}} \]  
\[ \gamma(\omega) = \alpha(\omega) + j\beta(\omega) = \sqrt{Z_S(\omega)Y_p(\omega)} \]

where

\[ Z_S(\omega) = R_C + j\omega L_m \] \hspace{1cm} (2.5a)
\[ Y_p(\omega) = \frac{j\omega C_{im}}{1 + j\omega C_{im}R_{sm}} + j\omega C_e \] \hspace{1cm} (2.5b)

It is useful to further approximated them as [68]

\[ Z_C \approx \sqrt{\frac{L_m}{C_{im} + C_e}} \] \hspace{1cm} (2.6a)
\[ \alpha \approx \frac{R_C}{2Z_C} + \frac{\omega^2}{2}C_{im}R_{sm}Z_C \] \hspace{1cm} (2.6b)
\[ \beta \approx \omega \sqrt{L_m(C_{im} + C_e)} \] \hspace{1cm} (2.6c)

Note that the dominant factor for the microwave loss is the second item, indicating that the intrinsic capacitance will play an important role, especially in the high frequency range.
2.2.2 Transmission Line Analysis and Optimization

When the electrical frequency goes up to over 40 GHz, the microwave wavelength falls into the millimeter scale comparable with the practical electrode length. Common circuit analysis based on lumped elements is no longer suitable for such case, instead, transmission line is introduced to tackle the issues of microwave propagation and reflection. Full-wave simulation based on Maxwell equations is a more general method but usually not convenient for the analysis and design.

One useful tool for the transmission line analysis is the transmission matrix, or called ABCD matrix. It is defined in terms of the total voltages and currents (see Fig. 2.3(a)) [69]:

\[
\begin{bmatrix}
V_1 \\
I_1
\end{bmatrix} =
\begin{bmatrix}
A & B \\
C & D
\end{bmatrix}
\begin{bmatrix}
V_2 \\
I_2
\end{bmatrix}
\] (2.7)

Fig. 2.3. (a). Definition of the transmission matrix; (b) The ABCD parameters of three basic circuits

Fig. 2.3(b) lists the transmission matrices of three basic two-port networks. They can be used to derive the equivalent ABCD matrix for the complex circuits.

Combining with the definition of the input impedance looking towards the load,
i.e., $Z_{in} = V/I$, we can get the following useful equations:

\[
V_2 = \frac{V_1 Z_{in,2}}{AZ_{in,2} + B}
\]

\[
Z_{in,1} = \frac{AZ_{in,2} + B}{CZ_{in,2} + D}
\]

Fig. 2.4. Various electrode designs. Only the signal electrode is shown for clarity. (a) basic lumped design; (b) basic traveling-wave design; (c) revised lumped design; (d) impedance-controlled electrode; (e) capacitively-loaded traveling-wave design; (f) segmented design.

Fig. 2.5. Cascaded two-port network representation of the electrode of an optical modulator.

Fig. 2.4 lists various electrode designs employed in the literature. Those electrodes include lumped, traveling-wave, capacitively-loaded and segmented types.
2.2. High Speed Electrode Design

can be essentially classified as a TML connection based on coplanar waveguide (CPW) or microstrip structure. All of them can be treated as a cascaded connection of several two-port network blocks as shown in Fig. 2.5. Each block can be described by a transmission matrix. Starting from the load, the input impedance looking towards the load at each node in Fig. 2.5 can be calculated from the left to the right by Equ. 2.8b. Then $V_1 = \frac{Z_{in,1}}{Z_s + Z_{in,1}} V_S$ according to Thevenin’s theorem. Using Equ. 2.8a, the voltage at the other nodes can be calculated one by one from the right to the left.

![Fig. 2.6. Sketch of a modulator segment. Dotted box shows equivalent quasi-static circuit model for the transmission line of an EAM](image)

Fig. 2.6 schematically shows a transmission line for a uniform modulator segment with a length of $l$, which can be seen as one box in Fig. 2.5. The voltage at position $z$ can then be given by

$$V(z) = \frac{V_2(e^{-\gamma z} - e^{\gamma z}) - V_1(e^{-\gamma (z-l)} - e^{\gamma (z-l)})}{e^{-\gamma l} - e^{\gamma l}}. \quad (2.9)$$

There will be a voltage drop over the series resistance $R_{sm}$ due to the doped cladding layers. Hence, the voltage over the intrinsic layer ($V_i$ in Fig. 2.6), which determines the field strength inside the MQW region and controls the absorption, is obtained by

$$V_i(z) = \frac{1}{1 + j\omega R_{sm} C_{im}} V(z) \quad (2.10)$$

Taking into account the walk-off effect due to the velocity mismatch, the actual $V_i$ the lightwave packet experiences should be written in the time dependent form as:

$$V_i(z, t = t_0 + z/v_o) = \frac{1}{1 + j\omega R_{sm} C_{im}} V(z)e^{j\omega t} \quad (2.11)$$

Here $t_0$ is the time when the lightwave packet reaches the position at $z = 0$. Then the small-signal response of the modulation contributed by this segment can be
approximated to be [68]

\[ I_{ac}(\omega) \approx \int_0^l V(z, t_0 + z/v_o)dz \]

\[ = \frac{1}{1 + j\omega R_{sm}C_{im}} \frac{V_1 e^{j\omega t_0}}{1 + \Gamma_r e^{-2\gamma l}} \left\{ \frac{e^{(j\beta_o - \gamma)l} - 1}{j\beta_o - \gamma} + \Gamma_r e^{-2\gamma l} \frac{e^{(j\beta_o + \gamma)l} - 1}{j\beta_o + \gamma} \right\} \]

(2.12a)

where

\[ \Gamma_r = \frac{V_2 e^{\gamma l} - V_1}{V_1 - V_2 e^{-\gamma l}} \]  

(2.12b)

Here, \( \beta_o = \omega n_o/c \) (\( n_o \) is the group refractive index of the lightwave, \( c \) is the light speed in vacuum) and \( \Gamma_r \) is the reflection coefficient at \( z = l \). If there are several modulation segments, the total modulation response is their sum and the output of the microwave power from the photodetector is the sum’s square. The modulation response is usually normalized to the ideal case with lossless microwave propagation, perfect impedance match and velocity match. It is easy to get that the voltage along the electrode in the ideal case is equal to a half of the driving voltage \( V_S \). Hence the final normalized E/O response can be given by:

\[ M(\omega) = \left| \frac{\sum I_{ac}(\omega)}{V_S} \right|^2 \]

(2.13)

This TML analysis is very general and comprehensive. The effects due to the impedance mismatch, microwave loss and velocity mismatch are all included. For a TW electrode with one modulation segment, Eqn. 2.13 will be reduced to the one in [68]; for a periodic segmented electrode, it will be reduced to the one derived by the Bloch-wave approach [25]. Fig. 2.7 shows the E/O response and reflection for the EAMs reported in [25]. The results calculated by this model agrees well with the experiment results. Besides, this model is not limited to the EAMs but also suitable for the III/V-based MZM design.

Fig. 2.7. Calculated E/O response and \( S_{11} \) for the EAM reported in [25]

The transmission matrix, the characteristic impedance and the complex propagation constant used in this model can be easily obtained by a TML simulator (e.g.
2.2. High Speed Electrode Design

HFSS) or converted from the measured scattering matrix of a straight transmission line as follows [69]:

\[
\begin{bmatrix}
A & B \\
C & D
\end{bmatrix} = \begin{bmatrix}
\frac{(1+S_{11})(1-S_{22})+S_{12}S_{21}}{2S_{21}} & \frac{Z_{ref}(1+S_{11})(1+S_{22})-S_{12}S_{21}}{2S_{21}} \\
\frac{1}{Z_{ref}} & \frac{(1-S_{11})(1-S_{22})-S_{12}S_{21}}{2S_{21}}
\end{bmatrix}
\]

(2.14a)

\[2 \cosh \gamma l = \frac{1}{S_{21}} + S_{12} - \frac{S_{11}S_{22}}{S_{21}}\]  

(2.14b)

\[Z_C = Z_{ref} \sqrt{\frac{(1+S_{11})(1+S_{22})-S_{12}S_{21}}{(1-S_{11})(1-S_{22})-S_{12}S_{21}}}\]  

(2.14c)

where \(Z_{ref}\) is the reference impedance and usually equal to 50 Ω for a standard instrument.

Thanks to the analytical formalism, the model developed above could quickly calculate the modulation response. However, if there are several modulation segments, the design is still very complex since too many parameters need to be optimized. For example, there will be at least six input parameters for a two-segmented EAM device (Fig. 2.4(f)). It is not easy to optimize those structures with traditional optimization methods. Here we use a genetic algorithm [70], a robust global optimization method and famous for the optimization with multiple parameters.

To apply the genetic algorithm, we need to construct an objective function, encode the chromosome as its input parameters and choose appropriate search ranges of the structural parameters. The 3 dB bandwidth (\(f_{3dB}\)) is the most important figure for the electrode design and can be directly derived from Equ. 2.13. It can be set as the objective function to be maximized. The lengths of the segments (including modulation segment (\(L_m\)), passive TML segment (\(L_p\)) and passive optical segment (\(L_o\))) and other interesting parameters such as the capacitances of the pads in Paper B are discretized in a binary-coded form as the input chromosome in the genetic algorithm. The critical parameter can have a fine resolution and the less important one can have a rough discretization so that the optimization can be fast.

In fact, a frequency response having a good bandwidth does not mean a better performance in practical transmission. A response with a drop in low-frequency range close to \(-3\, \text{dB}\) will still meet the definition of the bandwidth but is obviously not an expected design. Such case will occur especially when some capacitive component is introduced into the model. Additional limitation should be applied to get a good design. A simple empirical rule is adopted in our optimization. As schematically illustrated in Fig. 2.8, \(f_{1dB}\) is required to be larger than \(0.4 \, f_{3dB}\), while \(f_{2dB}\) should be greater than \(0.8 \, f_{3dB}\). Here, \(f_{1dB}\), \(f_{2dB}\) and \(f_{3dB}\) denote 1 dB, 2 dB and 3 dB bandwidths, respectively. Such a filtering mechanism could be easily implemented in the genetic algorithm by setting a smaller selection probability for an undesirable structure. More meaningful filter rules can be explored, such as keeping the fluctuation of the group delay less than 5 ps for a 50 Gb/s NRZ transmission.
Fig. 2.8. A typical frequency response of a segmented EAM with capacitive pads. \( f_{1dB}, f_{2dB} \) and \( f_{3dB} \) denote 1 dB, 2 dB and 3 dB bandwidths, respectively. As a limitation, \( f_{1dB} \) is set greater than 0.4 \( f_{3dB} \), and \( f_{2dB} \) is set greater than 0.8 \( f_{3dB} \).

### 2.2.3 Discussion of Electrode Design

To be simple, we first consider an EAM with only one active segment. Referring to Fig. 2.5, a lumped electrode in Fig. 2.2(a) is equivalent to the TW case with an open terminator (the center-driven lumped device can be seen as a case with two paralleled open transmission lines), i.e., \( \Gamma_r = 1 \) in Equ. 2.12a. Ignoring the microwave loss and applying \( e^{\gamma l} \approx 1 + \gamma l \) for a short active length, Equ. 2.13 can be simplified to [68]:

\[
M(\omega) \approx \left| \frac{2}{1 + j\omega Z_S C_m l + j\omega R_{sm} C_m} \right|^2
\]  

(2.15)

It implies that the bandwidth limit for a lumped device is related to the total RC time constant, where the item related to the source impedance is usually dominant. One way to increase the bandwidth is to add a shunt resistance \( R_{sh} \) at the feed-in point (e.g. Fig.2.2(c)). Then the E/O response of this revised lumped design can be given by:

\[
M(\omega) \approx \left| \frac{2/(1 + \frac{Z_S}{R_{sh}})}{(1 + j\omega Z_S C_m l/(1 + \frac{Z_S}{R_{sh}}) + j\omega R_{sm} C_m)} \right|^2
\]  

(2.16)
2.2. High Speed Electrode Design

Clearly, the RC limit is improved, equivalent to the case with a reduced source impedance and hence leading to a better bandwidth. However, the numerator revised from 2 to \(2/(1 + \frac{Z_S}{R_{sh}})\) in this expression indicates that the modulation efficiency is indeed reduced, meaning a penalty on the extinction ratio.

In order to improve the bandwidth, traveling-wave (TW) electrode (Fig. 2.2(b)) is proposed. Strictly speaking, traveling-wave describes a case where the wave travels unidirectionally without any reflection. That means \(\Gamma_r = 0\), \(Z_L = Z_C\), \(V_1 = V_S/(1 + Z_S/Z_L)\) and gives the normalized E/O response in the form of

\[
M(\omega) = \left| \frac{2/(1 + \frac{Z_S}{Z_L})}{(1 + j\omega R_{sm}C_{im})} \frac{e^{(j\beta_o - \gamma)l} - 1}{(j\beta_o - \gamma)l} \right|^2 
\]

(2.17a)

\[
\approx \left| \frac{2/(1 + \frac{Z_S}{Z_L})}{(1 + j\omega R_{sm}C_{im})} \right|^2 
\]

(2.17b)

Here, we can see that RC limit is further reduced to the internal RC penalty given by \(R_{sm}C_{im}\). The item related to the source impedance in Equ. 2.15 is totally eliminated in this case. This elimination of the dominant item means at least a double of the bandwidth compared with the basic lumped design. However, we should point out that this improvement is indeed based on the cost of modulation efficiency similar to the revised lumped design. A response drop of as high as 9.5 dB in low frequency range will occur for a typical TW configuration, with a characteristic impedance of 25 Ω or even lower. This degradation of the modulation response comes from two aspects. About 3.5 dB is caused by the drop of the feed-in drive voltage at the input port due to the impedance mismatch. In the TW case, only the forward microwave appears on the modulation part; while in the lumped design, the constructive overlap of both the forward and the backward microwave doubles the voltage over the modulation part. This difference leads to another 6 dB drop.

Someone may argue that the backward reflection, even if it is constructive, is harmful for the modulator’s performance due to the so called walk-off effect, which describes the expansion of the modulated pulse induced by the velocity mismatch between the microwave and the lightwave in time domain. This walk-off can be evaluated by analyzing the term \(\left| \frac{e^{(j\beta_o - \gamma)l} - 1}{(j\beta_o - \gamma)l} \right|^2\). If only considering the velocity mismatch, we get

\[
\left| \frac{e^{(j\beta_o - \gamma)l} - 1}{(j\beta_o - \gamma)l} \right|^2 \approx \left| \text{sinc} \left( \frac{\omega l}{2c} (n_o - n_e) \right) \right|^2 
\]

(2.18)

It will give a 3 dB bandwidth of

\[
f_{3dB} = \frac{1.39c}{(n_e - n_o)l} = \frac{0.417}{\Delta n l} \text{ [GHz]} 
\]

(2.19)

When the lightwave and the microwave propagate towards the same direction (corresponding to the forward microwave), a typical \(\Delta n\) would be less than 4 and
the bandwidth will be beyond 200 GHz for an active length \( l < 0.5 \text{ mm} \). For the counter-propagation case (corresponding to the backward microwave), the \( \Delta n \) can be as big as 12 and the walk-off is much more serious. But if the active length is shorter enough, e.g. \( l < 0.15 \text{ mm} \), the 3 dB bandwidth can be still pushed beyond 200 GHz. Therefore, the backward reflection should be avoided for a long device; but for a short device, the walk-off can be tolerable and it would be worthy to make constructive reflections happen to enhance the modulation efficiency. This enhancement would be more valuable in the high frequency range since it can compensate the response drop due to e.g. the RC limit and thus extends the bandwidth.

Fig. 2.2(d) presents an improved electrode design, called *impedance-controlled electrode* (ICE) in [71]. This technique was initially developed to achieve traveling-wave operation in a standard 50 \( \Omega \) RF system. Two passive TMLs with high characteristic impedances are introduced. One connects the source and the modulation segment’s input port; the other one connects the load and the modulation segment’s output port. With this configuration, the input impedance at the load side and that at the source side can be controlled by changing the lengths of the two passive TMLs. Good impedance match can be obtained at the source side and the troublesome microwave reflection to the source can be eliminated to some extent. The input impedance at the output port can be tuned to be the same as the characteristic impedance of the modulation segment for TW operation or moderately larger than that to get the constructive reflection. The latter condition has a good performance no matter on bandwidth or extinction ratio but works well only for a short modulation length.

Segmented electrode design shown in Fig. 2.2(f) can be seen as a variation of the ICE. The change is that here one modulation segment is divided into several small segments. This setup can keep the advantage of the ICE design in impedance control. Actually, it goes further with increased design freedom. The short segment enables a better use of the reflection and extends the bandwidth, and at the same time the total long active length can insure a good extinction ratio. It implies that segmented design can relax the tradeoff between the bandwidth and the extinction ratio, compared with the lumped and the TW designs. The disadvantage of the segmented design is that it needs redundant passive optical waveguides to connect the modulation segments. They will bring additional optical loss. A good choice is to use them to be the semiconductor optical amplifiers [56].

Similar segmented operation on the revised lumped design leads to the capacitive-loaded design in Fig. 2.2(e). With a small loaded segment, the RC limit will not be a problem. This electrode is frequently used for the MZM device with a very long modulation length. In those long devices, the traveling-wave feed line is preferable but the bonus from the reflection would no longer exist.
2.3 Segmented Hybrid EAM

As mentioned in Sec. 2.1.3, Kuo et al. have developed a hybrid silicon evanescent EAM with a lumped electrode (see Fig. 2.4(a)) [14]. The sketch of the cross section, the top view photo and the detailed epitaxial design can be found in Fig. 2.9.

For the hybrid EAM, the typical values for $R_{sm}$ and $C_{im}$ are $3 \Omega \cdot \text{mm}$ and $1 \text{pF/mm}$, respectively [14]. With these parameters, Equ. 2.15 predicts a 20 GHz bandwidth when $l = 100 \mu m$. It agrees well with the measurement. This bandwidth is still low for the mainstream transmission speed of 40 Gb/s. Higher bandwidth would be expected to reduce the power consumption per bit and simplify the setup of the optical interconnect.

We have shown that the basic lumped setup might be the worst choice for a higher bandwidth. Simply adjusting the electrode configuration can significantly improve the EAM’s bandwidth and reduce the microwave reflection, proven by many other studies [28, 63, 64].

In this thesis, we explore the segmented electrode designs to increase the bandwidth of the hybrid EAM. The sketch of a two-segmented hybrid EAM is shown in Fig. 2.10. The main process flow is depicted in Fig. 2.11. Our priority target for the first attempt is a bandwidth of 40 GHz, which can support at least a 50 Gb/s NRZ transmission rate. Under this premise, a longer modulation length is
Fig. 2.10. Sketch of structure of the two-segmented hybrid EAM. (a), (b), (c) show the schematic cross section of the modulation segment, the passive hybrid waveguide and the passive transmission line, respectively.

preferable for a low driving voltage and a high extinction ratio.

2.3.1 Structure Description

According to the difference in cross section, a segmented hybrid EAM can be basically divided into five components: modulation segment, passive microstrip, passive hybrid waveguide, passive silicon waveguide and taper section. The cross sections of the former three components are depicted in the bottom of Fig. 2.10. Among them, the modulation segment is certainly the key part. Its structure is similar to the previous one in [14]. We continue to use the old III/V epitaxy, with detailed parameters shown in the table of Fig. 2.9. The InGaAlAs/InGaAlAs MQW consists of 10 wells and 11 barriers with the photoluminescent (PL) peak at 1478 nm. According to the bonding flow shown in Fig. 2.1, this III/V epitaxial wafer is bonded over an SOI wafer including a 500 nm thickness silicon layer and a 1 μm silica buffer layer. The silicon rib waveguide is patterned on the wafer in advance of the bonding. The rib height is fixed to 250 nm but the rib width is variable with 1.5 μm for passive Si waveguide and 1.0 μm under III/V mesa. After the bonding, the mesa on top of the active region is first defined with a width of 4 μm by using
a CH$_4$/H$_2$/Ar-based plasma reactive ion etch. The width of the MQW layer, as well as the SCH layers, is then selectively under-etched down to around 2 $\mu$m. The under-etching is operated by using wet-etching recipe (H$_3$PO$_4$/H$_2$O$_2$), with circle pattern as the reference to indicate the etch speed. This waveguide configuration can release the intrinsic RC limit to some extent.
After the mesa definition, 0.5µm Ni/Au/Ge/Ni/Au alloy is deposited over the thin n-InP contact and forms the ground metal after lift-off process. The electrical isolation of the modulation segment and the passive hybrid waveguide is then realized by proton implantation. Microstrip is employed for the passive transmission line for a compact design. SU8 polymer is used for the dielectric layer, with its thickness increased to 4.75 µm for a high characteristic impedance. The top signal strip has a width of 4 µm and a height of 3 µm. The lengths of both the passive microstrips and the active modulation segments are carefully chosen to maximize the bandwidth and reduce the reflection to the source. GSG contact pads with 100 µm pitches are adopted at the input and the output ports for the RF testing.

![Fig. 2.12.](image)

(a). Optical mode distribution of the hybrid waveguide. (b). Effective refractive index v.s. wavelength

Fig. 2.12(a) shows the calculated optical fundamental mode distribution in the modulation segment. The mode power is mainly confined in the MQW/SCH region, with an evanescent tail in the silicon waveguide. That is different with the configuration for the hybrid evanescent laser since a high confinement factor is expected for an EAM to get a good extinction ratio. The confinement factor in the wells for this structure is around 19%. Further reducing of the silicon rib width can increase the confinement factor but will face the challenge from the process. The group index of the waveguide can be obtained by fitting the data of the wavelength-dependent effective indices. As shown in Fig. 2.12(b), the group index is about 3.5.

The transition from the passive silicon waveguide to the hybrid modulation part is achieved by a 60 µm bi-level taper structure, with III/V mesa’s width linearly tapered from 0.6 µm to 2.0 µm and the silicon rib width from 1.5 µm to 1.0 µm, and vice versa.

In this version, we did not do any special treatment to the passive hybrid waveguide. This hybrid part including the active III/V materials can be designed to be an SOA to compensate the EAM loss or to be replaced with a passive silicon waveguide and two tapers. Those adjustments can be tried in future.
2.3.2 Device Design

Tab. 2.5. Simulation parameters

<table>
<thead>
<tr>
<th>Layer</th>
<th>Thickness [µm]</th>
<th>Width [µm]</th>
<th>ε_r</th>
<th>n</th>
<th>σ [S/m]</th>
<th>Note</th>
</tr>
</thead>
<tbody>
<tr>
<td>p-contact</td>
<td>0.1</td>
<td>4</td>
<td>13.91</td>
<td>3.42</td>
<td>20</td>
<td>including contact resistance</td>
</tr>
<tr>
<td>p-cladding</td>
<td>1.5</td>
<td>4</td>
<td>12.5</td>
<td>3.1563</td>
<td>1520</td>
<td>doping-dependent mobility</td>
</tr>
<tr>
<td>p-SCH</td>
<td>0.15</td>
<td>2</td>
<td>13.39</td>
<td>3.431</td>
<td>0</td>
<td>depletion under bias</td>
</tr>
<tr>
<td>MQW</td>
<td>0.187</td>
<td>2</td>
<td>13.52</td>
<td>3.491</td>
<td>0</td>
<td></td>
</tr>
<tr>
<td>n-SCH</td>
<td>0.1</td>
<td>2</td>
<td>13.39</td>
<td>3.431</td>
<td>32500</td>
<td></td>
</tr>
<tr>
<td>n-contact</td>
<td>0.15</td>
<td>2</td>
<td>12.5</td>
<td>3.1681</td>
<td>110400</td>
<td>doping-dependent mobility</td>
</tr>
<tr>
<td>Metal</td>
<td>3</td>
<td>4</td>
<td>1</td>
<td>4.1e7</td>
<td>-</td>
<td>strip in the passive TML</td>
</tr>
<tr>
<td></td>
<td></td>
<td>3</td>
<td>10</td>
<td></td>
<td></td>
<td>strip in the modulation part</td>
</tr>
<tr>
<td>Si</td>
<td>0.5</td>
<td></td>
<td>11.9</td>
<td>3.46</td>
<td>0.1</td>
<td></td>
</tr>
<tr>
<td>SiO2</td>
<td>1</td>
<td></td>
<td>3.9</td>
<td>1.46</td>
<td>0</td>
<td></td>
</tr>
<tr>
<td>SU8</td>
<td>4.75</td>
<td></td>
<td>*1</td>
<td>1.7</td>
<td>*2</td>
<td></td>
</tr>
</tbody>
</table>

*1) \(2.23834 + \frac{1.04166}{(1 + 7.56216 \times 10^{-22} \times f^2)}\)

*2) \(\frac{0.00652385 + 0.0005359 \times 10^{-21} \times f^2}{1 + 7.56216 \times 10^{-22} \times f^2}\)

The EAM design is based on the transmission line model described in Sec. 2.2.2. The key is to get the characteristic impedances and complex propagation constants for the modulation segment and the passive segment. Those parameters can be calculated by a FEM-based electromagnetic simulator HFSS, with suitable input parameters. Table 2.5 summarizes the material parameters and the primary structure parameters used in our modeling. The conductivity of the p-contact is intentionally set to be much lower than the actual value in order to take into account the contact resistance. The calculation of the conductivity of the InP material has included the doping-dependent mobility. Although the p-SCH layer is slightly doped, it would be depleted under reverse bias. Hence, we set its conductivity to be zero. In contrast, the n-SCH layer is designed to be an insulation layer. But we believe that the heavy doping operation to the n-contact will inevitably make it at least partly doped and become conductive. A conductivity of 32500 S/m is assumed.

Fig. 2.13(a), (b) shows the calculated electrical parameters for the modulation segment. The characteristic impedance of our modulation segment is about 21 Ω, which is very typical for an EAM. The microwave group index is around 5. It is a bit larger than the optical group index 3.5. This small difference will not bring too much side-effect to an EAM according to Equ 2.19. The microwave loss in the modulation segment is a bit high due to the pretty big intrinsic capacitance (see Equ. 2.6b). A higher characteristic impedance of the passive microstrip is expected for the ease of the impedance match at the source side and to create the constructive reflection for the modulation segment. We tuned it to be around 80 Ω (see Fig. 2.13(c)) by using a 4.75 µm thick SU8 layer and a relative narrow strip.
with a 4 µm width. The optimization of the impedance is restricted by current process. The microwave group index is around 1.6 and the microwave loss is pretty low, usually smaller than 1 dB/mm in experiment for the frequency < 30 GHz. Those two issues allow us to use a long passive microstrip for the design.

Another concern is paid on the GSG pads. It will probably induce some parasitic capacitance. To minimize it, the ground metal under the signal pad is removed. During the design, we set a capacitance of 15 fF for each GSG pad in our model to cover its influence.

The remaining structure parameters are the lengths of each components. They are firstly optimized by applying the genetic algorithm as presented in Sec. 2.2.2 and then slightly adjusted for the layout reason.

We have put several segmented designs into our mask file. Those designs can be categorized into two types, the one-segmented structure (i.e., the EAM with an impedance-controlled electrode) and the two-segmented structure. They are labeled as S1Mxxx and S2Mxxx, respectively. Here, xxx presents the total modulation length.

Fig. 2.14 shows the calculated results of the electrical reflection ($S_{11}$), electrical transmission ($S_{21}$), the E/O modulation response (M) and the optical group delay.
2.3. Segmented Hybrid EAM

Fig. 2.14. Simulation evaluation of the designs. (a), reflection to the source; (b), electrical transmission; (c), E/O modulation intensity response; (d), group delay through the modulator.

after the modulator for each design. All devices are assumed to be connected with a standard 50 Ω source and a 50 Ω terminator. Over 40 GHz bandwidth is predicted by the plotting for the proposed design. Regarding the margins in the plotting, it should be safe to reach the target of 40 GHz, even suffering from some deviation due to the process.
Chapter 3

Silicon Grating Coupler

In this chapter, we will first focus on a standard grating coupler to discuss the basic principle using periodic analysis and antenna theory. The design, fabrication and characterization will then be briefly introduced. In order to improve the coupling efficiency, we experimentally demonstrated a nonuniform grating coupler by utilizing the lag effect in the dry etching process. We also proposed a polarization beam splitter based on a bidirectional grating coupler, as well as an improved design with a bottom Bragg reflector underneath.

3.1 Introduction

The idea to use a grating to couple light in or out of a slab waveguide goes back to the 1970s [72]. Interesting examples like the blazed grating coupler [73], the focusing grating coupler [74] and the nonuniform grating coupler [74, 75] have appeared for some decades although most of them were demonstrated only in theory due to the difficulty in sub-wavelength fabrication. The grating coupler becomes more and more valuable [76–78] when the silicon-on-insulator material system emerges to be the most popular PIC platform, together with the development of the high-resolution fabrication techniques e.g. the electron beam lithography (EBL) and the deep ultraviolet (DUV) lithography.

For silicon PICs, the enhanced ability of light confinement owning to the large refractive index contrast between the silicon core and the silica buffer can lead to a compact design and increased volume, on the other hand, it will also bring the challenge of the coupling between the silicon waveguide and the external fiber because of the large mode mismatch and the polarization-dependent loss. Direct butt-coupling will lead to an unacceptable loss of as high as 20 dB [79]. Various spot-size mode converters like the tapered fiber or the micro-lens from the fiber side and the lateral two-dimensional (2D) taper from the chip side are proposed to reduce the coupling loss. Less than 1 dB loss using lateral coupling including an polymer-cladded inverted taper and a tapered fiber has been demonstrated by some
groups \[80, 81\], although the critical alignment tolerance and the cumbersome facet polishing make it still far away from automatic packaging and mass fabrication.

Fig. 3.1. Setup for a grating coupler. No index-matching gel is shown for clarity

Grating coupler provides an alternative way to solve the coupling problem. Through a grating coupler, the light is coupled out of the wafer plane rather than from the lateral edge, hence it is called *vertical coupling scheme*. As the structure shown in Fig. 3.1 for an output coupler, the small waveguide mode is first expanded to a wide slab mode by an adiabatic lateral taper and then through a waveguide grating, it is diffracted into the top layer (usually an index-matching gel) with a tilt angle to the vertical direction, at the same time, the mode will be expanded in the longitudinal dimension to be compatible with the fiber mode and can be easily collected by a standard single mode fiber without any other special treatment. The input operation from the fiber to the waveguide is reciprocal to the output coupling. Such a vertical coupling scheme, moving the Input/Output (I/O) port from the wafer edge to the wafer surface, can bring a lot of benefits. The freedom to put the I/O port anywhere on the wafer will give great convenience to the mask layout and increase the wafer surface usage rate. With the grating coupler being an optical contact pad and the fiber being an optical probe, the PICs can be characterized in a similar way as the standard electronic wafer-scale testing. It indicates that a united testing platform for all the EIC, PIC and EPIC chips is possible. Furthermore,
the elimination of the operations like the wafer cleaving and facet polishing can significantly simplify the procedure of the testing and packaging, and the relaxed alignment tolerance can enable multiple I/O packaging without precise control. Those may lead to a potential cheap packaging solution [82].

3.2 Analysis and Simulation

3.2.1 Periodic Analysis

Grating Theory

For a basic understanding of the grating coupler, the simple ray model can provide an intuitive insight into the principle. Considering an output grating coupler, the light propagating along the grating waveguide will be diffracted upwards or downwards because of the effective index modulation by the etch grooves. As schematically shown in Fig. 3.2, the upward beams diffracted from adjacent grooves will experience a light path difference of

\[ \Delta = n_1 AC - n_{\text{eff}} \Lambda, \]  

where \( n_{\text{eff}} \) is the effective refractive index of the grating waveguide, \( \Lambda \) is the grating period. \( \theta \) in Fig. 3.2 is the radiation angle of the diffracted light with respect to the vertical direction. The principle of interference indicates that the radiation will be enhanced by constructive interference when the light path difference is an integer multiple of the wavelength \( \lambda \), i.e., under the in phase condition:

\[ (n_1 \sin \theta - n_{\text{eff}}) \Lambda = m\lambda, \quad m = 0, 1, 2, ... \]
If we rewrite it in the form of the wave vector, the well-known Bragg condition can be obtained as:

\[ \beta = k_1 \sin \theta - mk_g, \]  

(3.3)

where, \( \beta \) is the propagation constant of the guided mode, \( k_1 \) is the wave-vector of the diffracted light in the covering medium and \( k_g = \frac{2\pi}{\Lambda} \) is the grating vector. Clearly, \( \beta \) should be larger than \( k_1 \). Hence, only negative value of \( m \) can meet the Bragg condition. Typically, \( m = -1 \) will be adopted in general design.

![Wave-vector diagram](image)

**Fig. 3.3.** Wave-vector diagram for an output grating coupler at the grating interface with (a) a small positive output \((\theta > 0)\); (b) a small negative output \((\theta < 0)\); (c) a perfectly vertical output \((\theta = 0)\); (d) a large positive output where the second-order diffraction presents

The Bragg condition can be further described in a graphic representation, i.e., the wave-vector diagram. It will be very straightforward to get the possible diffraction order as well as the radiation angle from the wave-vector diagram. As shown in Fig. 3.3(a), when the grating vector is smaller than the propagation constant, the first order diffraction will be directed towards right. This radiation angle is positive, with the same sign as the projection of the diffraction in the input light’s direction. On the contrary, if the projection is opposite to the input light (e.g. the case of Fig. 3.3(b)), the angle is negative.

There will probably exist the second-order diffraction or even higher orders. For the case with a small grating vector in Fig. 3.3(c), the first-order diffraction has a
positive radiation angle but the second-order diffraction leads to a negative angle. Obviously, one fiber can’t collect both of them. Hence, higher-order diffraction should be eliminated for an efficient coupling. In this sense, negative radiation angle is preferable since the higher-order diffraction will be out of the radiation range, but it may have practical alignment problem. Another special case is the one shown in Fig. 3.3(d) for a perfectly vertical coupling when $\theta = 0$. It is attractive due to the simplification in coupling setup and assembly. However, the reflection due to the second-order diffraction in this case is inevitable. The perfectly vertical coupling can only be possible when the grating vector equals the propagation constant. The forward guiding mode and the backward guiding mode will meet the Bragg condition at the same time as depicted in Fig. 3.3(d). It means that we can’t expect a high vertical coupling efficiency without additional structure to suppress the backward reflection.

In the above analysis, we do not show the downward light. By replacing $n_1$ with $n_3$, Eqn. 3.2 can be also applied to the downward light. For a grating coupler, the presence of the downward light means a coupling loss since only the upward light can be possibly collected by the fiber positioned over the grating. The way to reduce the downward light will be discussed in Section 3.4.2.

### Photonic Crystal

The grating theory can reveal the relation between the radiation angle and the grating period for a grating coupler. The use of the Equ. 3.2 relies on the effective refractive index of the grating waveguide. But it is not easy to get that parameter. One simple way is to use the average of the effective indices of the etched part and the blank part. However, such a rough approximation can only be useful as an initiate value in practical design. For a two dimensional grating coupler $[76, 83, 84]$, it is out of the scope of the grating concept. The term photonic crystal coupler will be more suitable to name such devices.

Since a grating is a special photonic crystal structure, the general tools e.g. the band analysis (Bloch-wave analysis) widely used for the photonic crystal analysis can also work on the grating coupler, no matter it is based on a one dimensional (1D) grating or a two dimensional (2D) hole array. A unit of a 1D grating coupler for band analysis is shown in Fig. 3.4(a). Periodic boundary condition (PBC) and perfectly matched layer (PML) are applied to the vertical and horizontal boundaries, respectively. This unit is defined as a grating cell in this work. The cell width (corresponding to the grating period $\Lambda$) is labeled by $a$, following the practice of photonic crystal study. The width and the depth of the etch groove are marked by $w$ and $d$, respectively. Fig. 3.4(b) depicts the band diagram of a grating cell for TE polarization (with electric field parallel to the grating cell). Essentially, the grating coupler is an application of the leaky mode above the light line. In order to meet the phase matching condition, the normalized wave vector of the leaky mode along the propagation direction should be the same as the leaky wave in the covering medium. For the light radiating with a tilt angle $\theta$ in the covering medium, the
normalized wave vector of the leaky mode can be given by

\[ k = n_1 \sin \theta \cdot \frac{a}{\lambda} = n_1 \sin \theta \cdot \omega \]  

(3.4)

where, \( \omega \) is the normalized frequency. The troublesome \( n_{\text{eff}} \) in Equ. 3.2 is disappeared here. It is actually implicitly included in the band diagram. We can further conclude that \( \omega/k \) is determined by the radiation angle. It implies that there is a leaky line through the original point and the \((\omega, k)\) locus in the band diagram. Any mode located on this line will have an equal radiation angle related to the slope. Furthermore, the power flux propagation direction is determined by the sign of the group velocity \((v_g = d\omega/dk)\) of the leaky mode. Hence, for the intersection points A and B, they will have the same radiation angle but opposite propagation directions. That is the basis for a duplexer with three ports as demonstrated in [85], where a duplex based on a bidirectional grating coupler is reported to separate the 1520 nm and 1310 nm channels.

### 3.2.2 Array Antenna Theory

The above method, no matter the traditional grating theory or the more general photonic crystal tool, only see the high level of the periodic structure but being out of focus on the detailed structure inside one period. The use of the periodic...
boundary condition built on an assumption of infinite periods may magnify the specialty for a practical device with a finite length.

For a thorough understanding of the grating coupler deep into the physical mechanics of the radiation, especially considering those influence from the sub-period structure like the shape of the grating groove and the slab structure, we resort to the well-developed antenna theory assisted by the volume current method (VCM). Unlike the full-wave simulation, where the physical principle is hidden behind the result, an analytical solution obtained through the approximate VCM method can illuminate the diffraction process in essence and give a clear and intuitive physical picture. The behavior of radiation reveals that a grating coupler can be recognized as an optical phased array antenna. The concepts from well-developed antenna theory may offer new viewpoint on the design and the application of the grating coupler. For instance, the behavior of the beam steering using a grating coupler in [86] is obvious from the perspective of the array antenna [87].

The basic idea of VCM is to induce a virtual current to model the relative dielectric constant difference $\Delta \epsilon$ in certain region, e.g., the groove in a grating coupler. The current density $\vec{J}$ is defined by

$$\vec{J} = j\omega \Delta \epsilon \vec{E},$$

(3.5)

where $\omega$ is the field angular frequency and $\vec{E}$ is the electric field. The wave equation in the form of magnetic vector potential can then be given by [88]

$$(\nabla^2 + \omega^2 \mu_0 \epsilon) \vec{A} = \mu_0 \vec{J},$$

(3.6)

where $\mu_0$ is the vacuum permeability and $\epsilon$ is the material’s dielectric constant.

For the case of the grating coupler, the grating waveguide (see Fig. 3.5(a)) in a background of refractive index $n_1$ is modeled as a uniform waveguide with virtual current blocks replacing the grooves (Fig. 3.5(b)). The radiation basically scattering at certain discontinuity can be seen as the contribution of those current sources. To be simple, we start with the theory developed in [89] for the case of weakly confining grating, where the index contrast of the waveguide and the background is very small and we can assume the radiation is inside a uniform background ($n_1$). The whole structure can be hence simplified to be the phase array antenna system as shown in Fig. 3.5(c). In the far-field region, where the angular distribution of the radiation is independent on distance, the solution for a small tilt radiation angle to the vertical ($|\cos \theta| \approx 1$) takes the form of [89],

$$E_x(\vec{r}) \propto A_0 \sqrt{\frac{2}{\pi kr}} e^{-j\pi/4} e^{jz_0 \phi} \cdot F_D \cdot F_W \cdot F_A,$$

(3.7)
Chapter 3. Silicon Grating Coupler

Fig. 3.5. (a) Grating waveguide structure. (b) VCM approximation. (c) Array Antenna model for the analysis of a grating coupler.

where \( k = k_1 = n_1 k_0 \), \( \kappa = \sqrt{n_2 k_0^2 - \beta^2} \), \( \phi = \beta - k \sin \theta \),

\[
A_0 = \frac{\mu_0 \omega (n_2^2 - n_1^2) E_0}{4 \cos (\kappa t/2)}
\]

\[
F_D = \int_{t/2 - d}^{t/2} \cos (\kappa y) e^{-j k y \cos \theta} dy \approx de^{-j \frac{(t-d) k \cos \theta}{2}} \frac{\sin (dk \cos \theta/2)}{dk \cos \theta/2}
\]

\[
F_W = \int_0^w e^{j z \phi} dz = we^{j w \phi/2} \frac{\sin (w \phi/2)}{w \phi/2}
\]

\[
F_A = \sum_{n=0}^{N-1} (e^{j \Lambda \phi})^n = e^{j \frac{N-1}{2} \Lambda \phi} \frac{\sin (N \Lambda \phi/2)}{\sin (\Lambda \phi/2)}
\]
and \( \beta \) is the propagation constant of the mode in the grating waveguide and the structure parameters are defined in Fig. 3.5(a). The total radiation power can be computed from the Poynting vector:

\[
\vec{S} = \frac{\omega_0^2 A_0^2}{\mu_0 r} |F_D|^2 |F_W|^2 |F_A|^2 \hat{r}
\]  

(3.9)

**Fig. 3.6.** Power radiation patterns contributed by the depth element factor, width element factor, array factor and the total product for a grating coupler with 20 grooves. \( \theta = 15^\circ \), \( w = 200 \text{ nm} \), \( d = 90 \text{ nm} \), \( p = 629 \text{ nm} \) and \( t = 260 \text{ nm} \). Left, in Cartesian coordinate; right, in Polar coordinate.

Here, \( F_A \) is called *array factor*, the most important parameter in the common antenna literature [87]. The impact stemming from the groove shape is included in the parameters of \( F_D \) and \( F_W \), defined as *width element factor* and *depth element factor* in this thesis, respectively. Their product is the *element factor*. Using the formula above, the total power radiation in dB is plotted in the left of Fig. 3.6. The contributions from different factors are also depicted to give a rough evaluation of individual influence. We can see that among the three factors, the array factor plays the dominant role in the radiation behavior and the element factors acting like an angular filter sculpting the response.

More commonly, the power radiation pattern is typically plotted in the form of polar diagram as shown in the right of Fig. 3.6. Compared with the wave-vector diagram, the pattern diagram is much more useful and of course more vivid. It indicates not only the possible radiation angle but also the relative radiation strength. More detailed information like the side lobes due to the limited groove number can also be found in the vicinity of the main lobe.
3.2.3 Full-wave Simulation

Electromagnetic field behavior including the propagation, scattering, interference and diffraction can be simply but accurately described in the mathematical form of only a set of partial equations—i.e., the Maxwell equations, named after the founder James Clerk Maxwell. However, it is not trivial to get an analytical solution of those equations, even for some simple structure. A lot of numerical techniques like FDTD, FEM, MoM etc. are developed to solve the electromagnetic problem [90, 91]. Depending on the algorithms (based on e.g. grid discretization, basis function expansion, or statistical trial) and the practical problem (e.g. geometry complexity, solution scale), each method has its own pros and cons and can find its best application field [90].

Two rigorous methods are employed for the simulation of the grating coupler. One is the EigenMode Expansion (EME) method, mainly for the design and optimization; the other one is the well-known Finite Difference Time Domain (FDTD) method, mainly for the final evaluation of the entire device.

EME

The foundation of the EME method is the basis function expansion and mode matching technique. The electromagnetic field for a uniform waveguide in the propagation direction can be represented in terms of a series sum of local modes, i.e [92]:

\[
E(x, y, z) = \sum_{k=1}^{M} (a_k e^{i\beta_k z} + b_k e^{-j\beta_k z}) E_k(x, y), \quad \text{electricfield} \tag{3.10a}
\]

\[
H(x, y, z) = \sum_{k=1}^{M} (a_k e^{i\beta_k z} - b_k e^{-j\beta_k z}) H_k(x, y), \quad \text{magneticfield} \tag{3.10b}
\]

where \(\beta_k\) is the propagation constant of the waveguide mode (either guided mode or radiation mode), \(a_k\) and \(b_k\) presents the amplitudes for the backward and forward modes, respectively. This expansion can be seen as an exact solution of the Maxwell equations for a linear medium if the adopted mode number is big enough. It is very easy to set one slab source or more in the input waveguide by setting corresponding mode amplitude to be 1 and the others to be 0. At the abrupt joint in geometry or index distribution, the continuity boundary condition for the tangential components and the energy flow can be applied to calculate the reflection and transmission corresponding to each mode, which sew the field in each separate uniform section into a union.

Combined with the periodic boundary condition (PBC), the EME method can be applied to the periodic structure (e.g., grating, photonic crystal) and calculate its Blochmode’s propagation constant and related band diagram. This function has been embedded in the software of CAMFR [93], a two dimensional EME implementation. It is useful for the grating coupler design since we can further get
two important design parameters i.e. the effective refractive index and the leakage factor of the grating waveguide.

The EME method is pretty efficient in memory usage and calculation speed due to the $N_0$ scaling in propagation direction \[92\]. The inherent bidirectional property and wide-angle capability make it capable for many complex problems include diffractive elements, directional couplers, MMIs, bend, periodic structures and others. It is competitive in dealing with very fine structure like the nonuniform grating coupler in Paper G, although it may increase the computational time dramatically.

**FDTD**

Since the most interesting feature for a grating coupler is the coupling efficiency at a designated wavelength, a fast frequency domain solver like the EME method is preferable for the design and optimization fixing at one wavelength point. Once the design is finished, it is also necessary to evaluate the bandwidth and validate the design before fabrication by some other method, for that purpose we resort to the FDTD method.

![Fig. 3.7. Yee’s lattice in Cartesian coordinate (left) and leapfrog scheme in time domain (right)](image)

FDTD method might be the most popular simulation method in the field of integrated optics, due to the well understanding, easy implementation, universal capability and intuitive rendering of the light evolution in both space and time domain. The essence of the algorithm is the finite difference approximation of the Maxwell equations. According to the Yee’s lattice \[94\] as depicted in Fig. 3.7, the space discretization for the FDTD separates the electric field and the magnetic field into two interleaving grid networks with a half step offset. To ensure accuracy, but keeping a modest computational cost, the grid size must be moderately small,
typical within $\lambda/10 \sim \lambda/20$. The time domain usually adopts a *leapfrog scheme* (see Fig. 3.7) where the electric field and magnetic field are alternatively updated by each other with a maximum limit of the time step $\Delta t$ according to the *Courant stability condition*.

$$\Delta t \leq \frac{1}{v_{\text{max}} \sqrt{\frac{1}{\Delta x^2} + \frac{1}{\Delta y^2} + \frac{1}{\Delta z^2}}}$$

(3.11)

where $\Delta x$, $\Delta y$, $\Delta z$ are the lattice space steps, respectively, and $v_{\text{max}}$ is the maximum light phase velocity within the calculation region.

Although FDTD supports continuous wave (CW) operation and can get specific information at certain wavelength with revised algorithm combining with Z transforms [95], it is not efficient in time consumption and memory usage compared with the frequency domain simulator, e.g. EME. However, it is the dominant method in the time domain. It can vividly render the optical phenomena in animation and give an intuitive understanding of the field evolution. By setting a wideband pulse source as input, and assisted by the Fourier transform algorithms (FFT, DFT et al.), FDTD can be very efficient to output the wideband response after only one running.

### 3.3 Fabrication and Characterization

#### 3.3.1 Process Flow

Fig. 3.8 depicts the process flow for the fabrication of a standard grating coupler in our lab. Due to the difference in the etch depth, the definition of the waveguide ($1 \sim 5$) and the definition of the grating ($6 \sim 9$) are operated separately. Each definition requires one pattern generation and one dry etching.

The silicon-on-insulator wafer (1) can be usually obtained from commercial market. SOI wafers with high crystal quality can be made based on the Separation by IMplantation of OXygen (SIMOX) or wafer bonding (e.g. Smart Cut method) and presents pretty good optical, mechanical and thermal performance. A SIMOX SOI wafer is used to fabricate the grating coupler demonstrated in Paper G, with the purpose to eliminate the impact from the waveguide loss. However, such commercial wafer is usually in-batch for sale, with fixed structure parameters, which limit the freedom of design. As an alternative way, the amorphous SOI wafer can offer more flexibility in the developing period since the layer thickness of either the silicon or the silica can be accurately controlled during the depositions and it can be home-made by using PECVD in an ordinary clean room, although it typically suffers a higher waveguide loss.

The feature size of both the waveguide and the grating is in the scale of sub-wavelength, which is out of the capability of the regular photolithography technique. Those grating couplers in this thesis are all defined by using electron beam lithography (EBL) system (Raith 150). In the EBL, the use of electrons rather than the photons can push the diffraction limit from the micro scale to the nano scale.
With appropriate configuration, linewidths on the order of 10 nm or even smaller can be achieved by a dedicated EBL system. Due to its high resolution and flexibility, EBL emerges to be the dominant research tool in the field of nanophotonic although it can not support high volume production due to the low throughout. After exploring the prototype, the practical fabrication can be transferred to more general tools in industry like the DUV lithography.

In order to make full use of the EBL, high quality resist and suitable recipe are required. In Fig. 3.8, we showed the use of two resists: the negative resist HSQ for the waveguide and the positive resist ZEP520 for the grating. Here, the positive resist means that the part exposed to the electron beam will be removed by the resist developer and will be the window for the sequential etching ions; on the contrary, the part exposed to the beam for the negative resist will be kept after the developer and will be a shield to protect the material underneath during the etching. The choice of the resists usually depends on their performance like the resolution, contrast and sensitivity. Although the waveguide can be defined by using a positive resist, we prefer the negative HSQ for a better side-wall roughness. For the EBL system, due to the pixel by pixel scanning, we should also pay attention to the exposed area which is related to the total exposure time. Taking the grating as an

Fig. 3.8. Process flow for a standard grating coupler. 1. The original SOI wafer; 2. After the spin of HSQ; 3. After the first EBL exposure; 4. HSQ resist pattern; 5. Waveguide after the first etching; 6. After the spin of ZEP520; 7. After the second EBL exposure; 8. ZEP520 resist pattern; 9. Grating coupler after the second etching
example, it is unacceptable to use a negative resist even though it may provide a better performance. Once the pattern is generated on the resist, no matter the resist is positive or negative, the following etching process is the same. We use inductively coupled plasma reactive ion etching (ICP-RIE), an anisotropic dry etching method, which can achieve nearly vertical sidewalls and high aspect ratio. The etchant, high energy ions with a speed driven by an electrical field, will go through the open window of the resist and effectively eat the silicon underneath.

However, fabrication is actually not an easy work and needs patience and experience. There will be always some deviations between the fabricated pattern and the original design input to the EBL. For the fabrication of a shallow etched grating coupler, the etched line is usually bigger than the input pattern due to the so called proximity effect related to the electron scattered by substrate or certain unknown impurity inside. Those scattered electrons may invade into the area adjacent to the focused spot and broaden the exposure region of the resist. Another difficulty is the control of the etch depth. The etch speed is affected by many factors like the portion of the etchant, the kinetic energy of the ions, the environmental condition of the chamber and the feature size of the etching structure. The former three factors mentioned depend on the condition of the ICP-RIE machine. We usually put some big etching windows together with the grating pattern, so that a surface profiler can be used to get a rough evaluation of the etch depth. The last one is related to the lag effect [96]. The result of it is the etching nonuniformity for different feature size (etch width of the groove for a grating coupler), i.e., the smaller the etch width, the shallower the etch depth.

![Graphs showing etch depth vs. etch width](image)

**Fig. 3.9.** (a) Proximity effect induced line expansion; (b) Lag effect induced etch nonuniformity. Both are demonstrated on two samples with different etch times

Both the proximity effect and lag effect could be compensated after a careful adjustment of the EBL parameters like the dose, the voltage or the optimization of the etching recipe, but the exploration of the rule would need numerous attempts and take lots of time. For our devices, we use a pre-compensation method to calibrate those deviations. What we need is the mapping relation from the designed
width to the practical width and the etch depth under a fixed etching condition. Fig. 3.9 shows the mappings for our amorphous silicon wafer fabricated in our lab. Two samples under different etch time are demonstrated. The width expansion can be well fit by a linear curve as depicted in Fig. 3.9(a). The etch time does not affect it a lot, proving that it is mainly related to the proximity effect of the resist. Lag effect can be clearly observed in Fig. 3.9(b). It becomes more prominent when the line width is small, e.g. less than 300 nm.

Although in most cases, those side effects are detrimental, they can be utilized to achieve some special structure which can not be achieved by a common way. For example, the lag effect can be employed in the nonuniform grating coupler design. The reducing etch depth followed by the shrinking etch width can achieve more smooth transition between the blank waveguide and the grating region and give more freedom in the tuning of the radiation leakage. More detailed description can be found in Section 3.4.3.

3.3.2 Characterization

One promising applications of the grating coupler is for wafer-scale testing in a vertical coupling platform. The elimination of the cumbersome operation like the dicing and the facet polishing makes it convenient and time-effective. Fig. 3.10(a) shows the vertical coupling platform in our characterization lab. The setup is actually very simple, consisting of two normal three dimensional adjustors, a sample chuck and a pair of fiber clamps. Currently, the tilt angle of the fiber clamp is fixed at e.g. 15° in the photo although it can be replaced. It would be much better to use a tunable design so that it can adapt different grating designs. This vertical coupling platform can be possibly put into a standard electronic testing platform once a compatible fiber probe can be developed. That is just a mechanical problem.

Fig. 3.10. (a) Vertical coupling platform in our characterization lab; (b) Microscope photo of the setup for the characterization of a pair of grating couplers. Index-matching gel is used in the setup.
Fig. 3.10(b) is a close view of the probing part under a microscope during a testing. In order to reduce the reflection at the fiber’s facet, an index-matching gel with a refractive index close to the fiber is employed. The gel can be transparent if undisturbed and the component underneath is visible. But after the fiber facet touches the gel, the gel in the vicinity of the touching spot will look opaque and make the alignment difficult. The practical adjustment of the fiber position is hence separated into two steps: firstly, keep the fiber close enough to the gel but avoid the touching, then move the facet under the microscope to get a maximum output. Secondly, low down the fiber to let it enter the gel, at the same time, fine adjustments are needed to track the position for a maximum output. The first step is very important or we may easily lose the target in the second step.

Grating coupler is a polarization sensitive device. The polarization of the input light should be manipulated to be compatible with the grating coupler, so is the device under test (DUT). Fig. 3.11(a) shows the common experimental setup using a tunable laser (TLD) and a polarization controller (PC). Although the polarization in the fiber is still unknown after the PC, it is reasonable to assume that the right polarization input is achieved when we get the maximum output at the photode-
3.4 Improving the Coupling Efficiency

3.4.1 Introduction

An ideal interface between the silicon circuits and the fiber network would be expected to be lossless. It points out that the target for a coupler design is a coupling efficiency as high as possible. However, as the most important merit of figure, the coupling efficiency of the grating coupler is still not competitive. The low coupling efficiency, with a maximum of typical 20%–35% [82], is still not competitive for practical applications.

Considering an output coupling setup as depicted in Fig. 3.12, part of the light will be reflected back into the waveguide once it reaches the etched grooves. The part entering the grating region will be diffracted upwards or downwards at a radiation angle determined by Eqn. 3.2. The attenuation in the grating waveguide can be defined by \( \alpha(z) = -\frac{dP(z)}{2P(z)dz} \). \( \alpha \) is usually called as leakage factor or coupling.
Chapter 3. Silicon Grating Coupler

strength for the grating couplers. $P(z)$ is the power inside the grating waveguide and $dP(z)$ presents total radiation power at $z$. Obviously, only the power going towards the fiber facet can be collected by the fiber. But there will be a coupling loss due to the mode mismatching between the radiation mode and the fiber mode.

Typically, the coupling efficiency of a grating coupler can be calculated by

$$
\eta = \frac{P_{\text{in}} - P_R - P_T}{P_{\text{in}}} \cdot D \cdot O_f
$$

(3.13)

The first item presents the fraction of the power diffracted by the grating, where $P_{\text{in}}$, $P_R$, $P_T$ is the input power, reflected power and transmitted power, respectively. The second item $D$ presents the directionality, defined by the portion of the upward power over the whole diffracted power. The directionality is given by

$$
D = \frac{P_{\text{up}}}{P_{\text{up}} + P_{\text{down}}}
$$

(3.14)

The last item presents the extent of the mode matching. It can be calculated by using the overlapping integral between the upward light and the fiber mode

$$
O_f = \frac{\left| \int \int \mathbf{E} \times \mathbf{H}_{\text{fib}}^* \right|}{\Re \int \int \mathbf{E} \times \mathbf{H}^* \cdot \Re \int \int \mathbf{E}_{\text{fib}} \times \mathbf{H}_{\text{fib}}^*}
$$

(3.15)

A lot of research effort has been devoted to improving the coupling efficiency [97–100]. Those progresses can be mainly divided into two categories: the increase of the directionality $D$ and the improvement of the mode matching $O_f$. 

**Fig. 3.12.** Sketch of an output grating coupler.
3.4.2 Directionality and Improved VCM Method

In Sec. 3.2.2, we have introduced the antenna theory based on VCM approximation to explore the physical mechanics of the grating coupler. The radiation from the current source is assumed to freely enter the background material, without any interface reflecting or scattering. The radiation is symmetric along the horizontal line and the directionality is $0.5$. That can be seen right for a weak confined waveguide since the reflection at the interface is very weak. However, for an SOI grating coupler, the typical reflection at the interface of e.g. silicon and silica can be as high as $8\%$ due to the large index contrast. The reflected light may constructively or destructively interfere with the radiated power and in result affects the radiation distribution greatly. Such phenomena are proved by both the full wave simulation and the experimental results [100].

![Fig. 3.13. Depiction of the radiation including the interface reflection, for the case with an infinite buffer layer. (a), upward radiation; (b) downward radiation.](image)

Considering the two parallel beams $a$ and $a'$ from the same radiation point $r'$ as shown in Fig. 3.13(a), it is easy to find the path difference between them for TE polarization is $2(y + t/2) \cos(\theta)$, where $y$ is the radiation point’s coordinate along the vertical axis and $a'$ experiences one more reflection. Hence, the depth factor in Equ. 3.7 for beam $a'$ can be given by:

$$F_D(a') = \sqrt{R} \int_{t/2-d}^{t/2} \cos(ky)e^{-jk_0 y}e^{2jk(y+t/2)} \cos \theta \, dy \approx \sqrt{Req_{2} e^{\Delta_{up}}}$$

where, $k = k_2 = n_2k_0$ should be the wave-vector in the core material and $R$ is the reflection coefficient, given by Fresnel equation [101]. The total upward field can be written as:

$$E_x(r) \propto F_D(a) + F_D(a') = F_D(1 + \sqrt{R}e^{\Delta_{up}})$$

$$\Delta_{up} = (2t - d)k \cos \theta$$
Similarly, the path difference between the downward beams $b$ and $b'$ (see Fig. 3.13(b)) is $2(y - t/2) \cos \theta$, the depth element factor for $b'$ should be revised as:

$$F_D(b') = \sqrt{R} \int_{t/2 - d}^{t/2} \cos(\kappa y)e^{-jky\cos \theta} e^{2jk(y-t/2)\cos \theta} dy \approx \sqrt{R}Re^{-j\frac{t-d}{2}k\cos \theta} \frac{\sin (dk\cos \theta/2)}{ dk\cos \theta/2}$$  

(3.19)

And the total downward light will take the same form of Equ. 3.17, except that the phase difference should be replaced by

$$\Delta_{down} = -dk\cos \theta$$  

(3.20)

Fig. 3.14. Comparison of the directionality enhanced and reduced cases. (a),(b), $t = 340$ nm, $d = 5$ nm; (c),(d), $t = 340$ nm, $d = 190$ nm.

To be clear, we take a new case with a waveguide thickness of 340 nm. For Fig. 3.14(a),(b), with the etch depth of 190 nm, the total phase difference for the upward beams is about $1.05\lambda$ and the difference for the downward beams is around $0.41\lambda$; For Fig. 3.14(c),(d), with the etch depth of 5 nm, the total phase difference for the upward beams is about $1.46\lambda$ and the difference for the downward beams is around $0.01\lambda$. It is obvious that the upward radiation and the downward radiation are not equal due to the filtering of the depth element factor. Over 5 dB suppression ratio
can be found in both cases. Going back to the grating coupler design, we expect that the radiation power can be directed to the most at the main lobe upwards or counted by the directionality as big as possible. So the radiation shown in Fig. 3.14(b) is preferable since it may lead to a high coupler efficiency.

**Fig. 3.15.** Directionality as a function of the etch depth for TE polarization calculated by (a) Improved VCM method; (b) EME simulation for a series slab thickness. The other structure parameters are: $w = 200\, \text{nm}$, $\theta = 15^\circ$ (fixed by tuning the grating period), $n_1 = n_3 = 1.46$, $n_2 = 3.48$, $\lambda = 1.55\, \mu\text{m}$.

More general studies of the relation between the directionality, the slab thickness and the etch depth are shown in Fig. 3.15, calculated by using both the improved VCM method and the numerical EME simulation. The results are plotted as the function of the etch depth for a silicon waveguide thickness of 220 nm, 260 nm, 300 nm, 340 nm, and 380 nm, respectively. The output angle of every structure is fixed to be 15° with a variable grating period for a TE polarization at $\lambda = 1.55\, \mu\text{m}$. We can conclude that the improved VCM model works well for a theoretical guidance of the grating design. The value of the directionality and the trend of the curves actually agree with the numerical output. The deviation between the two methods occurs for the case with a large etch depth, where it seems that the residual silicon slab drags the light spot downwards a bit faster than what the VCM model predicts.

From Fig. 3.15, a high directionality can be achieved by appropriate choice of the slab thickness and the etch depth. The enhance originated from the constructive interference inside the slab waveguide is revealed by the improved VCM model and confirmed by the EME simulation. However, such points corresponding to the slab thickness beyond 300 nm and the deep etch depth of about 200 nm are usually not a good choice for the PIC circuits since it may introduce the high-order modes and serious reflections. Selectively increasing the slab thickness only in the grating region can avoid this problem to certain extent and have the benefit of the enhanced directionality, although suffering an increased fabrication complexity [97]. For a shallow etched grating coupler on a common wafer with a slab thickness less
than 300 nm, the slab thickness of around 260 nm will be preferable for a better directionality.

![Diagram of upward radiation](image)

**Fig. 3.16.** Depiction of the upward radiation including the interface reflection, for the case with a finite buffer layer.

In common, the practical grating coupler is usually defined on an SOI wafer with a finite buffer layer. The buffer layer will strongly affect the directionality and the coupling efficiency of a grating coupler. It is easy to extend the above model to include the buffer layer as shown in Fig. 3.16. For the upward radiation, the direct influence of the buffer layer is the reflection coefficient in the Equ. 3.17. Using the formula of multiple-beam interference with a plane-parallel plane [101], Equ. 3.17 will be revised as:

\[
E_x(\vec{r})_{up} \propto F_D (1 + \frac{(1 - e^{j\Delta_{buf}})^{\sqrt{R}} e^{j\Delta_{up}}}{1 - Re^{j\Delta_{buf}}}) 
\]

(3.21)

where, \( \Delta_{buf} = 2t_{buf}k_1 \cos \theta \). Correspondingly, the downward power should be multiplied by the transmission ratio:

\[
E_x(\vec{r})_{down} \propto F_D (1 + \sqrt{Re^{j\Delta_{down}}} \frac{1 - R}{1 - Re^{j\Delta_{buf}}}) 
\]

(3.22)

**Fig. 3.17** shows the relation between the directionality and the buffer thickness, which is actually a familiar thing for a grating coupler designer [99, 102]. The good fit between the improved VCM model and the EME method reveals that the buffer layer can be treated as a reflection enhanced film or anti-reflection film depending on the thickness. The change of the directionality is mainly due to the multiple beam interference in the film system.
3.4. Improving the Coupling Efficiency

![Graph showing directionality as a function of buffer layer thickness.](image)

**Fig. 3.17.** Directionality as a function of the thickness of the buffer layer for TE polarization at $\lambda = 1.55\,\mu m$ with the design parameters: $w = 200\,\text{nm}$, $t = 260\,\text{nm}$, $d = 90\,\text{nm}$, $\Lambda = 629\,\text{nm}$, $n_1 = n_3 = 1.46$, $n_2 = 3.48$

Following the same flow as the buffer layer, more advanced structure like the bottom Bragg reflector used in our paper $H$, the metal mirror in [99] or the top mirror [102] can be easily included in this improved VCM model.

### 3.4.3 Mode Matching and Nonuniform Design

The output from a fiber facet can be approximated to be Gaussian-profile. However, for a standard uniform grating coupler, the radiation light renders an exponential profile due to the uniform leakage. The mode incompatibility between them will bring at least 1 dB coupling loss [102]. A better mode matching requires a variable leakage factor along the propagation direction. It can be achieved by a nonuniform grating design with difference in either the groove depth [75] or the groove width [74, 98]. The presence of the lag effect shown in Sec. 3.3.1 makes it possible to change the groove depth and the width simultaneously, indicating an enhanced variation range of the leakage factor with reasonable feature sizes. Fig. 3.18 depicts the proposed nonuniform structure.

In order to investigate the lag effect, we use a Raith 150 EBL system to define a series of lines with step-increased widths in the ZEP520 resist. The pattern was transferred into the SOI structure by the ICP-RIE etching. Then the fabricated sample was cleaved and investigated by using a scanning electron microscope (SEM). The width and the depth of each etch groove were measured to get the etching nonuniformity. All the data together with a SEM photograph of the cross section of the testing sample are shown in Fig. 3.19, where we can see the obvious phenomenon of the lag effect especially when the etch width is smaller than 300 nm.
**Fig. 3.18.** Schematic structure of the present nonuniform SOI grating coupler. Blue curves (solid) represent the input and output power distributions when light is propagating upward from the waveguide to the fiber; green curve (dashed) represents the leakage factor distribution in the grating region in order to achieve a Gaussian profile output; the dashed box shows a grating cell.

**Fig. 3.19.** Relation between the etch width and etch depth (lag effect) obtained through SEM measurement. Inset: SEM micrograph of the cross section of a testing sample.

The experimental data are fit by a third order polynomial expression, which de-
3.4. Improving the Coupling Efficiency

terminates the relation between the etch width and the etch depth in the following design.

It is not trivial to design a nonuniform grating coupler since every groove should be individually treated and the corresponding grating pitch should be carefully tuned to achieve an expected radiated angle for the upward wave. For simplicity, we treat the nonuniform grating structure as a sequential combination of different grating cells, which can be analyzed by the Blochmode analysis using CAMFR. For a grating cell with known etch width (as well as etch depth according to Fig. 3.19) and output angle, we can tune the cell width to match the radiation mode at the designated wavelength. Once the cell structure is fixed, the leakage factor of the grating cell can be extracted by numerically monitoring the propagation loss in the corresponding uniform grating coupler. Note that only the leakage loss is included during the propagation simulation.

![Graph showing etch width vs. cell width and leakage factor]

**Fig. 3.20.** Calculated mapping from the etch width to the cell width and the leakage factor to achieve 15° tilt of output beam at 1520 nm.

Fig. 3.20 summarizes the mapping from the etch width to the grating cell width and the leakage factor. Based on the mapping, a nonuniform grating coupler can be sequentially assembled one grating cell by one grating cell according to the leakage factor distribution. The groove width in our design is between 60 and 340 nm. Fig. 3.21 shows the calculated waveguide-to-fiber coupling spectrum for the designed nonuniform grating coupler, as well as the power radiated upward and downward. We can see a maximum coupling efficiency of 74% (due to the good mode matching) near 1520 nm. The reflection at the waveguide to the grating interface is almost eliminated and the coupling loss is mainly due to the downward leakage, which implies that the efficiency can be improved further if the directionality of the grating coupler is optimized.

The inset in Fig. 3.21 shows the SEM top view of a fabricated nonuniform grating coupler. The grating was etched under the same condition as the testing sample for etching nonuniformity. The coupling efficiency of a grating coupler
Figure 3.21. Theoretically and experimentally obtained fiber-to-chip coupling spectra for TE polarization. Maximum coupling efficiency of 64% (−1.9 dB) and 1dB bandwidth of 43 nm were obtained experimentally. Inset: SEM top view of a fabricated nonuniform grating coupler.

can be obtained by characterizing the transmission spectrum of a pair of grating couplers, provided that the propagation loss due to the bridge waveguide between the two gratings is small enough to be ignored. The measured coupling spectrum for the TE polarization was also shown in Fig. 3.21 for comparison. One can see that a maximum coupling efficiency of 64% (about −1.9 dB) is achieved at 1524 nm and the 1 dB bandwidth is about 43 nm. The coupling efficiency is almost two times larger than that for a standard uniform grating coupler and the wideband makes it possible for application both at 1490 and 1550 nm.

3.5 Coupling for Different Polarizations

3.5.1 Polarization Issues

In fact, the so called single mode fiber supports two fiber modes. The two modes are orthogonal and can be used to transmit optical signals individually. Based on this feature, Polarization-Division Multiplexing (PDM) is widely employed to double the spectrum efficiency in a fiber communication system [103, 104]. Typically, fiber-compatible device should be polarization insensitive in order to avoid the polarization-dependent loss (PDL).

However, it is critical to fabricate polarization insensitive silicon waveguide, especially for the circuits based on strip waveguide due to the large refractive index contrast. Usually only one interesting polarization (mainly TE polarization) is addressed for practical device. The coupling for the TM polarization is required by
some applications of e.g. the surface plasmon device [105] and the horizontal slot waveguides [106, 107]. Those devices may have potential advantages in the field of optical sensing, bio-photonics and nonlinear optics.

The grating coupler discussed above only supports the TE polarization. But the design method and tools are general and can be also applied to the grating coupler for TM polarization. The main difference between the two polarizations is the effective index. For the SOI wafer, \( n_{\text{eff}} \) for TM polarization is usually much smaller than that for TE polarization, leading to a larger grating period. Besides, the confinement capability of the TM polarization is a bit weaker than the TE polarization. Correspondingly, the coupling strength is usually stronger for a similar structure. However, with appropriate design parameters, the coupling efficiency of a grating coupler for TM polarization can be competitive with that for the TE polarization.

3.5.2 Polarization Beam Splitter Using a Bidirectional Grating Coupler

Polarization beam splitter (PBS) plays an indispensable role in some occasions where polarization control is required, e.g. polarization-division multiplexed optical communication [103, 104], polarization diversity systems [108], polarization-based signal processing [109]. For the SOI PICs, PBSs based on directional coupler [110], Mach-Zehnder interferometer [111], engineering photonic crystal structure [112], two-layer taper [108, 113] and arrayed waveguide grating (AWG) [114] have been reported before. However, those devices still need polarization-insensitive interface including couplers and waveguides to transmit the light from outside optical fiber to their input ports. Such interface is still critical in design and fabrication.

A promising method is to achieve the polarization splitting and the fiber-to-chip coupling simultaneously by one device. Through a 2D grating coupler [76, 77], the two orthogonal polarizations from one fiber can be coupled into two waveguides with identical modes. Such a 2D grating coupler, indeed a PBS, polarization rotator and coupler, is perfect for the polarization diversity systems used to eliminate the polarization-dependent loss. However, the design and optimization for such a 2D grating coupler (a cumbersome 3D problem) is relatively complex in order to obtain high coupling efficiency for both polarizations [102]. One dimensional grating coupler can also be used as a PBS, through which the two polarizations are coupled into the grating waveguide but propagating towards opposite directions. The two polarizations can then be utilized together with e.g. a polarization rotator in a polarization diversity system or treated separately for their own functions for e.g. wavelength exchange [109].

Principle and Simulation

Fig. 3.22 depicts the proposed structure consisting of a bidirectional grating coupler on an SOI platform. Bottom Bragg reflector is shown as an optional way to improve
Chapter 3. Silicon Grating Coupler

Figure 3.22. Proposed polarization beam splitter based on a bidirectional grating coupler.

Figure 3.23. Wave-vector diagram for the polarization splitting.

the coupling efficiency. Paper C has clearly revealed the principle from the viewpoint of an input grating coupler. As a supplement, here we give the explanation from the viewpoint of an output grating coupler. Referring to the Bragg condition (Equ. 3.3) and the wave-vector diagram shown in Fig. 3.23, the difference for the two polarizations in the output case is the sign of the radiation angle rather than the diffraction order. We can further conclude that the necessary condition for a PBS is

\[ \beta_{TE} + \beta_{TM} = 2k_g \]  

(3.23)

where \( \beta_{TE} \) and \( \beta_{TM} \) are the propagation constants of the grating waveguides for TE polarization and TM polarization, respectively. For the fundamental mode, typically, \( \beta_{TE} > \beta_{TM} \). Hence, the radiation angle for the TE polarization is positive.
and that for TM is negative.

Figure 3.24. The wavelength dependence of the coupling efficiency and the extinction ratio for a PBS using an amorphous SOI wafer. The design parameters are $a = 60 \, \text{nm}$, $\Lambda = 596 \, \text{nm}$, $N = 18$, $\theta = 15^\circ$, $p_1 = 4.0 \, \mu\text{m}$, $t_{wg} = 260 \, \text{nm}$ and $t_{buf} = 3.0 \, \mu\text{m}$.

Paper C has demonstrated the detailed design procedure and a PBS design for a crystal silicon-on-insulator wafer. For the experimental study, we use a home-made amorphous silicon-on-insulator wafer to make the PBS. The main difference is that the refractive index for an amorphous silicon is higher than the normal crystal silicon. A value of 3.6 is used in our design. Fig. 3.24 shows the wavelength dependence of the coupling efficiency and the extinction ratio for a basic PBS design, whose structure parameters are listed in its caption. The coupling efficiency of 50\% is achieved for both the TE polarization and TM polarization. The extinction ratio is better than $-20 \, \text{dB}$ in almost the whole $3 \, \text{dB}$ band of over 70 nm. The polarization splitting behavior is clearly rendered by the field distributions ($E_x$ for TE polarization and $H_x$ for TM polarization) in Fig. 3.25. Those results are obtained by 2D FDTD simulations, where the grating is illuminated from the top by a Gaussian beam with a beam diameter of 10.4 $\mu\text{m}$ at $\lambda = 1550 \, \text{nm}$ as schematically shown in Fig. 3.22. The performance of this PBS can be further improved by using the Bragg reflector under the buffer layer, where the downward leakage can be effectively reduced. The Bragg reflector is made by alternatively depositing a layer of 105 nm thick silicon and a layer of 265 nm thick silica. The fabrication of this structure is compatible with the process for the amorphous silicon wafer. Fig. 3.26 shows the calculated coupling spectrum and the extinction ratio. Although only one pair of Bragg reflector is included, the coupling efficiency has been significantly increased to 63\%, about 1 dB improvement. Coupling efficiency of over 70\% is expected if employing more pairs.
Figure 3.25. (a) Field distribution ($E_x$) for TE polarization; (b) Field distribution ($H_x$) for TM polarization at $\lambda = 1550$ nm. The design parameters are listed in the caption of Fig. 3.24.

Layout and Characterization

For the proposed PBS, although additional functionality of polarization splitting is added to the grating coupler, the fabrication is the same as an ordinary grating coupler. However, since the negative angle is introduced, we should pay attention to the mask layout for ease of characterization. The common layout of connecting a pair of grating couplers with a straight waveguide will suffer practical problem of alignment. Instead, neighboring grating couplers are bridged by a S-bend waveguide with an offset as shown in Fig. 3.27 to avoid touching of fibers. Each polarization is designed to couple into one port of a grating coupler, but there might be a little part going to the other port as shown in Fig. 3.27(a). For TE polarization, the portion of the power at the expected port over total input power is defined as $a$ and the portion at the unexpected port is defined as $a'$. Similarly, $b$ and $b'$ are defined for the TM polarization. Since the coupling is reciprocal, the definitions above can also be applied to the output grating coupler. Fig. 3.27(b) presents the pattern for the characterization of the coupling efficiencies of the PBS. Since both polarizations are involved, it is more convenient to employ the setup using an ASE source as shown in Fig. 3.11(b). During the characterization, the fiber connecting to the ASE source is kept on the middle grating coupler as the input. The other fiber can be flexibly switched between the other two grating couplers, collecting the
3.5. Coupling for Different Polarizations

Figure 3.26. The wavelength dependence of the coupling efficiency and the extinction ratio for a PBS using the amorphous SOI wafer, with a pair of Bragg reflector underneath. $N = 16$, $p_1 = 3.3 \, \mu m$, $t_{Si} = 0.105 \, \mu m$, $t_{SiO_2} = 0.265 \, \mu m$, the other design parameters are the same as that shown in the caption of Fig. 3.24.

Figure 3.27. (a) definitions; (b) pattern for characterization of coupling efficiencies for both TE and TM polarizations; (c) pattern for characterization of the extinction ratio of the PBS. Blue arrow presents the TE polarization; red arrow presents the TM polarization.

output power $P_{TE}$ at left or $P_{TM}$ at right (see Fig. 3.27(b)). If the total input of
the ASE source is $P_{in}$, the output can be written as:

$$P_{TE} = 0.5 T_S P_{in} (a \times a + b' \times b)$$  \hspace{1cm} (3.24a)

$$P_{TM} = 0.5 T_S P_{in} (b \times b + a' \times a')$$  \hspace{1cm} (3.24b)

$$P_{TE} = 0.5 T_S P_{in} (a \times a + b' \times b')$$  \hspace{1cm} (3.24c)

where, $T_S$ is the transmission ratio through the S-bend waveguide. For a good design, $a >> a'$, $b >> b'$, the coupling efficiencies for the two polarizations can be given by:

$$\eta_{TE} = a = \sqrt{2P_{TE} / (T_S P_{in})}$$  \hspace{1cm} (3.25a)

$$\eta_{TM} = b = \sqrt{2P_{TM} / (T_S P_{in})}$$  \hspace{1cm} (3.25b)

$$\eta_{TM} = b = \sqrt{2P_{TM} / (T_S P_{in})}$$  \hspace{1cm} (3.25c)

Another important feature for a PBS is the extinction ratio, defined by $b'/a$ for TE polarization and $a'/b$ for TM polarization. The PBS based on the 2D grating coupler in [76] was characterized in a 'Fiber in - cleaved facet out' platform, where the light is input by vertical coupling, but collected laterally at the wafer edge. It needs build a new setup for the testing of the extinction ratio. Here we develop an easy way using the standard vertical platform to characterize the extinction ratio of our PBS. It is achieved by using a special pattern design. As shown in Fig. 3.27(c), for the right part, it is the same as that for the coupling efficiency; for the left part, the power collected at the output ($P_{ER}$) can be described by:

$$P_{ER} = 0.5 T_S P_{in} (a \times a' + b' \times b)$$  \hspace{1cm} (3.26)

Combining with Equ. 3.25, we can get

$$ER_{sum} = 10 \log_{10} (a'/b + b'/a) = 10 \log_{10} (P_{ER} / \sqrt{P_{TM} P_{TE}})$$  \hspace{1cm} (3.27)

Although extinction ratio at a given port can not be distinguished in the above equation, $ER_{sum}$ as the upper limit of the extinction ratio at each port is significant for the evaluation of the performance.

Two generations of the PBS have been fabricated in our lab. They are both based on the home-made amorphous silicon on insulator wafers. Fig. 3.28 shows the SEM image of the sample from the first generation. Detailed structure parameters can be found in Paper E for the first generation and Paper H for the second generation. Fig. 3.29 summarizes the measurement results. For the first generation, the maximum coupling efficiencies for TE and TM are 43% and 33%, respectively. They are improved to over 50% for the second generation with a Bragg reflector underneath. The extinction ratios for both generations are as low as around $-20 \text{ dB}$. 
3.5. Coupling for Different Polarizations

Figure 3.28. SEM top-view of the fabricated PBS based on a bidirectional grating coupler. Inset: the shallow etched grating.

Figure 3.29. Measurement results for the fabricated PBSs. (a) coupling spectrum and (c) extinction ratio for the first generation; (b) coupling spectrum and (d) extinction ratio for the second generation with Bragg reflector underneath.
Chapter 4

Summary, Conclusions and Future Work

In summary, electroabsorption modulators and grating couplers for silicon platform have been investigated in this thesis. They are the important components for the optical interconnects, especially for the inter-chip interconnects using the silicon-on-insulator wafer.

Recent progress in optical modulators based on either silicon substrate or InP substrate has been reviewed. Significant developments have been done for silicon modulators, with the performance improvements of pure silicon modulators and the emerging technologies of silicon-compatible modulators. However, the progress of silicon-based modulator is still trivial, if compared with the performance of its InP-based counterpart. Hybrid silicon evanescent platform, with the III/V materials being transplanted on the SOI wafer, provides the basis to use the mature InP-based techniques on silicon platform.

In this thesis, a general and comprehensive small-signal model based on the transmission line analysis, as well as the optimization method based on the genetic algorithm, has been developed for the modulator design with lumped, traveling-wave, capacitively-loaded or segmented configuration. This method has been applied to the design of the segmented hybrid EAMs, with significant improvements contributed by the following three aspects: (a) the RC limit related to the source impedance is eliminated by the distributed electrode; (b) good impedance match can be achieved at the source side by tuning the lengths of the passive high-impedance transmission lines; (c) constructive reflection can be obtained by the cascaded connection alternating between the active low-impedance modulators and the passive high-impedance TMLs. The first item can effectively increase the bandwidth to a double or more. The latter two items can improve the feed-in efficiency and hence reduce the driving voltage for a certain extinction ratio. They can also extend the bandwidth to some extent if the efficient enhancement occurs in the high frequency range. The fabrication of the designed hybrid EAMs is still in progress.
and the experimental result will be reported in the future work.

We have revealed that a grating coupler can be essentially treated as an optical phased array antenna with the radiation spots inside a multiple film system. An improved volume current method is developed to give a rough analytic evaluation of the directionality. It can be used to choose the initial wafer structure as the first step for a grating coupler design. The band analysis can then be employed to get the right grating period for an expected radiation angle, as well as the leakage factor of a grating cell. We can then find the optimal grating structure for a uniform grating coupler or the appropriate cells to assemble a nonuniform grating coupler. The final structure is recommended to be evaluated by using a FDTD simulator, with the coupling spectrum obtained after one running.

We have explored two methods to improve the performance of the grating coupler. For the direct way to improve the coupling efficiency, we have successfully fabricated a nonuniform SOI grating coupler by utilizing the lag effect during the ICP-RIE etching. The maximum coupling efficiency is improved to 64% (−1.9 dB), nearly a double of a standard uniform grating coupler. For the other way to add more functionalities to the grating coupler, a polarization beam splitter (PBS) based on a bidirectional grating coupler has been proposed and experimentally demonstrated. An extinction ratio of around −20 dB, as well as a maximum coupling efficiency of over 50% for both polarizations, is achieved by such a PBS with a Bragg reflector underneath.

The future work can include:

- For the hybrid modulators discussed in this thesis, there is still room for the improvement in bandwidth and insertion loss by optimizing the structure. Firstly, the passive microstrip can be replaced by using the coplanar waveguide (CPW) with the convenience of tuning the characteristic impedance by simply changing the gap between the signal strip and the ground metal rather than the change of the polymer layer thickness. The CPW structure may also lead to a lower loss for a high characteristic impedance. Once using CPW structure, with electrode on the top surface of the polymer and the optical circuits beneath, the main electrical circuits and the photonic integrated circuits can be vertically separated in two different layers. This isolation will greatly reduce the influence of the lossy metal to the optical circuits, increase the effective layout area for the optical components and ease the whole device design. Secondly, as we mentioned before, the passive optical segment in the segmented EAM can be designed to be a semiconductor optical amplifier to compensate the static optical loss from the modulation parts; we can also try to remove the lossy III/V material in that region to reduce the optical loss. Thirdly, the current hybrid waveguide is indeed not good enough for a more higher bandwidth e.g. towards 100 GHz due to the strict intrinsic RC limit. Optimization of the waveguide structure such as increasing the thickness of the intrinsic layer can be attempted, together with careful evaluation of the compromise of the bandwidth, the extinction ratio and the optical insertion
loss. Furthermore, a significant vision is to develop the hybrid electroabsorption modulated lasers (EML) for wide-band optical interconnects used for e.g. high speed computations. It can be the next work after a mature hybrid EAM is available.

- Although the hybrid EAM can be pushed further to support a higher speed, the spectra efficiency is actually very low due to the on-off keying modulation. By embedding the hybrid EAMs in an MZI, we may get a compact and high speed silicon-based advanced modulation format generator, which can support a data transmission rate multiple times of the bandwidth of the modulator and the driving circuits. To simplify the system, such devices can be further monolithically integrated with the electronic driver, where either the silicon or the InP can be used for the electric design although new infrastructure should be explored to reasonably arrange the electrical layer and the optical layer. Putting the driver and the modulator close can reduce the microwave loss in the lossy feed line and improve the system performance in the power dissipation, packaging complexity etc.

- Applying the advanced electrode (capacitive loaded type) to the pure silicon modulator based on carrier depletion is a possible way to achieve over 40 GHz silicon modulation, with the advantages of simple process and seamless interface with other silicon PICs although the large footprint may be an issue for compact integration.

- One of the disadvantages of a standard grating coupler is that it needs an additional etching operation. Nonuniform concept can be applied to the fully etched grating coupler to reduce the reflection loss. It can be also extended to the 2D grating coupler. But more efficient numerical simulation method is required for the design and the optimization of the 2D grating coupler. The popular FDTD method is not a good choice for the 3D simulation of a grating coupler due to the large computation scale. Integral-equation-based method like the MoM with certain fast algorithm would be more useful for this problem. Some other PhC grating couplers with more complex lattice can be explored, e.g., for a possible triplexer using a rectangle-lattice photonic crystal.

- The bidirectional-grating-based polarization beam splitter can be further integrated with detectors to be the receiver for the polarization division multiplexed optical communication system, probably needing a polarization tracking scheme.
Chapter 5

Summary of the Original Work

**Paper A:** Modeling and Optimization for Segmented Transmission-Line Electroabsorption Modulators with Asymmetrical Electrodes

Normalized RF link gain is derived for a segmented transmission-line electro-absorption modulator. Genetic algorithm is used to optimize the electrode structures which are asymmetrical and non-periodic. Performance of the optimized EAM design is analyzed.

*Contributions of the author:* The main original idea, all simulations, and the manuscript.

**Paper B:** Design and Optimization of an Arbitrarily Segmented Traveling Wave Electrode for an Ultrahigh Speed Electroabsorption Modulator

The modulation response of an arbitrarily segmented traveling wave electroabsorption modulator (TW-EAM) with a non-periodic electrode layout is analyzed with an effective approach based on the transmission line theory. An optimization method based on a genetic algorithm is used to optimize the segmented electrode with additional capacitive pads for a TW-EAM. Optimized electrode structures are given for 165 µm-length devices under standard 50 Ω RF termination condition and 210 µm-length devices with smaller integrated resistors. Great enhancement of performances in modulation response is achieved as compared to the periodic design reported previously. The simulation with the parameters extracted from measurement results suggests that a 3 dB-bandwidth over 100 GHz could be achieved. The eye diagram from the simulation demonstrates that such devices would be competent for a 100 Gb/s optical network.

*Contributions of the author:* The main original idea, all simulations, and the first draft of the manuscript.
Paper C: Proposal for a Grating Waveguide Serving as Both a Polarization Splitter and an Efficient Coupler for Silicon-on-Insulator Nanophotonic Circuits

In this paper, a grating waveguide is introduced and designed to serve as both a polarization splitter and an efficient vertical coupler between a fiber and a silicon-on-insulator (SOI) nanophotonic waveguide. Through this grating waveguide, the light from the fiber can be efficiently coupled into an SOI chip where the two orthogonally polarized waves are separated to travel towards the opposite directions along the waveguide. According to our simulations, the optimized structure can give a high coupling efficiency of about 50% for both polarizations, as well as a large bandwidth of over 70 nm and a very low polarization crosstalk below $-22$ dB at the output ports.

Contributions of the author: The original idea, device design and simulation, and the first draft of the manuscript.

Paper D: Ultracompact Low-loss Coupler between Strip and Slot waveguides

We present both theoretical and experimental results of an ultracompact waveguide coupler that is capable of highly efficient coupling of light from strip waveguides to slot waveguides, and vice versa. By optimizing the geometrical parameters, it is possible to achieve extremely low-loss coupling. A coupling efficiency of 97% has been obtained experimentally while keeping the overall size down to the range below 10 $\mu$m. Further analysis shows that the proposed coupler has relatively high tolerance to fabrication errors and is wavelength insensitive.

Contributions of the author: Grating coupler design, Participate in the characterization.

Paper E: Experimental Demonstration of an Ultracompact Polarization Beam Splitter based on a Bidirectional Grating Coupler [Best student paper award]

This is the experimental extension of Paper C. A bidirectional grating serving both as a polarization beam splitter and a vertical coupler for silicon on insulator nanophotonic circuits is fabricated and characterized. The measured coupling efficiency is as high as 43%. The demonstrated device has a large 3 dB bandwidth and a high extinction ratio between two orthogonal polarizations.

Contributions of the author: The original idea, device design, simulation and characterization, and the revision of the manuscript.

This is a preliminary work of Paper G, where a nonuniform is designed and fabricated for an amorphous SOI wafer utilizing the lag effect in dry etching. Over 80% (> −1 dB) coupling efficiency is theoretically obtained and experimental coupling efficiency of 55% is achieved for TE polarization.

Contributions of the author: The original idea, device design, simulation and characterization, the first draft of the manuscript and the presentation.

Paper G: Highly Efficient Nonuniform Grating Coupler for Silicon-on-Insulator Nanophotonic Circuits

In this paper, we present design, fabrication and characterization of a silicon-on-insulator grating coupler of high efficiency for coupling between a silicon nanophotonic waveguide and a single mode fiber. By utilizing the lag effect of the dry etching process, a grating coupler consisting of nonuniform grooves with different widths and depths is designed and fabricated to maximize the overlapping between the upward wave and the fiber mode. Measured waveguide-to-fiber coupling efficiency of 64% (−1.9 dB) for transverse electrical polarization is achieved by the present nonuniform grating coupler directly defined on a regular silicon-on-insulator wafer.

Contributions of the author: The original idea, device design, simulation and main characterization, and the first draft of the manuscript.

Paper H: Experimental Demonstration of a High Efficiency Polarization Splitter based on a One-Dimensional Grating with a Bragg Reflector underneath.

This is a continue work of Paper C. Bragg reflectors are employed to improve the coupling efficiency. Over 50% efficiency for both polarizations are achieved experimentally, and the extinction ratio between them is also high (−20 dB).

Contributions of the author: The original idea, and the revision of the manuscript.
Bibliography


