Integrated photonic chip for quantum key distribution

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2019

Zhang, G. (2019). Integrated photonic chip for quantum key distribution. Doctoral thesis, Nanyang Technological University, Singapore.

<https://hdl.handle.net/10356/105274>

<https://doi.org/10.32657/10220/47830>

**NANYANG
TECHNOLOGICAL
UNIVERSITY**

SINGAPORE

INTEGRATED PHOTONIC CHIP
FOR QUANTUM KEY DISTRIBUTION

**INTEGRATED PHOTONIC CHIP
FOR QUANTUM KEY DISTRIBUTION**

ZHANG GONG

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SCHOOL OF ELECTRICAL & ELECTRONIC ENGINEERING

2019

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**INTEGRATED PHOTONIC CHIP
FOR QUANTUM KEY DISTRIBUTION**

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School of Electrical & Electronic Engineering

A thesis submitted to the Nanyang Technological University
in partial fulfillment of the requirement for the degree of
Doctor of Philosophy

2019

Statement of Originality

I hereby certify that the work embodied in this thesis is the result of original research, is free of plagiarised materials, and has not been submitted for a higher degree to any other University or Institution.

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Acknowledgment

First and foremost, I would like to express my sincere gratitude and deepest appreciation to my supervisor, Professor Liu Ai Qun, for his guidance, encouragement, patience and enthusiasm. Whenever I had a problem, about research or life experience, the door of Prof Liu's office was always opening. He acts as a prudent advisor consistently guiding me and inspiring me towards higher achievement. He taught me to face life with a smile no matter how everything is. His help will affect me for a lifetime.

I would also express my profound gratefulness to Professor Kwek Leong Chuan, from Centre for Quantum Technologies, National University of Singapore, for his immense guidance and suggestive discussions on quantum physics since we started our quantum project. He is an excellent teacher with great patience and generosity.

I would like to express my sincere gratitude to Dr Cai Hong, from the Institute of Microelectronics, who is an expert in fabrication technology. She shares me with extensive knowledge about fabrication and experimental testing. I would also like to thank Dr Song Junfeng, Dr Yu Mingbin, Dr Tao Jifang and Dr Ren Min, from the Institute of Microelectronics, for their support and helpful advice.

I would like to thank Professor Xu Feihu for his invaluable advice on experiment and paper writing. I would also like to thank the Professors, Ser Wee, Zhou Xiaoqi, Su Xiaolong, Jia Xiaojun, Wu Jian, Yu Song, Muhammad Faeyz Karim, Sunil Bhawe, Joseph Fitzsimons, Fan Weijun, Gong Jiangbin and Alexander Szameit for their advice and guidance on the theoretical modeling and experimental study. I am also grateful to my collaborators Dr Haw Jingyan, Dr Zhang Yichen and Dr Syed Assad,

who are the experts in continuous variable quantum key distribution. They gave me much support on the quantum key distribution part of this research.

I would also thank my group members, Dr Chin Lip Ket and Dr Liu Yang for their assisting on paper and thesis writings. Also, I would like to thank the seniors of my group, Dr Dong Bin, Dr Huang Jianguo and Mr Lin Jianxiao who guided me when I just started. At the same time, I would also thank the other group members, past and present, for their kindness whenever I had a problem to ask them. They are Dr Zhu Weiming, Dr Lei Lei, Dr Feng Xueling, Dr Wu Pin-Chieh, Dr Xiong Sha, Dr Chen Yan-Yu, Dr Zhang Wu, Dr Fu Yi, Dr Zhao Haitao, Dr Shi Yuzhi, Dr Song Qinghua, Ms Shakeela Selvarajah, Mr Xia Yang, Mr Chen Hongfu, Mr Hsieh Chao Mao, Mr Yan Libin, Ms Zheng Shaonan, Ms Zhang Hui, Mr Li Zhenyu, Mr Wan Lingxiao, Mr Cao Lin, Mr Ding Henghui, etc.

I would also like to extend my thanks to all the technicians from the Institute of Microelectronics, Valens lab, Nanomaterials Lab, Photonics lab I and II and N2FC Lab for their help in offering the resources and technical support in this research.

In particular, I am deeply grateful to my parents and my family for their constant support whenever I faced any challenge. I must thank my girlfriend. Your love, support and encouragement are the torch in the dark that helps me out throughout these years.

Finally, I must express my gratitude to Nanyang Technological University for providing me the opportunity to work in such an excellent research environment.

Zhang Gong

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Summary

Silicon photonic technologies exploit silicon as an optical material to build a wide array of devices and systems. It has a vast and growing market in fields such as telecommunications, sensors, displays, and biomedical instrumentations. The flexibility, modularity and stability of silicon photonic systems provide unprecedented opportunities for next-generation quantum key distribution (QKD) and quantum networking. This doctoral thesis focuses on the design, simulation, fabrication, and experiment of QKD chips based on silicon photonics. Crucial components for QKD chips are designed and fabricated, including the grating coupler, polarization beam splitter, polarization rotator, modulator, photodetector, ring resonator and interferometer. Using these components, chip-based continuous variable quantum key distribution (CV-QKD) is demonstrated for the first time.

The thesis focuses on three parts: the passive components, the active components, and the integrated QKD chip. For the passive components, the two-dimensional grating coupler can be used as a path-to-polarization mode converter, the polarization beam splitter separates or combines two polarization states on-chip, and the polarization rotator rotates the polarization state between transverse electric and transverse magnetic modes. All the polarization manipulation components are essential for expanding the available degree of freedom of the photon. The Mach-Zehnder interferometer and ring resonator are the core components used in the modulator. Both are designed, simulated, fabricated and experimentally characterized.

Active components, including modulators and detectors are also studied in this thesis. For modulator, a thermo-optic modulator and a carrier injection modulator are

designed and experimentally studied. For detector, an on-chip germanium detector is also designed, fabricated, and characterized. As required in the continuous variable quantum information experiments, a balanced homodyne detector using the on-chip germanium detector is developed and analyzed in detail.

Finally, the QKD chip using the Gaussian-modulated coherent state (GG02) protocol is designed and fabricated. Two separate chips are used as the transmitter (Alice) and the receiver (Bob) for QKD. All the required passive and active components are integrated onto the photonic chips. The basic working principle of the GG02 protocol and those of the post-processing protocols are analyzed. The secure key rate is calculated based on the total loss and excessive noise of the system, with tests showing that the secure key-rate under collective attack can potentially reach 0.83 MHz at 0 km and 2.3 kHz at 100 km distance.

Based on the standard silicon photonic fabrication platform, the QKD chip can be integrated onto a single chip together with the classical optical communication systems and even with the controlling integrated circuit. The hybrid QKD chip will ultimately lead to the next generation of hybrid quantum-classical communication networks with more compact components.

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Nomenclature

ALD	Atomic layer deposition
AM	Amplitude modulator
ASE	Amplified spontaneous emission
BARC	Bottom anti-reflective coating
Bps	Bits per second
CMOS	Complementary metal-oxide-semiconductor
CMP	Chemical-mechanical planarization
COW	Coherent one way
CVD	Chemical vapor deposition
CV-QKD	Continuous variable QKD
CW	Continuous wave
DHF	Dilute hydrofluoric acid
DPS	Differential phase shift
DV-QKD	Discrete variables QKD
FEM	Focus exposure matrix
FIB	Focused ion beam
FPGA	Field-programmable gate array
HDP-CVD	High density plasma CVD
LO	Local oscillator
MDI-QKD	Mesurement-device-independent QKD
MEMS	Micro-electro-mechanical system
MMI	Multimode interferometer

MZI	Mach-Zehnder interferometer
NA	Numerical aperture
OSA	Optical spectrum analyzer
PBS	Polarization beam splitter
PD	Photodiode
PE-CVD	Plasma-enhanced CVD
PM	Phase modulator
PVD	Physical vapor deposition
QDR	Quick dump rinse
QKD	Quantum key distribution
RIE	Reactive ion etching
RMS	Root mean square
RTA	Rapid thermal annealing
SEM	Scanning electron microscope
SNU	Shot noise unit
SOI	Silicon-on-insulator
SPM	Sulfuric peroxide mix
TE	Transverse electric
TEM	Transmission electron microscope
TM	Transverse magnetic
USG	Undoped silicon glass
WDM	Wavelength-division multiplexing

Chapter 1 Introduction

1.1 Motivation

This PhD research project focuses on a single-chip solution for quantum key distribution (QKD). This first-time demonstration is a significant innovation for quantum networking. The realization of miniaturized and low-cost QKD system based on silicon photonic fabrication, though challenging, is desired. Chip-based QKD is the ultimate solution for the next generation QKD networking thus far. The successful development of a practical QKD chip may open a new era for on-chip QKD and secure communication network.

Information and communication network security are vital in modern life, and at the operation core of the financial systems, communication networks, industry, military units, diplomatic departments, and governments. They play a crucial role in national and social security. However, there is almost no cost-effective way to ensure the absolute security of information. One of the main security loopholes is the untrusted communication channel. Eavesdropping on the communication channel is unavoidable and, in most case, undetectable. To ensure the security of transmitted information, different encryption methods have been developed. The goal is to generate a cryptogram from the original message and some additional information known as a key. The message must remain secret without the key.

The one-time pad method, proposed in 1926, is one of the first modern encryption methods [1]. It has proven to be the only theoretically unconditionally secure encryption method. It requires each party use the same set of totally random keys of the

same length as the information being encoded. Due to the random nature of the keys, the result as noticed by the eavesdropper is random. Because they use the same secret key for both encryption and decryption, this method is categorized as a secret-key cryptosystem. Although this method ensures unconditional security, the requirement of a large number of random keys raises a new problem in the secret distribution of keys. A trusted courier or even personal meetings are required to avoid loopholes, which would be expensive and complicated in daily life. Another class of cryptosystems is the public-key cryptosystem. Different from the secret-key cryptosystem, it uses different keys for encryption and decryption. The first and the most successful public-key cryptosystem is the RSA (Rivest-Shamir-Adleman) cryptosystem. For RSA, the security is based on the difficulties of factorization of the product of two large prime numbers. However, in 1994, Peter Shor developed an efficient algorithm for integer factorization on a quantum computer with polynomial time [2], which threatened the RSA cryptosystem. Hence, it becomes essential to find a new efficient cryptosystem with unconditional security.

After the development of Shor's algorithm, QKD started to gain attention. As mentioned earlier, the one-time pad method offers unconditional security in the message encryption/decryption process. It would be a perfect match if not for the lack of a highly efficient method to distribute secure keys with unconditional security. Quantum systems are generally too lossy to transmit information directly, but they are perfect to generate random secret keys. The scheme is fully functional as long as the remaining keys between the two parties are the same. QKD utilizes the Heisenberg uncertainty principle and the quantum no-cloning theorem, to ensure the security of the distributed key [3]. The sender prepares a quantum state with some known information encoded. The receiver selects a basis to measure the state and gain part of the

information. After basis reconciliation, they could have a perfectly correlated key. However, if someone tries to eavesdrop in between, not only would it be impossible for the eavesdropper to clone the quantum state, but also the state would collapse into a new state, rendering part of the information lost. Hence, the sender and receiver can detect this information change and know the presence of an eavesdropper.

The development of QKD starts from the year 1984 when Charles H. Bennet and Gilles Brassard proposed the first quantum key distribution protocol, known as the BB84 protocol [4]. After more than 30 years of development, QKD systems have become increasingly stable and practical. Some of the world's leading large enterprises have invested heavy R&D capital in quantum information, contributing significantly to the development of quantum information technology in the industry. This include GOOGLE, INTEL, Microsoft, AT & T, Bell Labs, IBM and Hewlett-Packard in the United States; Philips in the Netherlands; Hitachi, NEC, NTT and Toshiba in Japan; British Telephone and Telegraph Company in UK, and Siemens in Germany. In October 2010, there was a joint exhibition in Tokyo by NEC, Toshiba, Mitsubishi Electronics and several other companies to demonstrate their commitment to the construction of a quantum communication network in Japan. Thus, while several developed countries around the world closely link quantum information with national security, at the same time, large international conglomerates have expressed deep interest in the potential for the commercialization of quantum communication. Nowadays, new requirements are found for QKD such as small size, low cost, high stability, high reliability and the potential for mass production. Like the revolution of the integrated circuit which gave rise to modern information technology, many people believe that integrated photonics is the core technology for the next generation of

quantum information. Among the various possible materials, silicon is still the most attractive one.

Researchers gradually discovered the promising capabilities of silicon photonics in information technologies, such as optical communication, data centers and signal processing, following applications in pressure sensors and gyroscopes in the 1990s. Various groups demonstrated successful integration of all the required components, including power splitters, multiplexers and demultiplexers, modulators, photodetectors, and lasers. Many of the advantages of silicon in semiconductor information technologies were now showing their potential in integrated optics. Silicon is abundant in the earth, has excellent mechanical, chemical and thermal stability, can be made ultrapure and most importantly, is fully compatible with the specially designed equipment for semiconductor technologies. Moreover, electronics and photonics can be integrated together on the same chip to offer a truly integrated photonic system [5]. All these advantages are leading to the realization of a silicon photonics chip with compact size, low cost, high stability, high reliability and ease of mass fabrication. In recent years, commercialized high-speed optical transceivers with rates up to 400 Gbps based on silicon photonics technologies have been developed by many companies such as Intel, Cisco, Macom, Luxtera, Mellanox, and Acacia.

The mature silicon photonic platform opens a new gateway for quantum information applications. Some recent results have already proven the possibility of using silicon photonics for QKD [6-8]. All the components required are available from previous research efforts. Thus, by using silicon photonic technology, a fully packaged QKD chip with all the necessary optical and electrical components can be built with low-cost and high yield. The history of chip-based quantum information will be covered in detail in Chapter 2. However, to protect the ultra-weak quantum signals, most of the

QKD systems now are designed using “dark fiber” for transmission. These requirements necessitate the installation of dedicated fiber for the QKD network, creating the fiber lease and maintenance costs and limiting the scale of the network. Integrating QKD with classical communications in the existing fiber infrastructure remains the optimal solution. Thus, some proof-of-principle works have shown that the QKD channel could coexist with classical optical communications to form a hybrid quantum-classical network using wavelength-division multiplexing (WDM) techniques [9-12]. With this technique, it is possible to integrate the QKD system together with the classical optical communication apparatus to form a hybrid quantum-classical communication chip for the next generation of low-cost and scalable secure quantum networks.

In summary, the development of chip-based QKD system is a goal pursued not only by academic institutions but also by industry parties. In this PhD research, the related design, theoretical analysis, fabrication, and experiments will be covered. The studies of such a system may offer a new possibility for next-generation quantum networks.

1.2 Objectives

The primary objective of this research is to innovate a chip-based QKD system by using various photonics components with silicon nano-photonic fabrication technology. Specifically, a silicon photonic chip platform is developed, including the required active and passive components. The platform is then used to construct a single chip for continuous variable QKD using the Gaussian-modulated coherent-state protocol. The research investigations include theoretical modeling, component design and simulation, silicon photonic fabrication and experimental studies.

The chip-based QKD system consists of a transmitter chip (Alice), a receiver chip (Bob) and the software for post-processing. The transmitter chip generates a Gaussian-modulated coherent state signal and the receiver chip performs homodyne detection on the transmitted states. All the essential components are integrated onto a single chip using the nano-silicon photonic fabrication process. After data reconciliation, error correction and privacy amplification, both the parties should share an identical secret key string. All the losses and noises in the system are carefully tested to ensure the safety of the final secure key.

The passive components are the backbone of the QKD chips. This includes all the components that do not require an external power supply, such as the fiber-chip couplers, power splitters, polarization manipulation components, interferometers, and resonators. The governing equations for the directional coupler, Mach-Zehnder interferometer and ring resonator are derived. Then all the required components are designed, simulated, fabricated, and experimentally verified. For each component, the critical specifications are calibrated to proper integration and function within the in QKD system. Finally, the fabrication process for passive components further developed.

The active components are also studied, including the modulators and detectors. Three different design of modulators, including the thermo-optic modulator, the carrier injection or depletion modulator and the electro-optic modulator, are analyzed theoretically and experimentally to compare their relative advantages. The on-chip germanium detector is also designed and fabricated, which is integrated into a homodyne detector for continuous variable QKD applications. The fabrication process for active components is also further developed.

1.3 Major Contributions

The major contributions of this PhD thesis lie in various aspects of the theoretical model construction, component design and simulation, chip fabrication and experimental studies of a chip-based QKD system. The details are listed below:

- 1) The fiber-chip couplers, including the taper coupler and grating coupler, are designed and experimentally tested. Notably, a two-dimensional grating coupler is designed as a polarization beam splitter/combiner (see Chapter 3).
- 2) The directional coupler is analyzed in detail, serving as the basis for all the mode coupling devices, including the polarization manipulation components, Mach-Zehnder interferometers (MZI), and ring resonators. The MZI and ring resonator are also designed and analyzed in detail (see Chapter 3).
- 3) The multimode interference (MMI) device is studied. Both 1-to-N and 2-to-2 MMI devices are designed and tested (see Chapter 3).
- 4) Two important polarization manipulation components, including the polarization beam splitter and polarization rotator, are designed and tested (see Chapter 3).
- 5) Three types of modulators are designed and studied in detail, including the thermo-optic modulator, carrier injection or depletion modulator and electro-optic modulator. Among them, the doping properties are also calculated in detail to guide the doping profile design. The performance of all three modulators under both static and dynamic responses are tested (see Chapter 4).
- 6) The on-chip germanium photodiode is designed and tested. It is then integrated into a homodyne detection system for quadrature field measurement. (see Chapter 4).

- 7) The fabrication processes for both passive components (see Chapter 3) and active components (see Chapter 4) are developed and analyzed in detail, using the facility in Institute of Microelectronics, A*STAR.
- 8) The CV-QKD Gaussian-modulated coherent state protocol, including the data reconciliation, error correction, and privacy amplification protocols are studied. The first chip for CV-QKD is designed, fabricated and tested. All the noise sources and losses in the system are measured and analyzed in detail. The result is used for the secure key rate analysis of the system. (see Chapter 5).
- 9) A proof-of-principle chip-based CV-QKD test is conducted. The experimental results show a secure key rate of 0.83 MHz at 0 km and 2.3 kHz at 100 km transmission distance under the collective attack assumption, which is comparable with previous results in bulk and fiber optics. (see Chapter 5).

1.4 Organizations

The thesis is organized into six chapters. The introduction of the thesis covers the motivation, objective, and major contributions as presented in this chapter. The motivation section explains why and how the PhD research is carried out. The objective states the primary focus of this thesis, and the contribution section lists the innovations and important findings in both the theoretical and experimental aspects.

In Chapter 2, a literature survey on the current development of silicon photonics, including both passive components and active components, is introduced. Then, a review on the roadmap of quantum key distribution is shown, with a general survey on the two major categories, i.e., continuous variable QKD and discrete variables QKD. The third part presents a review of the chip-based quantum information development, especially the chip-based QKD research. All these background works provide a

framework and inspiration for this entire research project. This review lays the technological foundation of this thesis.

In Chapter 3, the passive components are designed, simulated, fabricated and experimentally tested in detail. The design of the fiber-chip coupler is introduced first, including the special two-dimensional grating coupler for polarization multiplexing in QKD. The governing equations for the directional coupler, Mach-Zehnder interferometer, and ring resonator are derived. Then the multimode interferometer for power splitting and the polarization manipulation devices are analyzed in detail. The final part of this chapter explains the fabrication process development and process flow for the silicon photonic passive components.

In Chapter 4, the active components, including modulators and photodetectors are studied. First, the design of modulators based on three different effects, namely the thermo-optic effect, carrier injection or depletion effect, and electro-optic effect, are theoretically and experimentally studied. Then, the germanium photodiode is designed and tested. It is then used in a balanced homodyne detector for the receiver in continuous variable QKD. Finally, the fabrication process flow for the silicon-germanium active components is explained and discussed.

In Chapter 5, an innovative single-chip solution for continuous variable QKD is demonstrated for the first time, based on the previously developed components. Two chips, to serve as the transmitter and receiver, are designed and fabricated, with all the essential components integrated onto the single chip using nano-silicon photonic fabrication processes. In this chapter, quantum optics theory and information theory related to the continuous variable QKD are provided first, including the necessary post-processing protocols. Then the chip based QKD system is demonstrated theoretically and experimentally.

Chapter 6 summarizes the major contributions of this thesis and concludes with the recommendations for future work.

Chapter 2 Literature Survey

The literature survey in this chapter is divided into three parts. The first part focuses on the roadmap of quantum key distribution, with a general survey on the two major categories, i.e., the continuous variable QKD and discrete variables QKD. Then, a review of the current developments in silicon photonics, including both the passive components and active components, is covered. The third part presents a review of the developments in chip-based quantum information, with particular emphasis on chip-based QKD works.

2.1 Literature Survey on Quantum Key Distribution

The development of Quantum key distribution (QKD) starts from the year 1984, when Charles H. Bennet and Gilles Brassard proposed the first quantum key distribution protocol, which is known as the BB84 protocol [4]. Many QKD systems have been demonstrated during the past 30 years. Some loopholes have been closed, the security has been analyzed in detail and protocols have become more accessible to implement. In most QKD protocols, light is used as the medium to exchange quantum information. Depending on the dimension of encoded information, QKD protocols can be categorized into discrete variables QKD (DV-QKD) and continuous variable QKD (CV-QKD). DV-QKD encodes information into the polarization, phase, or time-bin of a photon with a limited dimension of Hilbert space. CV-QKD, on the other hand, encodes information into the quadrature field of the electromagnetic field with an infinite and continuous dimension of Hilbert space. Different from DV-QKD, the information here is a random number with a continuous probability distribution. Both methods use optical fiber or free space as the optical link. In this section, both methods will be reviewed in detail.

2.1.1 Discrete variables QKD

DV-QKD is a relatively mature technology, where a single photon is used as the information carrier. Hence, single photon detectors are a requirement in DV-QKD implementations. The previously mentioned BB84 protocol and most of the practical QKD experiments belong to this category. As shown in Fig. 2-1 [13], in the BB84 protocol, the transmitter named Alice encodes her information randomly in one of the four states, $|0\rangle$, $|1\rangle$, $|+\rangle$, $|-\rangle$, which practically can be the polarization or phase in the

Figure 2-1: Schematic of the BB84 protocol. The keys are encoded in the polarization of photons.

angle of 0 , π , $\pi/2$, $3\pi/2$. The receiver named Bob would measure the information randomly in the $|0,1\rangle$ basis or $|\pm\rangle$ basis. When they use the same basis, they can get the same information, but when they use a different basis, there is a 50% chance that the result is wrong. Thus, both will discard the information when they use a different basis. In this way, they can share a series of secure keys with minimum error.

One variation for BB84 is the entanglement-based protocol proposed by Artur Ekert in 1991, which is known as the E91 protocol [14]. In this protocol, an entanglement source is introduced to generate maximum entangled photon pairs as $\frac{1}{\sqrt{2}}(|\uparrow, \uparrow\rangle + |\downarrow, \downarrow\rangle)$. Each of the two parties, Alice and Bob, measures one of the photons. Alice measures her photon in 0 , $\pi/4$ and $\pi/2$ angle base. Bob measures his photon in $\pi/4$, $\pi/2$, and $3\pi/4$ angle base. Therefore, they are able to establish a secure communication by testing Bell's inequality. However, due to the difficulties in constructing a high-efficiency entanglement source, this protocol is not widely adopted.

To simplify the true single photon preparation process, weak coherent pulse with average photon number < 0.1 per pulse can be used to simulate a single photon source. However, this may open a severe loophole for the photon-number-splitting attack [15]. Because there is always a non-zero probability of getting more than one photon per pulse, the eavesdropper could locate these pulses and extract one photon without being noticed. This led to the development of the decoy states protocol, which have been reported to close this loophole [16, 17]. Alice, in this case, prepares her photon pulses with multiple average photon numbers. The eavesdropper attempts to maintain the overall error rate but inevitably changes the error rate distribution at each different photon number level. Thus, Alice and Bob can detect the eavesdropper by analyzing the error rate at each photon number level. The decoy state method has significantly increased the security level of QKD using weak coherent pulses and quickly become the most adopted protocol. State-of-the-art results have shown a 200 km transmission distance based on this protocol [18, 19].

Another source of practical loopholes is from the detector side. Many attacks are proposed targeting the practical single photon avalanche diode, such as the blinding attack, after-gate attack, super linearity attack and laser damage attack. [20]. To overcome this, device-independent QKD have been proposed. However, it is unpractical due to its requirements of detectors with near-unity efficiency and photon number resolution. In 2012, a measurement-device-independent (MDI) QKD was proposed, as shown in Fig. 2-2a [21]. It removes all the side channels of the detector without sacrificing the transmission distance. Experimental result in 2016 demonstrated an MDI-QKD system with a secure transmission distance of 404 km [22].

Distributed-phase-reference protocols are another type of DV-QKD protocol, in which the information is encoded as the relative phase difference between adjacent

Figure 2-2: Schematic of (a) MDI-QKD protocol [21], (b) COW protocol [26] and (c) the DPS protocol [23].

photon pulses. As shown in Fig. 2-2c, in the differential phase shift (DPS) protocol [23, 24], a single photon is split into $N (\geq 3)$ time slots, while each of them is modulated randomly with 0 or π phase shift. The receiver detects the phase difference between adjacent pulses using an unbalanced MZI. Another protocol is the coherent-one-way (COW) protocol [25], which encodes information in the intensity of adjacent photon pulses (i.e., “01” for bit 0, “10” for bit 1 and “11” for coherence check), as shown in Fig. 2-2b. The security of information is verified by checking the quantum coherence between adjacent bits. State-of-the-art implementations show a 307 km transmission distance for COW protocol [26] and 260 km for DPS protocol [27].

Up to the year 2018, many research groups have shown the maximum achievable transmission distances greater than 100 km fiber length and secure key rates above the Mbps (Megabits per second) level, which is already good enough for a metropolitan area communication scenario. Many countries have started the testing of their own QKD networks. The U.S. agency DARPA began its QKD network testing in 2002, makes it the first quantum network in the world [28]. Europe demonstrated the SECOQC quantum network in 2008. In total, 41 research and industrial organizations are involved. Multiple protocols are tested in the network, including BB84 with decoy state, COW, entanglement-based protocols and CV protocols [29]. In 2011, Japan started its QKD network construction with six nodes in Tokyo, which has so far demonstrated a GHz-clocked QKD system and a QKD smartphone, providing a secure communication channel for government and important infrastructures [30]. China deployed a QKD network between the Beijing and Shanghai in 2013. It is based on 32 trust nodes, and its total length is over 2000 km. The secure key rate over the whole length is more than 20 kbps. In 2014, the UK also started to investigate the Quantum Communication Hub to develop the first QKD network in the UK. Chip-scale integration is also one of their targets.

2.1.2 Continuous variable QKD

CV-QKD uses quadratures of the electromagnetic field to encode information. It has gained much attraction due to the much simpler measurement techniques required. To detect the quadrature of light, only a homodyne detector built using the classical photodiode is required. There are three types of CV-QKD, which are achieved by using the Einstein-Podolsky-Rosen (EPR) state [31], squeezed state [32], or the coherent state [33]. The coherent state protocol is the most popular one due to the simplicity of the

source, where only a standard telecom laser is required. Thus, coherent state CV-QKD protocol is fully compatible with classical optical communications equipment. Through the use of the advanced equipment, the cost, stability and reliability of CV-QKD become significantly better than those of DV-QKD.

The first coherent state CV-QKD protocol, known as the GG02 protocol, was reported by Frédéric Grosshans and Philippe Grangier in 2002. In the GG02 protocol, the coherent state is Gaussian-modulated in both the x and p quadrature at the sender side. Like the BB84 protocol, the receiver then chooses to measure one of the quadratures randomly. The first GG02 protocol was based on the forward-reconciliation scheme, which theoretically limits the transmission loss to 3 dB, or about 15 km distance. Then in 2003, the limit was broken by using the reverse reconciliation scheme in the protocol [34]. The proof of principle test produced a 75-kbps secure key rate at 3.1 dB loss. Another type of variation is the use of heterodyne detection, which is the so-called No-Switching scheme [35]. In this protocol, there is no need for the rapid basis switching. Both quadratures are measured at the same time. However, due to the Heisenberg uncertainty principle, additional shot noise is added to the measurement result. Thus, the performance for the No-Switching protocol remains the same as the original protocol.

In 2007, a much more practical system was constructed in France [36]. A 25-km length of optical fiber communication channel is used in this system. To reduce the phase noise generated in the fiber, the signal and local oscillator light are multiplexed both in time and polarization for fiber transmission. The final secure key rate was 2 kbps. At the same time, another all-fiber system is demonstrated in Canada with a similar strategy. The signal and local oscillator light are multiplexed both in time and frequency. A pair of acousto-optic modulator is used to upshift and downshift the

frequency. The final secure key rate is 0.3 bit/pulse in a 5 km fiber can operate stably with real-time error correction and privacy amplification [37].

Later in 2013, the group from Telecom Paris Tech, using a high-efficiency error-correction procedure, achieved a transmission distance of over 80 km, where they could provide 0.8 kHz secure key [38]. The previous results were limited by the low efficiency of error-correction at a low signal-to-noise ratio. In this work, a new type of multidimensional reconciliation protocol was introduced with significant improvement in the low signal-to-noise ratio scenario. As a result, the secure transmission distance was significantly boosted.

In the previously developed CV-QKD systems, the signal light and local oscillator are multiplexed in the same fiber. Due to the inevitable channel loss, the power of the local oscillator light must be high enough to ensure a good signal-to-noise ratio. Crosstalk in the fiber would be a big issue in this case. Moreover, the eavesdropper could attack the system by manipulating the local oscillator. In 2015, a new protocol suggested that the local oscillator light could be generated “locally” at the receiver side [39-41]. This protocol could significantly improve the security of CV-QKD.

In 2017, a group from Beijing University of Posts and Telecommunications demonstrated their CV-QKD system in a 50 km commercial fiber, with 8.7 kHz secure key rate. The 24-hour continuous test moved the CV-QKD towards a more practical secure network [42]. Later in 2018, a group from Huawei, Germany, demonstrated a CV-QKD experiment using the existing wavelength division multiplexed (WDM) optical communication network [43]. The secure key rate could reach 90 kbps over 20 km fiber together with 10×10 Gb/s WDM signal. The result proved the compatibility of CV-QKD with current telecom infrastructure.

2.2 Literature Survey on Silicon Photonics

The field of silicon photonics was first developed in the 1980s, when silicon was investigated as a suitable waveguide material for 1.3 to 1.6 μm wavelengths [44, 45]. A few years later, the study of silicon-on-insulator (SOI) wafer provided new potential for silicon photonics [46, 47], which started a new era for integrated optics. Various critical optical components have been successfully integrated on silicon photonic platform, including power splitters, multiplexers and demultiplexers, modulators, photodetectors and lasers. Many of the advantages of silicon in semiconductor information technologies are now showing their potential in integrated photonics as well. Silicon is abundant in the earth, has excellent mechanical, chemical and thermal stability, can be made ultrapure and most importantly, is fully compatible with the specially designed equipment for semiconductor technologies. Moreover, electronics and photonics can be integrated together on the same chip to offer a truly integrated photonic system [5]. All these advantages are leading to a silicon photonics chip with compact size, low cost, high stability, high reliability and simple fabrication.

Although silicon is a strong candidate for integrated photonics, many other fabrication platforms have been explored for chip-level integration together with the development of silicon. The III-V group compounds [48], lithium niobate (LiNbO_3) [49], and potassium titanyl phosphate (KTP) [50] are the common examples for nonlinear optics platforms. The high nonlinearity can be used to build on-chip lasers and frequency converters. High-speed modulation up to the GHz level is also achievable. However, the fabrication processes for these materials are highly specialized and too expensive for mass production. Another group is based on low index contrast materials such as silicon nitride [51], silica [52] and femtosecond laser

writing on glass [53, 54] techniques. This group typically offers low transmission loss down to 0.5 dB/cm and coupling loss lower than 1 dB, which is suitable for a low loss delay line and some other loss-sensitive applications. However, rapid modulation techniques even at MHz, are lacking in this group of platforms. Because of the low index contrast, the size of components is much larger than those in silicon, limiting the scalability and complexity of the final chip. Silicon, on the other hand, is a more balanced platform. Although the transmission loss is around 3 dB/cm, small devices at the micrometer scale still offer acceptable loss per chip. Modulators and detectors operating at the GHz level are also available. The main limitation is the lack of a high-efficiency on-chip laser, which can be solved by the laser die bonding technique.

In the following part of this section, the developments of the passive components, and active components are reviewed separately.

2.2.1 Passive components

Passive components are the class of fundamental components which cannot introduce energy into the system, including the fiber-chip coupler, power splitter, polarization manipulation components, filters, interferometers and resonators. Those related to the QKD work will be reviewed in detail below.

2.2.1.1 Fiber-chip coupler

Coupling loss due to large mode field mismatch between a sub-micron size waveguide and a single mode fiber with a mode-field diameter of about 10 μm , represents one of the major losses in a photonic system. If no special component is used to assist coupling, the average coupling loss is at the order of 20-30 dB. Thus, for a practical device, a mode converter between fiber and waveguide is a necessity. There are two commonly used coupling methods, the taper coupler and the grating coupler.

Based on the direction of the input light beam, these two methods are also known as side coupling and vertical coupling, respectively.

For the taper coupler, the cross-sectional area of the waveguide is changed, either shrinking (Fig. 2-3a) [55] or expanding (Fig. 2-3b) [56], to match the much larger mode field diameter of the fiber. If the tapered section is long enough, the conversion process is adiabatic and theoretically provides 100% efficiency. Limited by the fabrication process, the thickness of the waveguide is difficult to change and the only way to change the cross-section in practice is by changing the width of the waveguide. In this case, shrinking the width of the waveguide may provide a much better mode overlap in both lateral and vertical directions. One of the early results used a waveguide tip with width of 120 nm to achieve 0.5 dB coupling to a low mode field diameter (4.9 μm) fiber [57]. However, the use of uniquely designed fiber and high accuracy fabrication technology limited the practical use of this method. Later, a modified approach added another material overlay, such as a polymer (Fig. 2-3c) [58, 59], silicon oxynitride [60] or silicon dioxide (Fig. 2-3d) [61], covering the waveguide tip and acting as an additional stage for mode conversion. Direct coupling with <1dB coupling loss to the single mode fiber is possible with this approach. By combining the two methods, the input fiber was changed back to a lensed fiber with a mode field diameter of 2.5 μm , achieving a minimum 0.36 dB coupling loss [58, 62-64].

Grating couplers utilize a grating structure to achieve phase-matching between the input light and the waveguide guided mode by one of its orders of diffraction [65]. Their advantage is that they can be put anywhere on-chip, which makes in-line testing possible. One of the pioneering works in the early 2000s reported a grating coupler on SOI waveguide to provide a 74% simulated efficiency and 14% measured efficiency (Fig. 2-4a) [66]. A crucial source of coupling loss for the grating coupler is the

Figure 2-3: Taper coupler structures using (a) Shrinking taper, (b) Expanding taper, (c) Taper with polymer overlay and (d) Taper with SiO₂ overlay. Reproduced from Fig. 1 of Ref. [55], Fig. 2 of Ref. [56], Fig. 2 of Ref. [59] and Fig. 1 of Ref. [61].

diffraction of light in unwanted directions, which is characterized by the so-called directivity of the grating. The diffraction could be optimized by changing the shape of the grating to cancel the unwanted diffraction by destructive interference (Fig. 2-4b) [67], which gives a 1.6 dB tested coupling efficiency. Another way to achieve this is by adding a bottom reflection layer, such as Bragg grating [68] and metal (Fig. 2-4c) [69], to recycle the portion of light diffracted towards the substrate. A coupling efficiency of 69% is achieved in both the reflection methods. Further improvements utilize a non-uniform grating to match the Gaussian shape of the input light. Without the reflection layer, the coupling efficiency can reach 61% theoretically [70] and 34% experimentally (Fig. 2-4d) [71].

Figure 2-4: Grating coupler structures using (a) The original design, (b) Interference enhanced design, (c) Metal reflection layer enhanced design and (d) Non-uniform design. Reproduced from Fig. 1 of Ref. [66], Fig. 1 of Ref. [67], Fig. 7 of Ref. [69] and Fig.1 of Ref. [71].

2.2.1.2 Polarization manipulation

Due to the severe birefringence of silicon waveguides, silicon photonic chips are usually acutely sensitive to the polarization of input light. A general solution is implementing polarization diversity schemes [72, 73], which consist of a pair of polarization beam splitters (PBS) and polarization rotators, as shown in Fig. 2-5. With the polarization manipulation techniques available on-chip, light can be manipulated in one more degree of freedom, which is essential in some applications such as

Figure 2-5: Schematic of polarization diversity scheme. Reproduced from Fig. 1 of Ref. [73],

polarization-division multiplexing [74] and high-dimensional quantum information [75].

Chip-based PBSs have been reported using many different structures, such as Mach-Zehnder interferometers (MZI) (Fig. 2-6a) [76, 77], multimode interference (MMI) devices (Fig. 2-6b) [78, 79], photonic crystals (Fig. 2-6c) [80, 81] and directional couplers (Fig. 2-6d) [82-85]. The PBS based on MZI and, MMI tend to have an enormous size ($> 1000 \mu\text{m}$) even with SOI nano-waveguide, which is not practical in large-scale integrated photonic circuits. Photonic crystals and directional couplers could enable a compact size PBS, but they are sensitive to the fabrication errors. However, with the development of nano-fabrication technology, the effects of fabrication error gradually decrease, making them ideal for on-chip PBSs. There is another special case when the PBS is used at the input and output port interfaces, where a two-dimensional grating coupler can be used for both efficient light coupling and

Figure 2-6: Polarization beam splitter design based on (a) MZI structure, (b) MMI structure, (c) photonic crystal and (d) directional coupler. Reproduced from Fig. 1 of Ref. [77], Fig. 1 of Ref. [78], Fig. 1 of Ref. [80] and Fig.3 of Ref. [83].

polarization splitting together [86-88]. The coupling efficiency can reach 48%, while the polarization extinction ratio can reach 24 dB.

The polarization rotator is more difficult to fabricate on the waveguide platform because the waveguide usually has great polarization maintenance performance. Special asymmetrical coupling is introduced, such as off-axis core (Fig. 2-7a) [72], bilevel taper (Fig. 2-7b) [89], slanted angle (Fig. 2-7c) [90], corner cut waveguide (Fig. 2-7d) [91], triangular waveguide (Fig. 2-7e) [92], and asymmetrical directional coupler (Fig. 2-7f) [93, 94]. All the methods mentioned above, except directional coupler, requires at least one additional fabrication step, which would increase the complexity of the fabrication process and reduce yield. The directional coupler, on the other hand,

Figure 2-7: Polarization rotator designs based on (a) off-axis core, (b) bilevel taper, (c) slanted angle, (d) corner cut waveguide, (e) triangular waveguide, (f) asymmetrical directional coupler. Reproduced from Fig. 3 of Ref. [72], Fig. 1 of Ref. [89], Fig. 1 of Ref. [90], Fig. 1 of Ref. [91], Fig. 1 of Ref. [92] and Fig.1 of Ref. [94].

requires only one etching step, just like as the original waveguide structures. The extinction ratio in this case can reach > 20 dB.

2.2.2 Active components

Modulators and detectors are two major categories of active components, and they are the core components for optical-electrical-optical conversion in an optical

communication repeater. Currently, in a datacenter or communication network, the limitation for overall speed is the cost and power consumption per data bit, which leads to the requirements for modulators and detectors to be high speed, low voltage drivable (< 2 V), and high fabrication yield. The laser is another very important active component and it is possible to be integrated on-chip. However, it is beyond the scope of this project. There are a few possible ways to integrate laser on-chip. A fully integrated process could achieve by bandgap engineering of Ge, such as doping, induce strain and using Ge-Sn alloy, which could enhance the emitting efficiency but also increase the process difficulty. Some other material such as Er-related and III-V based material offers higher emission efficiency, but they are not compatible with CMOS fabrication process [95]. Another compromised plan is to bond the laser die to the silicon chip using wafer-to-wafer bonding or die bonding techniques after the silicon photonic fabrication is finished [96, 97]. This is also one of our future plans for a fully integrated miniaturized QKD chip. In this section, the current developments of silicon photonics modulators and detectors are reviewed.

2.2.2.1 Modulators and Switches

Silicon is not an ideal platform for active tuning because it lacks the electro-optic ($\chi^{(2)}$) effect commonly used in some other traditional optic materials [98]. Modulators based on the electro-optic effect can operate at up to 40 Gbps. Among them, aluminium nitride (AlN) is one emerging CMOS-compatible material with relatively strong electro-optic effect [99]. Modulation based on a ring resonator at 4.5 GHz has been demonstrated [100]. Although in silicon, there are some methods to induce $\chi^{(2)}$ nonlinearity such as adding a strong electric field [101] or a lateral strain [102], they are not stable enough for a practical device. Unique modulation methods must be introduced in silicon. The thermo-optic effect is one option, due to the large thermo-

optic coefficient of silicon (1.8×10^{-4} RIU/K) [103]. Based on Mach-Zehnder Interferometer (MZI) structure [104, 105] or ring resonator structure [106, 107], very compact and low loss thermo-optic modulators have been developed. An optical micro-electro-mechanical system (MEMS) device is another option for the modulator. The deformation of the waveguide could change the coupling to an adjacent waveguide [108] or change the phase of light [109]. However, both the heat and MEMS methods lack high-speed modulation capability at the MHz level. The most promising technique for fast modulation on silicon is the use of the plasma dispersion effect [110], where the refractive index is changed by free-carrier concentration. Due to the weakness of the effect, the first GHz device based on an MZI structure has up to 10 mm length [111]. By changing to a high Q-factor ring resonator, the device footprint could be reduced to micrometer level [112], while maintaining the GHz operation. The device here works at carrier injection mode, which is limited by the carrier recombination time and inherently slower than the carrier depletion or accumulation mode [113]. In recent years, the operation speed has been pushed to 50 Gbps and beyond [114-116]. The progress of ultra-fast modulators has compensated this weak point of silicon and proven the potential of silicon as a photonic platform.

2.2.2.2 Photodetectors

A good material for transmission is a bad material for the photodetector, because photodetection requires absorbing photons and generating electron-hole pairs. One way to solve this is by damaging the silicon lattice with ion doping to induce absorption at the desired 1550 nm wavelength [117-120]. However, the responsivity is not good enough (0.1 – 0.8 A/W), and high reverse bias voltage (> 10 V) is required. Hence, germanium and III-V group materials can be a good supplement to silicon. Wafer or die bonding techniques can be used to form a hybrid integration chip [121-123]. To

simplify the fabrication process and reduce cost, foundries still prefer using a fully CMOS compatible scheme. Thus, the hetero-epitaxy germanium growth on top of silicon is attracting increasing attention, despite the difficulties in the fabrication process. Due to the large lattice mismatch between germanium and silicon (4.2%), many efforts have been made in the development of a high-quality germanium layer. Several tens of nm of silicon-germanium [124, 125] or low-temperature germanium [126, 127] buffer layer are used to relax the stress. Today, the germanium detector is one of the most mature devices on the silicon photonic platform. The state-of-the-art specifications are: responsivity of 1 A/W at 1550 nm, bandwidth > 60 GHz and dark current $< 1 \mu\text{A}$ [128-132].

2.3 Literature Survey on Chip-based Quantum Information

Due to the requirements for small size, low cost, high stability, high reliability and the potential for mass production, integrated photonic technology provides an unprecedented opportunity for next-generation quantum information [133-135]. It is possible to have hundreds to thousands of waveguides on a single chip, which could form a vast network impossible with traditional platforms.

Early demonstrations used silica waveguides to achieve various applications. In 2008, a group in University of Bristol reported chip-based two-photon quantum interference with 94.8% visibility and controlled-NOT gate with the fidelity of 94.3% [52]. Later in 2010, the same group demonstrated the first chip-based quantum walk in 21 optical modes [136]. The result showed a 76-standard deviation violation over the classical limit. In 2011, a reconfigurable photonic circuit was reported for generating, manipulating and measuring entanglement and mixture states with 8 phase shifters [137]. That same year, another group showed that the integrated photonic chip is also

capable of manipulating polarization qubits [138]. Then in 2013, four groups realized chip-based boson sampling simultaneously [139-142]. They demonstrated chips with five to six waveguide modes and two to three photons inputs. Due to the computational complexity of the boson sampling problem, these results opened a new way to demonstrate quantum supremacy using integrated photonics. In 2014, a quantum teleportation chip was demonstrated [143], showing a fidelity of 89%. In 2015, a universal quantum computer based on linear optics was published [144]. The chip consists of 15 tunable MZIs across six waveguide modes, which can be tuned to achieve an arbitrary 6-dimensional unitary transformation. Continuous variable entanglement on-chip was also demonstrated in 2015 [145]. However, due to the relatively low refractive index contrast for the silica waveguide, the fabricated chip had a huge size, which limited the scalability of such a chip. Moreover, the lack of high-speed modulation and single photon source also limited the development of silica chips.

Silicon waveguides has a much higher refractive index contrast ($\Delta n = 2$) than doped silica ($\Delta n = 0.02$), leading to a 1000 times smaller waveguide bending radius [146, 147]. A group from Massachusetts Institute of Technology demonstrated a silicon-based reconfigurable photonic processor with 88 MZIs, where the total chip size is about 5 mm [148]. As a comparison, the previous work on silica chip has only 15 MZIs on a chip with a size of several centimeters. Moreover, the photon source and squeezed state generator mainly rely on the second order ($\chi^{(2)}$) and third order ($\chi^{(3)}$) nonlinear conversion of light. Silicon, being a centrosymmetric crystal, has no second-order nonlinearity naturally, but its high third order ($\chi^{(3)}$) nonlinearity can be used for the efficient generation of single photon states [149] or possibly squeezed state. The first photon pair generation experiment in a silicon waveguide was reported in 2006 [150]. Then silicon-based photon sources were demonstrated using various structures,

such as a single ring resonator [151-153], PIN junction assisted ring resonator [154, 155], coupled-resonator optical waveguides (CROW) [156], high Q-factor micro-disks [157], photonic crystal slow light waveguide [158], photonic crystal coupled-resonator [159], and one-dimensional photonic crystal cavity [160]. State-of-the-art photon sources can generate photons at MHz speed with about 1 mW coupled pump power. Regarding squeezed state sources, some work on silicon nitride [161, 162], periodically poled potassium titanyl phosphate (PPKTP) [163], and lithium niobate [164] waveguides have been demonstrated. However, there is no direct implementation of squeezed state on the silicon waveguide due to the strong nonlinear losses such as two-photon absorption and free-carrier absorption [162]. The nonlinear two-photon absorption is one of the biggest problems limiting the performance of silicon in quantum optics. In high-power nonlinear optics, the two-photon absorption effect may create parasitic free carriers, together with thermal effects further deteriorating the transmission efficiency.

By combining the photon source and the large circuit on silicon, more and more complex functions can be achieved. Experimentally, this began with on-chip quantum interference experiments in 2014 [153], in which a chip-based photon pair source demonstrated a Hong–Ou–Mandel (HOM) interference visibility of 95%. In 2016, a chip-to-chip quantum photonic interconnect was reported [165], which provided a new way for path-to-polarization conversion and entanglement distribution between different chips. In 2017, a photonic chip was designed for simulating a real Hamiltonian from diamond nitrogen system [166]. The uncertainty of approximation reached 10^{-5} . Later in 2018, the same group demonstrated a new chip for 15-dimensional quantum entanglements [167]. The chip has over 550 photonic components, which provide full tunability over the 15 dimensions. The same year, they published another large-scale

photonic chip for arbitrary two-qubit processing [168]. More than 200 photonic components are integrated onto the same chip, and the average quantum process fidelity is about 93.2%.

QKD, as a particular quantum information application, could perfectly adapt to the well-developed silicon quantum photonic platforms. The early works on miniaturizing QKD devices started with the compact design of optical components [169-171]. Then in 2014, a client-server QKD scheme based on the reference-frame-independent QKD protocol was demonstrated [49]. The laser and detectors are contained inside the server while the client integrates only the on-chip polarization rotator. In 2015, a hand-held QKD transmitter device was reported [172]. In this work, four vertical cavity surface emitting lasers (VCSEL) and four wire-grid polarizers were used to generate the required BB84 states. Then, a femtosecond laser written waveguide is used to erase the path information. Later in 2015, a silicon nitride chip for time-bin entanglement generation was published, in which the two-photon interference fringe is measured to have 88.4% visibility [51]. In 2016, the group from the University of Bristol demonstrated their work on chip-based QKD, which is by far the most comprehensive chip-based QKD system [173]. As shown in Fig. 2-8a, the transmitter chip is fabricated on the indium phosphide (InP) platform, which integrated the laser and fast modulators on the chip. The receiver chip uses silicon oxynitride (SiO_xN_y) to provide a low-loss delay line with 0.5 dB/cm loss. The chip was tested for three different protocols (time-bin encoded BB84, COW, and DPS) to show the reconfigurability, and the maximum secure key rates were 345 kbps, 311 kbps and 565 kbps for the BB84, COW and DPS protocols, respectively. However, research utilizing silicon platforms has not stopped. Later in 2016, a silicon chip designed for polarization-encoded BB84 transmitter was demonstrated (Fig. 2-8b) [8]. The proof-

Figure 2-8: State-of-the-art chip based QKD systems. (a) InP and SiON system for three protocols. (b) Silicon transmitter for BB84 protocol. (c) Silicon transmitter for three protocols. (d) Silicon chip for high-dimensional QKD. Reproduced from Fig. 1 of Ref. [173], Fig. 2 of Ref. [8], Fig. 1 of Ref. [7] and Fig.1 of Ref. [6].

of-principle test shows a secure key rate of 0.95 kbps at 5 km of fiber distance. In 2017, another silicon transmitter was designed for time-bin encoded BB84, polarization encoded BB84 and COW protocol (Fig. 2-8c) [7]. The results showed a simulated secure key rate of 329 kbps for polarization encoded BB84 and 916 kbps for the COW, respectively. Also in 2017, another group demonstrated a high-dimensional QKD chip

based on multicore fiber (Fig. 2-8d) [6], which could provide more than 1 bit/photon and drastically increase the secure key rate.

2.4 Summary

This chapter provided an overview of the reported research work on silicon photonics, QKD, and the new works that combine the two fields. In Section 2.1, the roadmap of QKD, including DV-QKD and CV-QKD were reviewed in detail. Different generations of protocols targeting different practical problems were introduced. Then, the history and various components of silicon photonics were reviewed in Section 2.2, including the fiber-chip coupler, polarization manipulation devices, modulators and photodetectors. Finally, in Section 2.3, some recent developments of chip-based quantum information were reviewed. The previous development proved the feasibility of using silicon photonics for difficult quantum information processing jobs. Furthermore, this review has shown the limitations of the previously developed QKD chips. Thus it leaves plenty of space for further exploration of various QKD applications on-chip and for the improvements in the performance.

Chapter 3 Silicon Photonic Passive Components

Passive components are the devices that incapable of controlling the light using another electronic signal, which are the underlying devices to construct a quantum integrated photonic chip, includes the input & output interface, transmission medium and manipulation components. All the components are based on the waveguide structures, which uses a higher refractive index material as core and a lower refractive index material as cladding. In this chapter, the fundamental working principle of the waveguide is introduced first. Followed by the design and simulation of fiber-chip coupling interface. Then the light manipulation components, including the directional coupler, multimode interferometers, polarization manipulation devices, Mach-Zehnder interferometers and ring resonators, are designed, simulated, fabricated and tested. The final part of this chapter explained the fabrication process for all the silicon photonic passive components.

3.1 Fiber-chip Coupler

Fiber-chip coupler act as the optical interface between the photonic chip and the connecting optical fiber. Categorized by the relative angle between the input light beam and the waveguide, there are two types of coupling techniques, lateral coupling and vertical coupling. In this section, the most commonly used technique for both lateral coupling and vertical coupling are studied in detail.

3.1.1 Taper coupler

Taper coupler is a kind of lateral coupler that couple light from the fiber to the end of the waveguide with optical axis aligned. To get an efficient coupling, the mode size of waveguide and fiber are carefully matched. A typical single mode fiber has about 10- μm mode field diameters, while the waveguide has about 0.5- μm mode field diameter. The fiber mode is reduced to about 2.5 μm by using ultra-high NA fiber or lensed fiber. The waveguide mode is expanded by introducing a taper structure at the end, as shown in Fig. 3-1a. The width of the waveguide will reduce from w to w_t . The orange curve illustrates the mode profile at the two ends of taper. While decreasing the waveguide width, the mode size increases. By carefully optimize the value of w_t , mode matching between fiber and waveguide could achieve. The effective refractive index and power coupling are calculated with waveguide tip width w_t from 80 nm to 240 nm, as shown in Fig. 3-1b and Fig. 3-1c. The power coupling measures the amount of power that can couple from one mode into a forward propagating wave with another mode profile. The fiber mode is a Gaussian distribution with 2.5- μm beam width. The result shows that for TE mode coupling, a waveguide width of 160 nm is optimum, while the 100-nm width waveguide is best for TM mode coupling. However, restricted by the

Figure 3-1: (a) Schematic of taper coupler structure. (b) Effective refractive index and fiber to chip power coupling efficiency of waveguide with different width w_t .

Figure 3-2: Simulated TE and TM mode electric field distribution in waveguide with different width.

minimum linewidth of the lithography process, the waveguide width of 200 nm is chosen, which corresponded to a coupling loss of 3 dB. The TE and TM mode profile for 160-nm, 200-nm and 450-nm waveguide width are shown in Fig. 3-2. 160-nm width is the optimum coupling width for TE mode, for which the electric field is expanding far beyond the waveguide region. 200-nm is the actual coupling width in our experiment. 450-nm is optimized for single mode transmission, which shows much better light confinement inside the waveguide region compares to other narrower designs. For TE mode, most of the evanescent field is at the vertical edges, while for TM mode, most of the evanescent field is at the top and bottom edges.

To reduce the coupling loss between the 200-nm tip and 450-nm single mode waveguide, the taper length should be sufficiently adiabatic or in another word long enough. Thus, the total transmission from the 200-nm tip and 450-nm waveguide is simulated as a function of taper length L . As shown in Fig. 3-3a, the transmission

Figure 3-3: (a) Waveguide taper transmission as a function of taper length for TE mode and TM mode. (b) SEM photo of a waveguide tip with silicon dioxide hard mask on top.

increases while the taper length increases. At $L = 10 \mu\text{m}$, the loss is less than 0.5% for TE mode and less than 0.15% for TM mode. At $L = 100 \mu\text{m}$, the loss is less than 0.35% for TE mode and less than 0.13% for TM mode. In our design, $L = 100 \mu\text{m}$ is chosen.

A scanning electron microscopic photo of the waveguide tip is shown in Fig. 3-3b. The waveguide structure is made by silicon with a cross-section of 200×220 nm. There is a 70-nm thickness SiO_2 on top of the silicon structure, which is used as the hard mask for silicon etching. In Fig. 3-3b, the SiO_2 layer is false-colored in blue. The tip shows a round shape because of the limitation of lithography resolution.

3.1.2 1D grating coupler

Different from the taper coupler, the grating coupler is a vertical coupling component that could couple light with a large angle between fiber and waveguide. 1D denote the periodic grating structure distribute only in one dimension (1D), as compared to the 2D grating coupler in the next section. As shown in Fig. 3-4, the input light comes from a nearly vertical direction with a small angle θ . With the help of the grating coupler, the light can end up transmitting in the horizontal direction. It is necessary for output from the grating coupler to be phase-matched with the targeting mode inside the waveguide with $k_z = kn_3 \sin \theta = \beta$. Since the refractive index of waveguide n_1 is larger than the cladding material n_3 , the phase-matching condition could never achieve without any auxiliary structures. A grating coupler with proper period could reduce the propagation constant inside the waveguide by [174]

$$\beta_g = \beta + \frac{2\pi m}{\Lambda}, \quad (3-1)$$

where β_g is a series of propagation constants available from the grating structure. Λ is the period of the grating. m is an integer, which is called the diffraction order of the grating. Usually the first diffraction order with $m = -1$ is used for phase match. Thus, the phase-matching condition can be expressed as

$$k_z = \beta_g = kn_3 \sin \theta = \beta - \frac{2\pi}{\Lambda}. \quad (3-2)$$

Figure 3-4: Basic working principle of a grating coupler. The input light has a small tilted angle θ .

By solving Eq. (3-10), the grating period is given by

$$\Lambda = \frac{\lambda}{n_{eff} - n_3 \sin \theta} . \quad (3-3)$$

The grating coupler is fabricated by etching the waveguide surface with small trenches. The trenches have a period of Λ , empty ratio of E/Λ and depth of D . Since the trenches will slightly affect the propagation mode of waveguide, the three parameters require careful fine tuning to maintain the maximum coupling efficiency. As a result, the grating period value directly calculated from Eq. (3-11) could only act as a reference. Detailed analysis of the effect of etching must be performed.

The coupling efficiency is simulated by Lumerical FDTD Solutions software. A Gaussian beam with 10.4- μm beam width is used to simulate the single mode fiber output. The tilting angle θ of fiber is chosen as 8 degrees, which is a widely used polish angle for fiber communication devices. The waveguide structure is silicon and cladding material is SiO_2 . The top cladding above waveguide has a thickness of 1 μm . A detector is placed at the end of the waveguide output to calculate the coupling efficiency. Three

Figure 3-5: Simulated coupling efficiency as a function of grating period and etching depth.

main parameters including the period Λ , empty ratio of E/Λ and etching remain R is optimized. First, the empty ratio of E/Λ is fixed at 0.5. The result is shown in Fig. 3-5. A global maximum at $R = 110$ nm and $\Lambda = 636$ nm are observed. The 90% tolerance range for the period is 35 nm, and for etching depth is 25 nm. It means that in order to get a coupling efficiency difference of less than 10% for multiple devices, the fabrication error for the period has to be less than 35 nm and error for etching depth has to be less than 25 nm. The non-consistency of optimum grating period value at different etching depth also confirmed the effect of etching on the waveguide mode. By fixing $R = 110$ nm and $\Lambda = 636$ nm, the effect of empty ratio of E/Λ is also calculated. As shown in Fig. 3-6a, maximum coupling efficiency occurs at about $E/\Lambda = 0.48$. The 90% tolerance range for empty ratio is 0.38 to 0.54, which correspond to the trench width of $E = 242 \sim 343$ nm.

Figure 3-6: (a) Coupling efficiency as a function of empty ratio. (b) Coupling efficiency as a function of beam position.

The position of the beam can also affect the coupling efficiency, as shown in Fig. 3-6b. The coordinates of the grating is defined in Fig. 3-4, where the origin is at the edge of the grating. The coordinate of the beam is defined at the center of the beam. The result shows the maximum coupling is at about 3.2- μm away from the grating edge. The 90% tolerance range is 2.5 μm . Thus, the moving resolution for fiber alignment stage should be lower than 2.5 μm to get an extra loss of less than 10%.

Figure 3-7: (a) Electric field distribution for the grating coupler structure. (b) (c) SEM photos of two grating coupler designs.

The electric field distribution of an optimized grating coupler is shown in Fig. 3-7 (a). The primary coupling loss comes from the incomplete mode coupling. A small portion of the light is diffracted towards the wafer substrate. One solution is to add a reflecting material under the waveguide structure to “recycle” the diffracted light, such as a metal layer or Bragg grating layer. In this way, coupling loss down to 1 dB can be achieved. Two SEM photos of the fabricated grating coupler are shown in Fig. 3-7 b-c. One with a curved structure and one with a straight structure. Both work well in our experiments.

3.1.3 2D grating coupler

Similar to the 1D grating coupler, the two-dimensional (2D) grating coupler contains a periodic structure in two dimensions. As shown in Fig. 3-8a, two identical

Figure 3-8: (a) Schematic of a 2D grating coupler. (b) SEM photo of fabricated 2D grating. (c) Basic working principle for 2D grating coupler.

1D grating couplers overlap orthogonally with each other and form a matrix of small pillars. Both the 1D grating couplers could work independently. Since the TE and TM mode have a different effective refractive index, phase matching condition cannot achieve for TE and TM mode together on the same grating coupler. This property can be used as a polarizer and select only one polarization state from incoming light. When incoming light contain orthogonal polarization components (red and yellow arrows), the orthogonal components can be separated into different paths. Both the paths still maintain TE mode. The 2D grating acts as a polarization beam splitter in this case. It can also work as a polarization beam combiner also. When two TE mode light comes from different paths, even though both lights are TE polarized, the output from the 2D grating contains orthogonal polarization components.

The basic working principle of the 2D grating coupler is shown in Fig. 3-8c. Similar to the 1D grating coupler, the input light is tilted with angle θ from vertical direction. The input plane (yellow color) has a 45-degree angle with x - z and y - z plane. The projected propagation constant along two output direction is

$$k_{z1} = k_{z2} = kn_3 \sin \theta \cos 45^\circ. \quad (3-4)$$

Since the 2D grating coupler requires a full 3D-FDTD simulation, which is very time consuming, the 2D grating coupler can be simulated as two independent 1D grating couplers to simplify the calculation with 2D-FDTD. Then 3D-FDTD is used only to make some fine-tuning. To project the 2D grating coupler into the 1D grating coupler, an effective angle α is introduced as

$$k_{z1} = k_{z2} = kn_3 \sin \alpha. \quad (3-5)$$

The effective angle is used as the input angle in the 2D-FDTD simulation. For the standard polished fiber with $\theta = 8^\circ$, the effective angle $\alpha = 5.647^\circ$. Fixing the empty ratio of E/Λ at 0.5, the effect for tuning etching remain R and period Λ is investigated. As shown in Fig. 3-9a, the maximum coupling efficiency occurs at $R = 116$ nm and $\Lambda = 620$ nm, which is slightly different than the 1D case. The 90% tolerance range is 25 nm for Λ and 20 nm for R . Fig. 3-9b shows the effect of empty ratio E/Λ . Maximum empty ratio occurs at $E/\Lambda = 0.455$ and the 90% tolerance range is 0.335 to 0.515, which correspond to a tolerance for trench width of $E = 208 \sim 319$ nm.

With the result from the 1D grating coupler, a full 3D-FDTD simulation is performed. The input light follows the direction in Fig. 3-8 with an 8° angle tilt and the light is linearly polarized in y -axis. The result is showing in Fig. 3-10. The top part is the side view and bottom part is the top view. The yellow circle marked the position of input light. From the result, it is clear to see that although same grating structure is

Figure 3-9: (a) Simulated coupling efficiency as a function of grating period and etching depth. (b) Coupling efficiency as a function of empty ratio.

possessed in horizontal and vertical direction, the light can only couple to the horizontal direction.

Figure 3-10: 3D-FDTD simulation result of electric field distribution for 2D grating coupler. (a) side view (b) top view.

To verify the performance of the 2D grating coupler working as a polarization beam combiner, the devices are fabricated and tested with experimental setup shown in Fig. 3-11a. A 1550-nm continuous wave (CW) laser (Santec TSL-510) is used as the light source. The light source first coupled into the chip via a 1D grating coupler (1D-GC). The laser is then split into two paths via a 50:50 beam splitter (BS). One of the paths is amplitude modulated at 7 MHz frequency using a function generator (Philips PM5193). Then the two paths are combined by the 2D grating coupler (2D-GC). The combined light is separated by an off-chip polarization beam splitter (PBS) (Thorlabs PBC1550SM-FC). The output from PBS is recorded by two photodetectors (MenloSystems FPD-510). The result is shown in Fig. 3-11b. Theoretically, after demultiplexing the light by another PBS, the 7 MHz frequency can only be measured in one of the PBS output. In reality, there is a 35-dB difference between the two

Figure 3-11: (a) Experimental setup for the testing of 2D grating coupler performance. (b) Performance result for 2D grating coupler working as a polarization beam combiner.

photodetectors at 7 MHz frequency, which indicated that the 2D grating provided a 35-dB crosstalk suppression between different polarization states.

3.2 Directional Coupler

When two waveguides are close to each other, the optical mode of each waveguide will couple or interfere with each other. A directional coupler is based on such phenomenon and becomes an essential component for various photonics devices. In this section, the basic working principle of the directional coupler will be in-depth analyzed.

Figure 3-12: Schematic of a general directional coupler with two different waveguides coupling with each other.

The coupling relationship between two arbitrary waveguides can be derived based on the coupled mode theory. Consider a general case when two waveguides have different shape and refractive index, as shown in Fig. 3-12. Two waveguides are named WG₁ and WG₂ with a refractive index of n_1 and n_2 respectively. The waveguide cladding has a refractive index of n_0 . The propagation constant for the two waveguides are β_1 and β_2 , respectively. When the input field at $z = 0$ is $A(0)$ and $B(0)$, the light field at length z can be expressed as [174]

$$\begin{cases} A(z) = \left\{ \left[\cos(qz) + i \frac{\delta}{q} \sin(qz) \right] A(0) - i \frac{\kappa}{q} \sin(qz) B(0) \right\} \exp(-i\delta z), & (3-6a) \end{cases}$$

$$\begin{cases} B(z) = \left\{ -i \frac{\kappa}{q} \sin(qz) A(0) + \left[\cos(qz) + i \frac{\delta}{q} \sin(qz) \right] B(0) \right\} \exp(i\delta z). & (3-14b) \end{cases}$$

where δ is the difference of propagation constant between two waveguides as

$$\delta = \frac{\beta_2 - \beta_1}{2}. \quad (3-7)$$

q is defined as

$$q = \sqrt{\kappa^2 + \delta^2}, \quad (3-8)$$

where κ is the coupling coefficient between two waveguides. For a more practical case, the light is only coupled to WG₁ at $z = 0$. In this case $A(0) = A_0$ and $B(0) = 0$. The power in both waveguide at length z can be derived as

$$\begin{cases} P_a(z) = \frac{|A(z)|^2}{|A_0|^2} = 1 - \sin^2(qz) \left(\frac{\kappa}{q}\right)^2, \\ P_b(z) = \frac{|B(z)|^2}{|A_0|^2} = \sin^2(qz) \left(\frac{\kappa}{q}\right)^2. \end{cases} \quad (3-9a)$$

$$(3-17b)$$

Eq. (3-17) shows that during transmission the power in both the waveguides is oscillating back and forth between each other. The oscillation amplitude can be defined as the power coupling efficiency F , which is

$$F = \left(\frac{\kappa}{q}\right)^2. \quad (3-10)$$

Fig. 3-13 plotted the relationship described by Eq. (3-17), with $F = 1$ and $F = 0.3$, respectively. In the case when $F = 0.3$, only a maximum 30% of power can transfer to the other waveguide. However, in the case when $F = 1$, the power can completely transfer from one waveguide to the other, which can provide larger flexibility for designing a practical waveguide coupling component. Thus, $\delta = 0$ or $\beta_2 = \beta_1$ is required for most of the coupler designs.

The length z where maximum power is coupled to the other waveguide is given by

$$z = \frac{\pi}{2q}(2m + 1), \quad (3-11)$$

where m is a non-negative integer. When $m = 0$, the length is defined as the coupling length L_c , which is expressed as

$$L_c = \frac{\pi}{2q} = \frac{\pi}{2\sqrt{\kappa^2 + \delta^2}}. \quad (3-12)$$

Figure 3-13: Power in both the waveguide as a function of qz when power coupling efficiency (a) $F = 1$ (b) $F = 0.3$.

For the particular case when $F = 1$, the coupling length becomes

$$L_c = \frac{\pi}{2\kappa} . \quad (3-13)$$

Although theoretically $P_a(z)$ and $P_b(z)$ always sum up to 1, real devices cannot. Fractional coupler intensity loss γ is defined to deal with this situation. The relationship is given by

$$|A(z)|^2 + |B(z)|^2 = (1 - \gamma)(|A_0|^2 + |B_0|^2). \quad (3-14)$$

This loss is significant and will largely affect the performance of some devices, such as the ring resonator. More detail will be discussed in the next chapter.

A directional coupler is usually designed as two waveguides with same refractive index and size so that $F = 1$ is always maintained. As shown in Fig. 3-14, two

Figure 3-14 SEM photo of a directional coupler. The coupling has two identical waveguides coupling with each other.

waveguides are brought together with a distance of 200-nm. A shorter distance gives higher coupling coefficient κ and a shorter coupling length L_c , which can reduce the overall components size. After a specific length z , those two waveguides are separated again. The coupling region length z is designed based on the requirement for directional coupler. In our case, the output power ratio of 50:50 and 1:99 is required, which correspond to a length of $z = L_c/2$ and $z = L_c$.

Since the waveguide structure has high dispersion, the effect of wavelength change to a directional coupler is calculated. As shown in Fig. 3-15a, each curve represents the power coupling from one to the other waveguide at different length of coupling region z . The coupling length L_c is the value of z where the curve reaches the maximum. The result indicates a significant drop in L_c when the wavelength is increasing from 1500 to 1630 nm. Consider the case for a 50:50 directional coupler at 1550 nm, the coupling length is fixed at $L_c = 11.94 \mu\text{m}$. When the different wavelength

Figure 3-15: (a) Waveguide power coupling as a function of length of coupling region and wavelength. (b) Power coupling ratio for different wavelength on a 50:50 directional targeting at 1550 nm.

is used on this device, the power coupling ratio will be different. As shown in Fig. 3-15b, the power coupling for 1500 to 1630 nm wavelength ranging from 0.35 to 0.8. As a result, the directional coupler can only be used for devices without much wavelength tuning requirements.

3.3 Multimode Interferometer

Multimode interferometer (MMI) is based on the self-imaging effect of multimode waveguides. The input field is repeated in one or more images at a fixed interval along the waveguide. It can be used for multiple beam splitting and combining. The main component for the MMI is a widened waveguide that can support a large number of modes. As shown in Fig. 3-16, the multimode waveguide has a width of W and length of L . The thickness $2a$ is based on the available wafer, which in our case is $2a = 220$ nm. The input and output ports for MMI are one or more single-mode waveguides with width $2d = 450$ nm. The refractive index for MMI and input & output waveguides are n_1 and for cladding material is n_0 . When the input is connected to the center of MMI, the interference pattern is symmetric in the x -axis, as shown in Fig. 3-17a. Because of the self-imaging effect, the input field has one identical image at the characteristic length L_{MMI} , two images at $L_{MMI}/2$, three images at $L_{MMI}/3$ and so on. Thus, by choosing a proper length and width, the device can be used as a 1 to N beam splitter. The characteristic length can be calculated by [174]

$$L_{MMI} = \frac{n_{eff} W^2}{\lambda}, \quad (3-15)$$

where n_{eff} is the effective refractive index in the y -axis. Using the n_{eff} in y -axis instead of n_1 for silicon structure, the calculation of the device can reduce to the calculation of a plane structure in the x - z plane. Here, the effective refractive index is $n_{eff} = 2.848$ for 220-nm thickness silicon. For an MMI structure with $W = 2.8$ μm , the calculated $L_{MMI} = 14.4$ μm .

When the input waveguide is not connected to the center, which corresponds to an asymmetric input case. The identical image appears at $4L_{MMI}$, and the two-fold

Figure 3-16: Schematic of a 1-to-N MMI device. The core is a multimode waveguide with width W . The input is a single mode waveguide and the output is multiple single mode waveguide.

image appears at $2L_{\text{MMI}}$, as shown in Fig. 3-17b, which is the basic working principle of a multiple input MMI. The length of $4L_{\text{MMI}}$ is used as 1:99 beam splitter and length of $2L_{\text{MMI}}$ is used for 50:50 beam splitter.

The effect of wavelength change is also calculated for the case of 1:99 beam splitter and a 50:50 beam splitter, which was designed for 1550 nm wavelength originally. As shown in Fig. 3-18, the transmission from one input port to both the output port is calculated. For the 1:99 beam splitter, the 90% tolerance range can cover the 100-nm wavelength range. For the 50:50 beam splitter, the result is even better. The whole spectrum range from 1500 to 1630 nm gives a splitting ratio tolerance of less than 3%. Compare to the beam splitter based on the directional coupler, which is discussed in the last section, the performance of MMI-based beam splitter is much more robust against wavelength drift. Comparing Fig. 3-15b with Fig. 3-18b, there is a 45% change for the directional coupler and a 3% change for MMI over the whole spectrum.

Figure 3-17: Interference pattern for a MMI device when (a) input is at the center and (b) input is not at center.

The stability of MMI makes it robust against fabrication error and environmental fluctuation.

The MMI device is then experimentally tested. Fig. 3-19a shows an SEM photo of the fabricated MMI device. The optimized device has a width of $W = 2.8 \mu\text{m}$ and length of $54 \mu\text{m}$, which is slightly shorter than the theoretically calculated value. The difference is mainly from the difference between the refractive index used in simulation and reality. The device is designed as a 1:99 beam splitter with two input ports and two output ports. Both the input and output ports are connected to the edge of MMI via a short taper. A broadband light amplified spontaneous emission (ASE) source (Amonics ALS-CL-15-B-FA) working at the C+L band with wavelength from 1500 to 1630 nm wavelength is used as light source. The input light is coupled to one input port of MMI.

Figure 3-18: The transmission of (a) 1:99 (b) 50:50 MMI as a function of wavelength.

Then both the output ports of MMI are monitored separately on an optical spectrum analyzer (OSA) (Yokogawa AS6370D). As shown in Fig. 3-19b, the spectrum from the two output ports has a maximum 17 dB extinction ratio, which corresponded to a splitting ratio of about 2:98.

(a)

(b)

Figure 3-19: (a) SEM photo of a MMI device. (b) Experimental output spectrum from a 1:99 MMI device.

3.4 Polarization Manipulation

Polarization manipulation components offer another degree of freedom to the photonic integrated circuit, which could potentially be used in data multiplexing,

information encoding and sensing. There are two major components covered in this section, polarization beam splitter and polarization converter. Both of them are studied theoretically and experimentally.

3.4.1 Polarization beam splitter

The propagation constants of the rectangle-shape waveguides have a significant difference between TE and TM mode. Due to the propagation constants difference, there is a significant difference in the coupling length between the two modes also. By carefully tuning the propagation constant (i.e., waveguide width), TE mode and TM mode may entirely separate into the different waveguides at a certain point, which is the basic working principle of such polarization beam splitter design. As shown in Fig. 3-20, the length corresponding the maximum and minimum transmission for both TE and TM mode are plotted as a function of waveguide width. The transmission here is measured from the same waveguide with input. There are six points in the Fig. 3-20, where the maximum of TE (or TM) mode coincides with the minimum of TM (or TE) mode. All the six points can be used for polarization splitting. Another concern for designing the couple region length is the coupling loss. To reduce the coupling loss, the couple region length should be as short as possible, and the width of the waveguide should be close to the 450-nm single mode design. As a result, point 1 and 3 are the priority.

Further simulations indicated that point 3 has a better extinction ratio than point 1. As shown in Fig. 3-21, the maximum transmission for TE mode is close to 1 and for TM mode is only 0.9. Minimum transmission for both TE and TM mode is close to 0. Thus, the case when TE mode has maximum transmission while TM mode has minimum transmission provides better extinction ratio. The result in Fig. 3-21

Figure 3-20: Maximum and minimum transmission for both TE and TM mode as a function of waveguide width.

corresponding to a transmission result at waveguide width of 467 nm. The gray area marked the operation point for PBS. A 3D-FDTD simulation also confirmed the result. The electric field distribution clearly shows the mode separation at the output port. The final design uses the waveguide width of 467 nm and couple region length of 55 μm .

To boost-up the PBS extinction ratio, a two-level cascaded design is used, as shown in Fig. 3-22a. The same structures are used in series to separate the two modes multiple times to get a higher extinction ratio. Light input from one of the input port is separated into two paths. The upper path contains the majority of TE mode and bottom path contains the majority of TM mode. A second stage removes the small amount of TM mode light in the upper path and the small amount of TE mode in the bottom path. The curve structure losses the light to avoid back reflection. In the end, TE mode comes out from path 1 and TM mode comes out from path 0.

Experimental testing verified the well-functioning of such PBS design. A broadband ASE source (Amonics ALS-CL-15-B-FA) is connected to a fiber-based PBS

Figure 3-21: Mode coupling relationship of the directional coupler-based polarization beam splitter.

and polarization controller to generate the required TE and TM mode light. Then the generated light is input in the chip-based PBS device. The output light passes through another set of calibrated polarization controller and fiber-based PBS. To select TE or TM mode measurement. The final spectrums from path 1 and path 0 are monitored on an OSA (Yokogawa AS6370D). When only TE mode light is input, the power output from path 1 is more than 40 dB higher than the output from path 0. Similarly, when only TM mode light is input, the power output from path 1 is more than 40 dB lower than the output from path 0, as shown in Fig. 3-22b. The optimum working wavelength is at about 1520 to 1525 nm for the current design. Minor tuning of the waveguide width could shift the optimum wavelength to our device operation wavelength at 1550 nm.

Figure 3-22: (a) SEM photo for the two-level cascaded PBS design. (b) Experiment result for the two-level PBS device.

3.4.2 Polarization rotator

As discussed in Section 3.2, maximum coupling could occur when phase matching condition is fulfilled between two modes, which means two modes can convert to each other when the effective refractive index for the two modes are the same. In the waveguide structure, the phase matching condition can be achieved easily by tuning the width of the waveguide. In this way, an efficient mode converter can be designed based on the directional coupler structure. As a special case, the converter for

Figure 3-23: (a) Phase matching condition for TE-TM mode coupling and electric field distribution at the phase match point for (b) TE (c) TM mode.

TE₀ and TM₀ can be used as a polarization rotator to switch between horizontal and vertical polarization states.

As shown in Fig. 3-23a, the effective refractive index for TE₀ is larger than TM₀ in the 200 – 700 nm range. Thus, a narrower waveguide is used for TE₀ input and output, while a wider waveguide with the same effective refractive index is used for TM₀ input and output. The horizontal dashed line marked the two widths that have a same effective refractive index, which is 290 nm for TE₀ and 525 nm for TM₀. All the width pairs with the same effective refractive index can be used here. The width pair of 290 nm and 525 nm here is chosen to guarantee the fabrication accuracy while maintaining an excellent

Figure 3-24: Mode coupling efficiency as the function of the (a) input and output waveguide width and (b) coupling length.

single mode performance. The mode profile of these two points is shown in Fig. 3-23b and Fig. 3-23c.

The maximum mode coupling efficiency for different waveguide width is calculated in Lumerical MODE solution software. As shown in Fig. 3-24a, different curves correspond to different input waveguide width from 288 to 292 nm, and the x -

axis is the output waveguide width from 500 to 550 nm. A very narrow coupling window is observed, which means that the maximum mode coupling efficiency is very sensitive to both the waveguide width. A 1-nm shift in input waveguide width can reduce the maximum coupling efficiency from 85% to 10%. For output waveguide width, the 50% tolerance range is 9 nm. The small tolerance ranges lead to a low yield for such devices. During our experiment, the wafer-level line width tolerance is about ± 5 nm. Testing structure with different width must include on the same wafer to confirm the final accuracy and get the optimum coupling efficiency.

The coupling length for such directional coupler is also calculated. Only TE mode light is input to the narrower waveguide, and the mode coupling efficiency for TE-TE coupling and TE-TM coupling are measured from the wider waveguide. As shown in Fig. 3-24b, the maximum coupling efficiency for TE-TM reaches 85% at about 600 μm coupling length, while the TE-TE coupling in the same waveguide is lower than 3%.

An SEM of the fabricated asymmetric directional coupler is shown in Fig. 3-25. A narrower waveguide is used for both input and output of TE mode and a wider waveguide is for TM mode input and output. The testing uses a similar experimental setup as the PBS testing. A broadband ASE source (Amonics ALS-CL-15-B-FA) is connected to a fiber-based PBS and polarization controller to generate the required TE mode light. Then the generated light is coupled to the narrower side of the polarization rotator device. The output light from the wider waveguide passes through another set of the calibrated polarization controller and fiber-based PBS to select the TE or TM mode. The final spectrums are monitored on an OSA (Yokogawa AS6370D). The testing result is shown in Fig. 3-25. A 15–25 dB difference between the TE output and

(a)

(b)

Figure 3-25: (a) SEM photo for TE-TM polarization rotator. (b) Testing result for TE-TM polarization rotator.

TM output from the same waveguide can be seen, which indicated a maximum 25 dB conversion ratio.

3.5 Mach-Zehnder Interferometer

Mach-Zehnder interferometer (MZI) is one of the essential waveguide devices. It has been demonstrated in different applications such as modulators, sensors, multiplexers and switches. The basic structure is shown in Fig. 3-26. It contains two directional coupler structures connected by two long waveguides, while the optical path length can be tuned to give an additional phase different ϕ between the two waveguides. The phase difference will cause constructive or destructive interference at the output. Due to the wavelength used for interference, only a nanometer level path length difference could give an observable output, which makes it a very accurate and sensitive sensor for the optical path length difference.

First, the governing equation for an MZI is derived based on Eq. (3-14) [174]. Consider a general case that two light waves A_0 and B_0 are input at two input ports of MZI. Both the directional couplers have maximum power coupling efficiency, which means $\delta = 0$ and $\beta_2 = \beta_1$. Thus, after the first directional coupler, the outputs are given by

$$\begin{cases} A_1 = A_0 \cos(\kappa l) - iB_0 \sin(\kappa l) , \\ B_1 = -iA_0 \sin(\kappa l) + B_0 \cos(\kappa l) , \end{cases} \quad (3-16a)$$

$$(3-24b)$$

where l is the coupling region length of the directional coupler. After the long waveguides with length L , the output becomes

Figure 3-26: Schematic of a waveguide-based Mach-Zehnder interferometer.

$$\begin{cases} A_2 = A_1 \exp(-i\beta L) , \\ B_2 = B_1 \exp(-i\beta L + i\phi) . \end{cases} \quad \begin{matrix} (3-17a) \\ (3-25b) \end{matrix}$$

After the second directional coupler, the output becomes

$$\begin{cases} A_3 = e^{-i\beta L} \left(-A_0 (e^{i\phi} \sin^2(\kappa l) + \cos^2(\kappa l)) \right) \\ \quad - e^{-i\beta L} (iB_0 (1 + e^{i\phi}) \sin(\kappa l) \cos(\kappa l)) \end{cases} \quad (3-18a)$$

$$\begin{cases} B_3 = e^{-i\beta L} \left(-A_0 (e^{i\phi} \sin^2(\kappa l) + \cos^2(\kappa l)) \right) \\ \quad - e^{-i\beta L} (iA_0 (1 + e^{i\phi}) \sin(\kappa l) \cos(\kappa l)) \end{cases} \quad (3-26b)$$

For a more practical case when light only couples to the upper waveguide, the input $A_0 = A_0$ and $B_0 = 0$. The output becomes

$$\begin{cases} A_3 = e^{-i\beta L} (-A_0 e^{i\phi} \sin^2(\kappa l) + A_0 \cos^2(\kappa l)) , \\ B_3 = e^{-i\beta L} (-iA_0 (1 + e^{i\phi}) \sin(\kappa l) \cos(\kappa l)) . \end{cases} \quad \begin{matrix} (3-19a) \\ (3-27b) \end{matrix}$$

From the Eq. (3-27), both the output A_3 and B_3 are oscillating as a function of phase shift ϕ while the total power remains unchanged. The product of coupling region length l and coupling coefficient κ of the directional coupler changes the maximum power coupled to the output B_3 , as shown in Fig. 3-27a. The maximum power coupling occurs when $\kappa l = 0.25\pi$, which corresponds to a 50:50 beam splitter. When $\kappa l < 0.25\pi$, the power from A_3 cannot fully couple to B_3 . The maximum and minimum value for A_3 and B_3 can be derived as

$$\begin{cases} \max|A_3|^2 = |A_0|^2 , \end{cases} \quad (3-20a)$$

$$\begin{cases} \min|A_3|^2 = |A_0|^2 \cos^2(2\kappa l) , \end{cases} \quad (3-28b)$$

$$\begin{cases} \max|B_3|^2 = |A_0|^2 \sin^2(2\kappa l) , \end{cases} \quad (3-28c)$$

$$\begin{cases} \min|B_3|^2 = 0 . \end{cases} \quad (3-28d)$$

Figure 3-27: (a) The output of MZI as a function of phase difference ϕ at different coupling region length l . (b) Extinction ratio at output A_3 when the coupling length of directional coupler changes.

Thus, the optimum design for MZI should have $\cos^2(2\kappa l) = 0$ to provide full tunability.

In a practical case, the directional coupler may not be perfect 50:50 ratio due to fabrication error. Is it better to use output B_3 if the maximum extinction ratio is critical, because the minimum is always 0. Similarly, if minimum transmission loss is critical,

Figure 3-28: Power difference between output A_3 and B_3 with different directional coupler coupling region length.

it is better to use output A_3 , because it can always reach maximum power. The effect of directional coupler power splitting ratio on the extinction ratio of A_3 is shown in Fig. 3-27b. Maximum extinction ratio occurs at $\kappa l = 0.25\pi$ as predicted. In order to get an extinction ratio above 20 dB, the acceptable range for κl is 0.03π , which is $\pm 6\%$. When $\kappa l = 0.25\pi$, the output power A_3 and B_3 becomes

$$\begin{cases} |A_3|^2 = |A_0|^2 \sin^2\left(\frac{\phi}{2}\right), & (3-21a) \end{cases}$$

$$\begin{cases} |B_3|^2 = |A_0|^2 \cos^2\left(\frac{\phi}{2}\right). & (3-29b) \end{cases}$$

During the experiment, MZI structure is also used to measure the actual coupling region length value for a 50:50 beam splitter. By fixing the phase difference $\phi = 0$, the output $A_3 = 0$ and B_3 reaches the maximum when the directional coupler has perfect 50:50 ratio. The power difference between A_3 and B_3 is shown in Fig. 3-28. Different curves correspond to different coupling region lengths l from 9 to 10.5 μm . The maximum power difference occurs at a wavelength of 1540 nm and 1558 nm for l

= 10 μm and 9.5 μm , respectively. As a result, the optimum design for 1550 nm beam splitter has $l = 9.7 \mu\text{m}$. The result is slightly smaller than the predicted value of 11.94 μm in Section 3.3. The reason is mainly that of the nonnegligible coupling at the curved input and output.

3.6 Ring Resonator

Ring resonator is another essential waveguide device for various applications including the sensor, modulator and laser. As shown in Fig. 3-29, ring resonator is made of a closed waveguide loop. A straight waveguide is placed close to the ring to form a directional coupler. Light can couple to the ring via evanescent coupling. The ring structure act as a Fabry–Pérot interferometer, in which only specific wavelength could resonant. The governing equation of ring resonator can be derived based on Eq. (3-14). We assume the feeding waveguide and ring resonator has the same propagation constant β . The coupling region can be treated as a directional coupler with coupling coefficient κ , coupling region length l and fractional coupler intensity loss γ . Output A and B is given by [174]

$$\begin{cases} A = (1 - \gamma)^{1/2} [A_0 \cos(\kappa l) - iB_0 \sin(\kappa l)] , & (3-22a) \\ B = (1 - \gamma)^{1/2} [-iA_0 \sin(\kappa l) + B_0 \cos(\kappa l)] . & (3-30b) \end{cases}$$

After the ring structure with length L , B_0 can be expressed as

$$B_0 = B \exp \left(-\frac{\rho}{2} L - i\beta L \right) , \quad (3-23)$$

where ρ is the intensity attenuation coefficient. The amplitude transmission from the straight waveguide can be expressed as

Figure 3-29: Schematic of a ring resonator. A straight waveguide is placed close to the ring to form a directional coupler. Light can couple to the ring via evanescent coupling.

$$\frac{A}{A_0} = (1 - \gamma)^{\frac{1}{2}} \left[\frac{\cos(\kappa l) - (1 - \gamma)^{\frac{1}{2}} \exp\left(-\frac{\rho}{2}L - i\beta L\right)}{1 - (1 - \gamma)^{\frac{1}{2}} \cos(\kappa l) \exp\left(-\frac{\rho}{2}L - i\beta L\right)} \right]. \quad (3-24)$$

To simplify the expression, new parameters x , y and ϕ are defined as

$$\begin{cases} x = (1 - \gamma)^{\frac{1}{2}} \exp\left(-\frac{\rho}{2}L\right), & (3-25a) \end{cases}$$

$$\begin{cases} y = \cos(\kappa l), & (3-33b) \end{cases}$$

$$\begin{cases} \phi = \beta L = \frac{2\pi}{\lambda} n_{eff} L. & (3-33c) \end{cases}$$

Then, the intensity transmission can be derived as

$$T(\phi) = \left| \frac{A}{A_0} \right|^2 = (1 - \gamma) \left[1 - \frac{(1 - x^2)(1 - y^2)}{(1 - xy)^2 + 4xy \sin^2\left(\frac{\phi}{2}\right)} \right]. \quad (3-26)$$

The relationship between T and λ is shown in Fig. 3-30a. A periodic drop in the transmission spectrum is observed, which is due to the constructive resonance of the ring resonator. The simulated electric field distribution at resonance wavelength and away from resonance wavelength is shown in Fig. 3-30b and Fig. 3-30c respectively. When the input is at resonance wavelength, light field enhancement inside the ring and

Figure 3-30: (a) The intensity transmission spectrum of a ring resonator structure. (b) The simulated electric field distribution when $T = T_{\min}$ and (c) $T = T_{\max}$.

a reduction in the feeding waveguide is observed. When the input is away from resonance wavelength, nearly no light couples to the ring resonator. The resonant condition of the ring can be expressed as

$$\phi = \beta L = 2m\pi . \quad (3-27)$$

The maximum and minimum value of the transmission can be derived as

$$\begin{cases} T_{\max} = \frac{(1 - \gamma)(x + y)^2}{(1 + xy)^2}, \\ T_{\min} = \frac{(1 - \gamma)(x - y)^2}{(1 - xy)^2}. \end{cases} \quad (3-28a) \quad (3-36b)$$

Extinction ratio ER is defined as the ratio between maximum and minimum transmission, which is $ER = 10 \log(T_{\max}/T_{\min})$. ER reaches the maximum when $T_{\min} = 0$, which is also referred to the critical coupling condition. Critical coupling occurs when $x = y$ or

$$(1 - \gamma)^{\frac{1}{2}} \exp\left(-\frac{\rho}{2}L\right) = \cos(\kappa l). \quad (3-29)$$

Cavity enhancement factor B measures the power enhancement inside the ring due to constructive interference, which can be derived as

$$B = \left|\frac{B_0}{A_0}\right|^2 = \frac{x^2(1 - y^2)}{1 + x^2y^2 - 2xy \cos(\phi)}. \quad (3-30)$$

The width of the resonant peak is an important parameter to evaluate the performance of the ring resonator. The full width at half maximum (FWHM) is defined as the width of the peak at half the peak height, which can be derived as

$$FWHM(\phi) = \frac{2(1 - xy)}{\sqrt{xy}}, \quad (3-31)$$

which is measured as the width in ϕ . In most case, finesse F or Q-factor is used instead of $FWHM$. Finesse is defined as

$$F = \frac{2\pi}{FWHM} = \frac{\pi\sqrt{xy}}{1 - xy}, \quad (3-32)$$

where 2π is the distance between adjacent resonance peak, namely free spectral range (FSR). Moreover, Q-Factor is defined as

$$Q = \frac{\phi}{FWHM} = \frac{\pi n_{eff} L \sqrt{xy}}{\lambda(1 - xy)} . \quad (3-33)$$

FSR could also be expressed as the distance with the unit of frequency or wavelength by

$$FSR_f = \frac{c}{n_{eff} L} , \quad (3-34)$$

$$FSR_\lambda = -\frac{\lambda^2}{n_{eff} L} . \quad (3-35)$$

The simple relationship indicates that FSR can be a great tool to assist the measurement of the effective refractive index.

Since x is a constant only related to material properties and fabrication accuracy, only y is tunable in devices design. Consider a 40- μm radius ring with coupling loss of 0.1 dB and propagation loss of 3 dB/cm, the fractional coupler intensity loss is $\gamma = 0.0228$ and the intensity attenuation coefficient is $\rho = 0.0172$. The Q-factor, extinction ratio and cavity enhancement factor at different coupling conditions are shown in Fig. 3-31. The critical-coupling point is marked with the dashed line. On the left side of the critical-coupling point is the under-coupling region, and on the right side is the over-coupling region. The cavity enhancement factor of the ring also reaches the maximum of $B_{max} = 36$ at the critical coupling point, which means the power inside the ring is 36 times higher than input power. The high power can be used to stronger light-matter interaction. The maximum $Q = 8 \times 10^4$ is at under-coupling region when $\kappa l = 0$. At the critical-coupling point, the Q-factor drops to 4×10^4 . As a result, there is a trade-off between a higher extinction ratio and a higher Q-factor depends on the application. Using as modulators require a higher extinction ratio, but using as sensor requires a higher Q-factor.

Figure 3-31: Q-factor, extinction ratio and cavity enhancement factor of ring resonator as different coupling conditions.

The transmission and cavity enhancement factor near the resonant peak is shown in Fig. 3-32. The two parameters are highly correlated. At under-coupling region, both the peaks are narrower and shallower compare to others. This corresponding to a higher Q-factor and lower extinction ratio. Similarly, at the critical-coupling point, the results show a higher extinction ratio and higher Q-factor compare to those at under-coupling region. At over-coupling region, both the Q-factor and extinction ratio drops significantly.

Figure 3-32: Transmission and cavity enhancement factor near the resonant peak at different coupling conditions

Although the Q-factor and cavity enhancement factor can be tuned by coupling conditions, the maximum value of them are also largely affected by the coupling loss and waveguide loss. Here the critical coupling Q-factor at different losses is investigated. The results are shown in Fig. 3-33a. The coupler loss is assumed in the range of 0.02 to 1 dB and waveguide loss is in the range of 0.2 to 10 dB/cm. The color

(a)

(b)

Figure 3-33: (a) Q-factor at critical coupling point as a function of coupler loss and waveguide loss. (b) SEM photo of an array of ring resonator.

bar in the graph shows the simulated Q-factor in the logarithm scale. The result indicated that Q-factor drops for an order of magnitude when coupler loss increases from 0.02 to 0.4 dB. The effect for waveguide loss is much lower. The drop is less than 30% when waveguide loss increases from 0.2 to 10 dB/cm. Thus, based on a known Q-factor, the coupler loss can be estimated. Or in a reverse way, if the coupler loss can be

Figure 3-34: Experimental result for Q-factor and extinction ratio at different coupling length.

measured precisely, the upper bond for Q-factor can be estimated. For a typical case in our experiment, the waveguide loss is about 2 dB/cm and the maximum Q-factor is at about 6×10^4 level. From the result, we can estimate the coupler loss to be around 0.1 dB.

In the experiment, the ring resonator with cross-section of 450×220 nm is built. The coupling gap is fixed at 300 nm. An array of ring resonators with different coupling length are included in the design, as shown in Fig. 3-33b. The transmission spectrums for each ring are measured. The Q-factor and extinction ratio are then interpreted from the spectrum, as shown in Fig. 3-34a and Fig. 3-34b. Both results have the same trend with the prediction. The fitting is not perfect mainly due to the limitation of measurement methods. There's some difficulty in measuring high Q-factor especially when extinction ratio is low. The peaks might be submerged by the background noise. The selected spectrum for coupling length of 0, 7 and 13 μm is shown in the inset of Fig. 3-34a. It is worth noticing the fluctuation of the spectrum and how it may affect the shape of the resonant peak.

3.7 Fabrication Process

In this section, the fabrication of passive devices is discussed in detail. The chip is fabricated using the nano-photonic fabrication technologies, which is compatible with the standard complementary metal-oxide-semiconductor (CMOS) fabrication. A CMOS compatible process can entirely make use of the different materials and fabrication equipment that has been developed over more than 50 years. Adopt the CMOS process may significantly reduce the fabrication cost and is suitable for mass production. The most commonly used material for silicon photonics is silicon-on-insulator (SOI) wafer, especially silicon on silicon dioxide (SiO_2) wafer, which is initially used in semiconductor devices to reduce parasitic capacitance. Silicon photonics builds waveguide structures based on the SOI structure, with silicon core and SiO_2 cladding. The refractive index of silicon is 3.48 at 1550 nm wavelength, while SiO_2 is 1.44. Due to the large refractive index difference between silicon and SiO_2 , light

Figure 3-35: Transmission window of some commonly used silicon photonic material.

can be strongly confined in the silicon layer. Minimum bending radius of about $5\ \mu\text{m}$ is achievable for silicon waveguide. As a comparison, the minimum bending radius for single-mode fiber is about 2 cm. Smaller bending radius ensures smaller components can be built using this technology. Another consideration for silicon photonics is the transmission wavelength. Fig. 3-35 shows the transmission window of some commonly used silicon photonic material. Silicon has a transparent window of about 1-8 μm and SiO_2 is transparent at 0.3-3.5 μm , which is perfect for 1310 nm and 1550 nm fiber communication wavelength. If visible light is required, silicon must be changed to other material such as aluminium nitride or silicon nitride.

The standard fabrication process for passive silicon devices is shown in Fig. 3-36. The process is a wafer level process that accepts only the 8-inch wafer. Each unit exposure field has a maximum size of $25 \times 33\ \text{mm}$. The SOI wafer we use has a 220-nm thickness crystallized silicon as the device layer and a 2- μm thickness SiO_2 as the insulation layer. Starting with this wafer, the first step is to add a laser mark on the

Figure 3-36: Fabrication process for passive components.

Figure 3-37: Photo resist (a) without and (b) with bottom anti-reflective coating.

wafer for future process tracking. Then the wafer is pre-cleaned with quick dump rinse (QDR) and sulfuric peroxide mix (SPM). SPM is a strong oxidizing agent that is used to clean organic residues on the wafer.

After preparing the wafer, step 2 is to pattern the grating structure. 193-nm deep UV lithography and reactive ion etching (RIE) steps are used in this step. Firstly, A layer of 90-nm bottom anti-reflective coating (BARC) and a 400-nm thickness photoresist is dispensed on the wafer surface. BARC is used to avoid backscattering and standing waves during the exposure, which could provide better control for the pattern profile. Then after exposure, Fig. 3-37 shows the photoresist with and without BARC layer. Typical standing wave pattern on photoresist sidewall is observed in Fig. 3-37a. Because the final device dimension is affected mainly by the energy and focus of exposing process, a focus exposure matrix (FEM) test is performed when a precise device dimension is required. As shown in Fig. 3-38, each unit cell on the wafer has a different lithography parameter, with the x -axis as exposure energy in the unit of millijoule and y -axis as focus height in the unit of μm . Then the resulting pattern is measured in a critical dimension scanning electron microscope (CD-SEM). Two critical

x/y	21	22	23	24	25	26	27	28	29	30	31	32	33	34	35
-0.375					NA	NA	NA	NA	NA	NA	NA				
-0.3			NA	187	176	34.5	NA	NA	NA	NA	NA	NA	NA		
-0.225		NA	211	182	182	NA	161	NA	NA	NA	NA	NA	NA	NA	
-0.15	NA	NA	216	197	182	179	167	157	151	NA	160	NA	NA	NA	NA
-0.075	NA	238	217	217	191	183	179	161	150	148	128	131	NA	NA	NA
0	NA	243	227	218	205	190	184	169	151	157	141	137	122	121	113
0.075	248	244	224	218	206	193	179	165	157	159	147	141	129	123	115
0.15	242	239	231	219	209	195	186	180	167	159	153	137	132	122	115
0.225		243	226	188	202	191	190	174	165	154	145	135	129	NA	
0.3			219	206	185	192	175	176	NA	156	136	123	NA		
0.375					187	137	165	142	143	NA	NA				

(a)

x/y	21	22	23	24	25	26	27	28	29	30	31	32	33	34	35
-0.375					1263	1240	1236	1218	1208	1203	1196				
-0.3			1289	1272	1253	1243	1231	1228	1214	1206	1207	NA	NA		
-0.225		1312	1292	1276	1263	1252	1242	1238	1223	1225	NA	NA	1213	1225	
-0.15	1332	1304	1297	1279	1265	1257	1250	1237	1229	1226	1207	1213	1211	1213	1218
-0.075	1330	1311	1297	1283	1269	1263	1255	1230	1228	1228	1228	1221	1218	1218	1232
0	1330	1313	1297	1285	1274	1267	1240	1221	1241	1233	1234	1222	1223	1220	1231
0.075	1331	1311	1301	1288	1250	1270	1261	1251	1246	1237	1234	1230	1228	1230	1241
0.15	1333	1319	1286	1288	1281	1270	1264	1254	1246	1241	1234	1231	1230	1236	1248
0.225		1311	1302	1280	1275	1275	1265	1258	1249	1241	1238	1236	1234	1248	
0.3			1315	1301	1288	1279	1269	1260	1250	1247	1245	1244	1248		
0.375					1296	1286	1275	1263	1257	1259	1256				

(b)

Figure 3-38: Focus exposure matrix wafer mapping for (a) 200 nm and (b) 1270 nm designed linewidth.

dimension measurements are shown. Fig. 3-38a shows the result when the targeting

linewidth is 200 nm and Fig. 3-38b has a designed linewidth of 1270 nm. The measured linewidth is marked in the center of each unit cell. “NA” corresponded to the case when the exposure step failed to generate any pattern with the specific parameter. Among the results, those in the ± 10 nm error range are marked with black rectangles, which are the optimum parameters. Different error range can be chosen also depends on the following process. For example, a -10~-20 nm error range is usually chosen for the silicon waveguide etching process. After the lithography step with the correct parameter, the silicon layer is partially etched for 110 nm by RIE. For grating structure, the RIE etch depth is controlled by precise timing and measured with a thin film measurement machine. Finally, after etching, the photoresist is stripped with oxygen plasma and the whole wafer is cleaned using dilute hydrofluoric acid (DHF) and SPM. The result for grating patterning step is shown in Fig. 3-39a.

The 3rd step is to deposit a 70-nm thickness SiO₂ using plasma enhanced chemical vapor deposition (PE-CVD). The SiO₂ layer is used as a hard mask for silicon waveguide etching, due to the high etching selectivity between them.

In the 4th step, the waveguide structure is patterned. A lithography process same as step 2 is used to pattern the photoresist. Then the hard mask is etched using RIE. After oxide thickness measurements confirm there is no oxide left, the photoresist is stripped and the wafer is cleaned. Using the remaining hard mask, the silicon is partially etched for 130 nm depth. The remaining 90-nm silicon layer is left for the rib structure in step 5. A photo of the etched waveguide structure with the false-colored hard mask is shown in Fig. 3-39b.

Step 5 is for the rib structure patterning. Another lithography step masked the rib region with the photoresist. The photoresist together with the hard mask protected the waveguide and rib structure in the final RIE etching step. Different from the

(a)

(b)

Figure 3-39: (a) SEM photo after grating patterning. (b) SEM photo for etched waveguide structure with hard mask on top.

previous RIE etching steps, the final RIE step can be controlled by monitoring the endpoint during the etching process, which may offer higher robustness against fabrication error. After this step, all the patterning for the silicon device layer is finished. As shown in Fig. 3-40, the three layers with different thickness are used in different devices such as MZI and grating couplers.

In step 6, after cleaning the wafer again, the cladding oxide layer is deposited. There are two types of oxide involved, plasma-enhanced chemical vapor deposition (PECVD) undoped silicon glass (USG) and high-density plasma chemical vapor deposition (HDP-CVD) oxide. HDP-CVD oxide is used as the majority of cladding material for its better gap filling performance, but the process may damage the silicon surface and cause corner erosion. The gap filling void from PECVD USG oxide is shown in Fig. 3-41a. To avoid both the corner erosion and gap filling void, 80-nm PECVD USG oxide is deposited first as protection and then follows by 1-2 μm HDP-CVD oxide. After this thick oxide layer deposition, the stress may cause the wafer bend. Another layer of 200-nm thickness amorphous silicon layer is deposited on the back side of the wafer to balance the stress.

Step 7 is for the first trench etching. The first trench is used to prepare a smooth chip sidewall for waveguide side coupling, which could minimize the coupling loss. Usually, the trench is designed with a width of 100 μm , which is slightly larger than the dicing blade with a thickness of 40 μm . In this step, the trench is etched through all the SiO_2 layer on top of the silicon substrate.

In the 8th step, a 40-nm aluminium oxide (Al_2O_3) is deposited by atomic layer deposition (ALD) as sidewall protection and hard mask for vapor HF release. Then, another deep trench with 100 μm depth is etched inside the SiO_2 trench using deep RIE. The deep trench cleared space for fiber side coupling. The resulting trench is shown in

(a)

(b)

Figure 3-40: SEM photo of the final silicon structures with rib layer colored in red, grating layer colored in yellow and waveguide layer colored in blue.

(a)

(b)

Figure 3-41: (a) TEM shown the gap filling void from PECVD USG oxide deposition. (b) SEM photo of etched trench.

Fig. 3-41b. The top SiO₂ layer has a smooth sidewall and the bottom silicon substrate has a typical periodic structure from deep RIE.

Finally, for step 9 and step 10, the window structure is patterned. The window structure is used to expose and even suspend some silicon structure, which is essential in some sensor or optomechanical applications. After patterning the photoresist, Al₂O₃ is etched first using metal RIE. Then the photoresist is stripped, and the wafer is cleaned. Finally, the oxide layer is removed by vapor HF isotropic etching. The etching depth is controlled by reaction time. As shown in Fig. 3-42, the window area is marked in purple. Some waveguide structure are exposed in the window area. In Fig. 3-42b, it is clear to see the undercut caused by isotropic vapor HF etching.

3.8 Summary

This chapter focused on the design, simulation, fabrication and experiments of various silicon photonic passive components. In section 3.1, the fiber-chip coupler is analyzed. The simulated coupling loss is 3 dB for side coupling and 4 dB for the grating coupler, which perfectly matched our testing result. To improve the performance, side coupling requires a tip width low than the lithography limitation. Thus, resolution enhancement technologies could be applied in the future works, such as a phase-shift mask or multiple patterning. For the grating coupler, the improvement could focus on reflecting the diffracted light at the substrate. In section 3.2, a theoretical model of the directional coupler is provided, which is the basis for all the mode coupling devices, such as MZI, ring resonator, mode converter and beam splitters. In section 3.3 the MMI structure is analyzed, which can be used as a robust beam splitting and combining device. The testing results also confirmed the wide operation bandwidth of more than

(a)

(b)

Figure 3-42: SEM photo after window opening and vapor HF release. The window areas are false colored in purple.

130 nm. In section 3.4 two important polarization manipulation components are designed and analyzed, the polarization beam splitter and polarization mode converter. The testing result shows a 40-dB extinction ratio for polarization beam splitter and a

25-dB extinction ratio for polarization converter. The successful operation of such devices opens the possibility for quantum information applications using polarization encoding. In section 3.5 and 3.6, the theoretical model for MZI and ring resonator are provided, which could offer guidance for optimizing the devices based on such structures. Finally, in Section 3.7, the fabrication process for all the passive components are summarized in detail. All the Si-based components are designed and integrated in the QKD chip. The selection of these components is based on extensively design and experiment. Grating couplers present at the input and output port of chip. Directional couplers are the core of MZI and ring resonator. MZIs are used in the modulators in Chapter 4. Polarization manipulation components are used for the polarization multiplexing of signal and local oscillator light described in Chapter 5. In this chapter, a design library and selection guideline are offered for future works when coupler, beam splitter and polarization manipulation are involved.

Chapter 4 Silicon Photonic Active Components

The active components are another group of fundamental components for quantum photonic chip, including source, modulator and detector. One property in common is that they have an intentional dynamic light-matter interaction. The design of active components is largely based on the passive components stated in the last chapter. In this chapter, two types of active components are covered, which is the modulator and detector. First, modulators based on thermo-optic effect, carrier injection or depletion effect and electro-optic effect are theoretically and experimentally studied. These are the three major modulators available on the silicon photonic platform. Then, the design and testing of a germanium photodiode are introduced, which is then used to build a balanced homodyne detector for quantum information applications. Finally, the fabrication process for silicon photonic active components is summarized.

4.1 Modulator

The optical modulator is the device that changes a beam of light based on the applied signal. The modulation could add on intensity, phase or polarization of the input beam. In a photonic circuit, this is typically achieved by changing the refractive index of the material. There are three options available for silicon photonic platform, thermo-optic modulator, carrier injection modulator and electro-optic modulator.

4.1.1 Thermo-optic modulator

Thermo-optic modulator utilizes the thermo-optic effect of material, in which the refractive index of the material is changed by changing the temperature. Silicon happens to have a relatively high thermo-optic coefficient of [175]

$$\frac{dn}{dT} = 1.86 \times 10^{-4} \text{ RIU/}^{\circ}\text{C}. \quad (4-1)$$

As a comparison, the thermo-optic coefficient for silicon nitride is 2.45×10^{-5} RIU/ $^{\circ}\text{C}$, and silicon dioxide is 0.95×10^{-5} RIU/ $^{\circ}\text{C}$. For a 500- μm length waveguide, only a temperature rises of 6 $^{\circ}\text{C}$ could shift the phase by 2π . Thus, a heater structure can be introduced near the waveguide to change the temperature. Generally, there are two ways to heat up the silicon waveguide. One is to introduce another layer of material as a heater on top of the waveguide. Another option is using doped silicon as heater beside the waveguide. Both the cases are simulated in COMSOL software. The model cross-section is shown in Fig. 4-1. The SiO_2 – Air interface is assumed to have a heat transfer coefficient of $h = 10 \text{ W}/(\text{m}^2 \cdot \text{K})$. The bottom of the silicon substrate is assumed to have a high-efficiency heat sink with a constant temperature of 20 $^{\circ}\text{C}$. The heater has a constant power output with the unit of $\text{mW}/\mu\text{m}$. The color map in the graph is the

Figure 4-1: Two types of heating methods. (a) TiN heater above waveguide structure (b) Doped silicon heater beside waveguide structure

temperature distribution with the unit of degree Celsius (°C). Then the center temperature of the waveguide is calculated at different heater power, as shown in Fig. 4-2a. The result shows that both the heating configurations have similar heating efficiency. The heater power of about 0.03 mW/μm is capable of increasing the waveguide temperature for about 6 °C, which is achievable on a CMOS chip.

Our measurement shows the electrical resistivity for TiN thin film is $1.55 \times 10^{-6} \Omega \cdot \text{m}$. A standard TiN design has a thickness of 150 nm and a minimum width of 1 μm.

Figure 4-2: (a) Temperature change as a function of the heater power for waveguide heater and TiN heater. (b) I-V curve for doped silicon with different doping concentration.

For a TiN heater with 500-μm length and 3-μm width, the voltage and current required for 2π phase shift are 5V and 3mA.

For doped silicon, the electrical resistivity is related to the doping concentration, which can be simulated by Silvaco TCAD software. Boron doping on 220-nm thickness

Figure 4-3: Simulated step response of the two types of heater. Both are capable of operating at 10 kHz frequency.

silicon with the energy of 20 keV and dose of 1×10^{13} , 1×10^{14} and 4×10^{15} ions/cm² are calculated. The resistance testing region has a length of 3 μm and a width of 1 μm. The resulting I-V curves are shown in Fig. 4-2b. The sheet resistivity for the 220-nm thin film at the three doses is 281 Ω, 1542 Ω and 7788 Ω, which corresponded to a bulk resistivity of 6.18×10^{-5} Ω·m, 3.39×10^{-4} Ω·m and 1.71×10^{-3} Ω·m. As a result, doped silicon heater could also provide similar performance as TiN heater. Both the methods can meet the requirements for the thermo-optic modulator.

The dynamic response is another important consideration for modulators. Step response for heater is also simulated in COMSOL, as shown in Fig. 4-3. The time constant is marked with the gray dashed line. Both the cases have a time constant of about 10 μs, while the waveguide heater is slightly faster. Thus, the thermo-optic modulator is capable of operating at about 10 kHz modulation frequency.

Fig. 4-4 shows the microscopic photo of the fabricated MZI modulator using Al and TiN as the heater material. The Al heater uses a snake shape to increase the

(a)

(b)

Figure 4-4: Microscopic photo of thermal optics modulator based on (a) Al heater and (b) TiN heater.

resistance. The dynamic response of the TiN-based device is then experimentally tested. The experimental setup is similar to the one used in PBS testing. A 1550-nm CW laser (Santec TSL-510) is connected to a fiber-based PBS and polarization controller to

Figure 4-5: Experimental test result for the thermo-optic modulator under 10 kHz square wave.

generate the required TE mode light. Then the input light is coupled to one of the input ports of MZI. The arm of MZI is tuned by applying a voltage signal from a function generator (Philips PM5193) via an RF probe. The output light is measured on a high-speed photodetector (MenloSystems FPD-510) connected to an oscilloscope (Tektronix MDO4104B-3). The modulation result at 10 kHz square wave signal input is shown in Fig. 4-5. The result fits the prediction very well.

4.1.2 Carrier injection or depletion modulator

The density of free carriers can also change the refractive index of a semiconductor material, which can be described by the Drude-Lorenz equation as follows [175],

$$\Delta n = \frac{-e^2 \lambda_0^2}{8\pi^2 c^2 \epsilon_0 n} \left(\frac{N_e}{m_{ce}^*} + \frac{N_h}{m_{ch}^*} \right), \quad (4-2)$$

where e is the electron charge, λ_0 is the free space wavelength, c is the vacuum speed of light, ϵ_0 is the permittivity of free space, n is the refractive index, N_e is the free

electron concentration, m_{ce}^* is the effective mass of electron, N_h is the free hole concentration, m_{ch}^* is the effective mass of holes. Soref and Bennett provide an experimental-based equation for 1550 nm wavelength [175], which is expressed as

$$\Delta n = -\left[8.8 \times 10^{-22} \Delta N_e + 8.5 \times 10^{-18} (\Delta N_h)^{0.8}\right]. \quad (4-3)$$

There are two major ways to actively tuning the concentration of free carriers. One is by depleting the pre-doped free carriers inside the waveguide with a reverse voltage on the p - n junction, which is referred to as the carrier depletion modulator. The other is by injecting excessive free carriers into the waveguide region from outside with a forward voltage on the p - i - n junction, which is referred to as the carrier injection modulator.

For a carrier depletion modulator, a p - n junction is first formed in the waveguide region. The free carrier concentration can be simulated by Silvaco TCAD software. As shown in Fig. 4-6a, a 70-nm thickness rib structure is built on the side of the waveguide as a connection to electrodes. Phosphorous is doped on the right-hand side with a mask stopping at the center of the waveguide and boron are doped on the left-hand side also with a mask stopping at center. Both the doping have a dose of 1×10^{14} ions/cm². The resulting dopant concentration distribution in logarithm scale is shown in Fig. 4-6a and Fig. 4-6b. A cutting line at 30-nm height for the two distributions are shown in Fig. 4-6c. The maximum concentration for both the dopants are about 5×10^{18} /cm³. The p - n junction will form in the center of the waveguide as designed. The simulated I-V curves are shown in Fig. 4-6d. Both the curves are the same only that the upper graph is shown in linear scale and the bottom graph shows the modulus of the current in the logarithmic scale. The I-V curve shows a typical p - n junction characteristic with a reverse breakdown at about -6.3V.

Figure 4-6: Phosphorous and boron doping concentration distribution (log scale) for a p - n junction modulator. (a) Phosphorous. (b) Boron. (c) Horizontal cutting line at 30 nm. (d) Simulated I-V curve of the p - n junction in linear and log scale.

Figure 4-7: (a) Electron and hole concentration distribution (log scale) at different voltages. (b) Average hole and electron concentration in the waveguide region at different voltage.

The electron and hole concentration distributions at different voltages are also simulated. Some selected results are shown in Fig. 4-7a. With a forward biased voltage

added, more electrons and holes are injected into the waveguide region, which is referred to as the carrier injection mode. With reversed biased voltage added, pre-doped electrons and holes are depleted from the waveguide region, which is referred to the carrier depletion mode. After the reverse break down, the electron and holes concentrations are increasing again. A weighted average concentration in the waveguide region using TE mode profile as weight is shown in Fig. 4-7b, which provides an estimation for the resulting refractive index change. The reverse biased region before breakdown provides an excellent linear relationship between voltage and concentration. When reversed voltage increases, the concentration decreases, and refractive index increases. At reverse breakdown voltage and forward voltage, the concentration increases with the applied voltage, which will cause the refractive index to drop in the waveguide.

Another type of modulation configuration can only work in carrier injection mode, which is based on a $p-i-n$ structure. In this case, doping is kept 500-nm away from the waveguide, as shown in Fig. 4-8a and Fig. 4-8b. A 70-nm thickness rib structure is also used to connect the doped region with the waveguide. The doping mask for phosphorous and boron ends at 2.75 μm and 1.25 μm , respectively, which is marked with gray dashed line in Fig. 4-8c. The waveguide region has only background level boron doping, which acts as the insulator region of the $p-i-n$ structure. The I-V curve of the $p-i-n$ structure is shown in Fig. 4-8d. There is no reverse breakdown happens in the 0 to -10 V region. The free carrier concentration for the $p-i-n$ structure is also simulated, as shown in Fig. 4-9a. At forward biased voltage, the free carrier concentration increases when voltage increase. The main difference for $p-n$ structure and $p-i-n$ structure is that for $p-i-n$ structure there is nearly no free carrier to deplete at the reverse bias voltage. The devices can only work in carrier injection mode. The weighted

Figure 4-8: Phosphorous and boron doping concentration distribution (log scale) for a p - i - n junction modulator. (a) Phosphorous. (b) Boron. (c) Horizontal cutting line at 30 nm. (d) Simulated I-V curve of the p - i - n junction in linear and log scale.

Figure 4-9: (a) Electron and hole concentration distribution (log scale) at different voltages. (b) Average hole and electron concentration in the waveguide region at different voltage.

average concentration in the waveguide region is shown in Fig. 4-9b. The result shows that the concentration at the reverse biased region is lower than $10^{-6} / \text{cm}^3$, which could only cause negligible refractive index change.

Another consideration for carrier injection or depletion modulator is the heat effect from electric current. Based on the I-V curve in Fig. 4-6d and the temperature power relationship in Fig. 4-2a, the absolute value of phase shift caused by the heater is plotted together with the phase shift caused by the free carrier in Fig. 4-10a. All the phase shift here is calculated relative to the lowest possible value to avoid any negative data in log-scale. One thing to notice is that the phase shift caused by the free carrier is always negative, and the phase shift from heat is always positive. Because of the joint effect of heat and carrier injection, the final phase shift will follow the upper bound in this figure, which is in the range of 10^{-3} to 10^1 . Thus, the result can be separated into 3 regions, which are marked with blue shade. Region 1 and 3 is dominant by heat, so when the absolute value of voltage in this region, the phase shift will increase. Instead, region 2 is dominant by the free carrier, which coincides well with the linear region in Fig. 4-7b. In this region, the phase shift decreases when the voltage increases. The experimental testing result of a *p-n* modulator is shown in Fig. 4-10b. The black curve is the I-V curve, which indicated the function of the fabricated *p-n* junction as predicted. The red curve shows the phase shift at different voltages, which shows a perfect match with prediction. At below -7 V, the *p-n* junction is at the reverse break down region, the phase increases significantly with heat when the reversed voltage increase. At about -7 to 4 V, the phase shift has a steady linear relationship with the voltage increase, which is dominated by the free carriers. When the applied voltage is above 4 V, the heating effect takes over again and causes the phase shift increases with the applied voltage. The phase shift here is measured by monitoring the resonant peak shift of a ring resonator while changing the applied voltage on the *p-n* modulator. The experimental setup is like the one used in thermos-optic modulator testing. A broadband ASE source (Amonics ALS-CL-15-B-FA) or a tunable laser (Santec TSL-510) is

Figure 4-10: (a) Simulated phase shift caused by heat and free carrier. (b) Experimental testing of the I-V curve and phase shift result for a *p-n* modulator.

connected to a fiber-based PBS and polarization controller to generate the required TE mode light. Then the input light is coupled to the feeding waveguide of the ring resonator. The ring resonator is tuned by applying a voltage signal from a function generator (Philips PM5193) or a power source (Keithley 236) via a RF probe. The output light is measured on a high-speed photodetector (MenloSystems FPD-510)

Figure 4-11: (a) Ring resonator peak shift as a function of applied voltage. (b) Experimental result of a MZI-based amplitude modulator.

connected to an oscilloscope (Tektronix MDO4104B-3) for the dynamic response or an OSA (Yokogawa AS6370D) for the spectrum. As shown in Fig. 4-11a, the resonant peaks at 6, 4, 0, -6 and -7.8 V are selected to show the overall trend. Blueshift or redshift of the resonant peak corresponded to a dropping or increase of phase, respectively. The ring resonator used in this testing is shown in Fig. 4-12a, the ring resonator has a total

(a)

(b)

Figure 4-12: Microscopic photo of carrier injection modulator based on (a) ring resonator and (b) MZI.

length of 250 μm. The cross section is 450×220 nm and coupling gap is 200 nm. About 178 μm of the ring is built with a p - n modulator. The phase modulation is placed on the spare part of the ring, where the coupling is not affected. When a voltage is added across

Figure 4-13: Dynamic response experimental result for a MZI-based carrier injection modulator.

the p - n junction, the phase shift can be read out by the resonant peak of the ring. The phase modulator can be used as an amplitude modulator using a ring resonator or MZI structure. For the ring resonator, when the input wavelength is fixed at one of the resonant peaks of the ring resonator, the output power can be tuned by shifting the resonant peak. However, ring resonator-based amplitude modulators can only operate in some limited wavelength. For an MZI structure, the operating wavelength is much broader. Fig. 4-12b shows the fabricated carrier injection amplitude modulator based on MZI structure. The phase modulation is placed on both the MZI arm to provide a balanced optical path. Both of the structure could be designed as a p - n or p - i - n modulator, only depends on the doping process. Fig. 4-11b shows the testing result of a MZI-based amplitude modulator. The tested MZI has an arm length of 200 μm . The maximum and minimum output occurs at 0 V and 6.01 V bias voltage. The device is capable of operating at the whole C+L band.

Dynamic response for carrier injection modulator is also tested. An MZI-based amplitude modulator is used in the testing. A 1550-nm CW laser is input to one of the

input port of MZI. A 10-MHz square wave modulation signal is added on the *p-i-n* structure. Then the output light is measured by a photodiode connected to an oscilloscope. The result is shown in Fig. 4-13. The 90% switching time is 2.5 ns, which means that 100 MHz modulation is achievable. It is fully capable for our QKD application in 1-10 MHz band.

4.1.3 Electro-optic Modulator

Electro-optic effect is the change of optical properties of a material when the applied electric field changes. There are two major effects that can cause refractive index change, i.e., the Pockels effect and the Kerr effect. The Pockels effect is also known as the linear electro-optic effect, which indicated the refractive index change is proportional to the applied electric field. The Kerr effect is called the second-order electric field effect, which means the refractive index change is proportional to the square of the applied electric field. Due to the symmetry of the crystal structure, only the Kerr effect is present in silicon. However, the refractive index changes due to the Kerr effect is too small. In silicon, to get a refractive index change of 10^{-4} , the electric field required is about 100 V/ μm . Since the breakdown voltage for silicon is about 300 V/ μm , this modulation for silicon is impractical. To utilize the higher Pockels effect, some other material must be used. On the CMOS platform, luckily there is one material available, which is the aluminium nitride (AlN). The crystal structure of AlN is shown in Fig. 4-14. It has a 6mm hexagonal symmetry shape. The Pockels effect can be described by the equation of index ellipsoid as [176]

$$\left(\frac{1}{n_x^2} + r_{1k}E_k\right)x^2 + \left(\frac{1}{n_y^2} + r_{2k}E_k\right)y^2 + \left(\frac{1}{n_z^2} + r_{3k}E_k\right)z^2 + 2yzr_{4k}E_k + 2zxr_{5k}E_k + 2xyr_{6k}E_k = 1, \quad (4-4)$$

Figure 4-14: Schematic aluminium nitride crystal structure.

which is also known as optic indicatrix. The solutions for x , y and z form an ellipsoid. The magnitude of each vector (x , y , z) is the refractive index for light vibrating in this direction. In the equation, $k = 1, 2, 3$ corresponding to x -, y - and z -axis. n_x , n_y and n_z are the refractive indices in x -, y - and z -axis. r_{ij} are the electro-optic coefficients, which can be written in a matrix in the form of

$$\begin{pmatrix} r_{11} & r_{12} & r_{13} \\ r_{21} & r_{22} & r_{23} \\ r_{31} & r_{32} & r_{33} \\ r_{41} & r_{42} & r_{43} \\ r_{51} & r_{52} & r_{53} \\ r_{61} & r_{62} & r_{63} \end{pmatrix} \begin{pmatrix} E_x \\ E_y \\ E_z \end{pmatrix}. \quad (4-5)$$

For a 6mm symmetry crystal structure, most of the elements in the matrix are zero. The resulting electro-optic coefficient matrix is

$$\begin{pmatrix} 0 & 0 & r_{13} \\ 0 & 0 & r_{13} \\ 0 & 0 & r_{33} \\ 0 & r_{51} & 0 \\ r_{51} & 0 & 0 \\ 0 & 0 & 0 \end{pmatrix} \begin{pmatrix} E_x \\ E_y \\ E_z \end{pmatrix}. \quad (4-6)$$

For the three remaining elements, r_{51} is too small to measure in AlN. In order to use the higher elements r_{13} and r_{33} , electric field along the z -axis E_z is applied, which is the c -axis for crystal structure. Since the deposited AlN layer has a c -axis out-of-plane orientation, the TE and TM mode refractive index can be calculated as [177]

$$\begin{cases} n_{TE} = \frac{n_o}{\sqrt{1 + n_o^2 r_{13} E_z}} \approx n_o - \frac{1}{2} r_{13} n_o^3 E_z, \\ n_{TM} = \frac{n_e}{\sqrt{1 + n_e^2 r_{33} E_z}} \approx n_e - \frac{1}{2} r_{33} n_e^3 E_z, \end{cases} \quad (4-7a)$$

$$(4-7b)$$

where n_o and n_e is the refractive index for ordinary and extraordinary light without the presence of the electric field.

Due to the limitation for metal in the semiconductor foundry, it is hard to make a pair of parallel electrodes in the vertical direction. A ground-signal-ground (GSG) electrode configuration is used to introduce the electric field in the z -axis. As shown in Fig. 4-15a, the physical vapor deposited (PVD) aluminium nitride layer has a thickness of 400 nm. The fabricated waveguide has a trapezoidal shape with a corner angle of 70.7° . The thickness for the bottom and top oxide cladding is 2 μm and 1.5 μm , respectively. The center electrode is positively biased at 1 V, and the two side electrodes are grounded. The color map in the graph shows the electric field strength and the arrow indicated the electric field direction. The E_z component along the white dashed line is plotted separately in Fig. 4-15a. The result shows in the AlN layer, E_z is about 0.9×10^5 V/m with 1 V applied voltage.

To test the modulation performance, a ring resonator is built with AlN. As shown in Fig. 4-15b-d, the width of the waveguide is designed as 1.3 μm , while the fabricated width is 1.22 μm on top and 1.51 μm on the bottom. The coupling gap for the ring is 500 nm. The ring has a radius of 130 μm and the total length for the metal

Figure 4-15: (a) Simulated electric field distribution with GSG electrode. (b) Cross section TEM photo of the AlN waveguide with electrode on top. (c) AlN based ring resonator as modulator. (d) Enlarged TEM photo for the AlN waveguide.

electrode on top is $476 \mu\text{m}$. It is worth noticing that the deposited polycrystalline AlN has well-aligned grains in the out-of-plane direction, but the crystal growth in-plane is random. Thus, the material performs isotropic in-plane.

A DC power supply is connected to the electrode while TE or TM mode light is coupled to the ring resonator. As shown in Fig. 4-16, the Q-factor for TE and TM mode is 3.2×10^4 and 1.1×10^4 , respectively. When a voltage is applied to the electrodes, the

Figure 4-16: Resonance peak shift for (a) TE and (b) TM modes at different bias voltage.

peak shift can be observed. The peak shift is about 10 pm with 60 V bias voltage for TE mode and about 7 pm with 90 V bias voltage for TM mode. The peak shift is then plotted as a function of applied voltage, as shown in Fig. 4-17, which shows a good linear relationship between each other. Since the peak shift is directly related to the refractive index change, using Eq. (4-7), these values can be then used to calculate the

Figure 4-17: Peak shift as a function of applied voltage for TE and TM mode.

electro-optic coefficients. The refractive index change can be derived from the resonant condition Eq. (3-35) of the ring resonator. The calculated electro-optical coefficient $r_{33} = 0.83$ pm/V and $r_{13} = -0.44$ pm/V. As a comparison, some commonly used nonlinear crystal such as gallium arsenide (GaAs) has $r_{14} = 1.5$ pm/V and lithium niobate (LiNbO₃) has $r_{33} = 33$ pm/V. The performance of AlN is comparable with these materials.

Then the dynamic response of the AlN modulator is also tested. A CW laser with wavelength at the resonant peak of ring resonator is used as light source. The device is modulated under a 5 MHz square wave modulation signal with a 20 V peak to peak voltage. The output signal is measured by a photodetector connected to an oscilloscope, as shown in Fig. 4-18. The red dots are the original data from the oscilloscope, and the black curves are the smoothed data. The 90% switching time at 5 MHz modulation is measured to be 45 ns, which is acceptable for QKD applications.

One of the advantages of the electro-optic modulator is that the power consumption is very low since it is driving by an electric field rather than electric current or power. Here we also estimated the power consumption for our design. There

Figure 4-18: Performance of AlN modulator at 5 MHz square wave modulation.

are two main contributions, one is the DC power leakage and the other is the AC power consumed by the parasitic capacitor. The DC power leakage can be estimated from the insulation resistance between the two electrodes. The dielectric material between the two electrodes is silicon dioxide, which has a resistivity of about $10^{15} \Omega \cdot \text{m}$. Thus, the estimated electric power consumption with 60 V bias voltage is at the order of 10^{-16} W . The AC power consumption can be estimated by the power required for charging and discharging the parasitic capacitor. The parasitic capacitance is at the order of 10^{-15} F . With 60 V bias voltage, the energy stored in the capacitor is about 10^{-12} J/bit . Consider a modulation speed of 5 MHz, the power consumption is about 10^{-6} W .

To further optimize the performance of such AlN modulator, one important consideration is to reduce the driving voltage. Higher driving voltage leads to higher power consumption for same device configuration. Moreover, the electrical signal with both high voltage and high speed requires more effort to generate. To reduce the driving voltage, one way is to reduce the thickness of cladding oxide on top of the AlN waveguide. However, there is a trade-off between the electric field and the additional

Figure 4-19: The electric field distribution when the cladding material changes to aluminium oxide.

loss in the waveguide. More testing is required to get an optimum result. Another way is to boost up the electric field by changing the dielectric constant surrounding AlN waveguide. Current cladding material is silicon dioxide with a dielectric constant of 3.9. If we could change to aluminium oxide with a dielectric constant of 9.8, the resulting electric field distribution is shown in Fig. 4-19. The E_z component is about 1.7×10^4 V/m, which is nearly doubled compared to the case with silicon dioxide cladding.

4.2 Photodiode

The photodiode is another important active component, which could convert light into electric current. In this section, the chip-based germanium photodiode is introduced, designed and tested. The homodyne detector used for quantum information applications is also demonstrated based on the chip-based germanium photodiode.

4.2.1 On-chip germanium photodiode

Since our targeting wavelength is the communication wavelength at 1310 and 1550 nm, silicon has a low absorption coefficient in this band. Thus, germanium (Ge)

with smaller bandgap energy is introduced as the absorption material. It has a much higher absorption coefficient comparing to silicon. Also, Ge is a CMOS compatible material, which makes it possible to build an on-chip photodiode.

The Ge photodiode is designed with a vertical *p-i-n* structure, as shown in Fig. 4-20a. Ge is epitaxially grown on top of the silicon layer. The light from the silicon waveguide can evanescently couple to the Ge layer. A longer photodiode could increase the absorption and increase the final detection efficiency. The thickness for the Ge layer is 500 nm. The photodiode is designed to have a width of 8 μm and length of 25 μm . Due to the limitation for Ge epitaxy process, the resulting Ge layer has a frustum shape. A top view microscopic photo of the fabricated Ge photodetector is shown in Fig. 4-20b. Different layers are false-colored to be distinguishable. The green layer is the silicon layer. Light couples to the photodiode via an adiabatic taper. The red layer is the Ge on top of the silicon layer. The yellow and white layers are two layers of aluminium for electronic connection. The reason for using two layers is the requirements for separating front-end-of-line (FEOL) and back-end-of-line (BEOL) process.

The evanescent coupling between silicon and Ge layer are simulated using Lumerical FDTD software. As shown in Fig. 4-21a, the light is input from the bottom silicon layer and directly couples to the Ge layer from the edge of Ge. Because Ge has a higher refractive index of 4.28 at 1550 nm compare to silicon with refractive index of 3.47, the light could travel stably in this layer. Due to the absorption of Ge at 1550 nm wavelength, the energy dissipated is simulated to be about 80% at a transmission distance of 25 μm . The results confirm the design principle of evanescently coupled detector structure. The concentration of electrons and holes in the detector under different bias voltages are also simulated using Silvaco TCAD software. As shown in Fig. 4-21b, the intrinsic Ge region was simulated to be depleted entirely under -1 V bias

(a)

(b)

Figure 4-20: (a) Schematic of germanium photodiode cross-section. (b) Microscopic photo of the chip-based Ge photodiode.

voltage, which means when the electron and hole pairs are generated after photon absorption, the applied electric field could effectively sweep them towards the electrodes and collect as photocurrent.

Figure 4-21: (a) Simulation result shows the light coupled from silicon layer to germanium layer. (b) Photodiode electron and hole distribution at 0V and -1V reversed bias voltage.

Then the dark current and responsivity of the fabricated photodiode are calibrated, based on the experimental setup shown in Fig. 4-22a. The light source is a 1550-nm wavelength CW laser. After a polarization controller, the laser is coupled to the waveguide via a grating coupler. The light is then splitting to two paths via a 50 :

Figure 4-22: (a) Experiment setup for detector responsivity testing. (b) V-I curve of photodiode under different light power. (c) Responsivity over 0 to -4 V. In this region, the detector gives similar responsivity of about 700-800 mA/W.

50 beam splitter. One of the paths is used to assist alignment, from which the output power is measured by an off-chip detector. The other path is directly coupled to the on-chip photodiode. The electrical signal from the on-chip photodiode is amplified and collected via a data acquisition (DAQ) equipment. To measure the responsivity precisely, it is important to know the precise power coupled to the photodiode. The coupling loss at the grating coupler and waveguide transmission loss can be calculated using the cut-back method [175]. The transmitted power of a series of the waveguide with different length and the same coupler is tested. The linearly fit of the output power-waveguide length relationship gives the transmission loss of waveguide as slope and the y-intercept as the total transmitted power excluding waveguide loss. In our case, the total loss before the on-chip detector is calculated to be 12.7 dB. The I-V curve measured for on-chip photodiode under different input light power is shown in Fig. 4-22b. When no input light is presented, the curve shows the dark current, which is the noise level of the photodiode. At -1 V reverse bias voltage, the dark current is 2.2×10^{-7} A. The responsivity over the whole range of 0 to -4 V bias voltage is shown in Fig. 4-22c. In this region, the detector gives similar responsivity of about 700-800 mA/W. The results indicate the photodiode could operate at a wide range of bias voltage with similar performance. At -1 V bias voltage, the detector gives a responsivity of 734 mA/W.

The high-speed performance of Ge photodiode is also measured. Using a fiber-based amplitude modulator (Covega Mach-10), the input laser is modulated by a square wave signal. The output light is measured by the germanium detector and the output electrical signal is measured on a high-speed communication analyzer (Agilent Infinium DCA-86100A). As a comparison, the input light is first measured by a commercialized high-speed photodiode, the resulting eye diagram is shown in Fig.

Figure 4-23: Eye diagram of (a) reference commercialized photodiode and (b) on-chip Ge photodiode. (Time: 50 ps/div, Voltage: 10.2 mV/div)

4-23a. A clear eye opening is shown. The measured rising time is 23 ps, falling time is 19 ps and the extinction ratio is 12 dB. The same signal is then measured by the on-chip Ge photodiode, the result is shown in Fig. 4-23b. The eye-opening is also acceptable. Both the rising time and falling time are measured to be about 300 ps, which is good enough for our QKD application at 10 MHz level.

4.2.2 Balanced homodyne detector

The balanced homodyne detector is an essential device for quadrature measurement, which can be designed and integrated into the silicon photonic chip. The basic working principle of the homodyne detector on quadrature measurement will be covered in Section A2.2. In this section, the design and performance of the Ge based homodyne detector will be discussed. Fig. 4-24a shows the schematic of a homodyne detector. It consists of a 50 : 50 beam splitter (BS) and two photodiodes (PD). The signal light is a weak light modulated in both the quadratures, while the local oscillator (LO) light is a stronger light used as phase reference and signal amplification. The electrical signals from two PDs i_1 and i_2 are subtracted and amplified using a transimpedance amplifier. From a classical perspective, the result is $i_1 - i_2 = 0$, because the 50 : 50 splitter splits the light into two identical paths. Thus, all the fluctuation contained in the LO light would be well balanced and cancel out. From the quantum point of view, the electrical signal i_1 and i_2 still contain shot noise because of the random Poissonian statistics of the photon. Due to the random statistic, the resulting $i_1 - i_2$ would add up the shot noise. In our chip, as shown in Fig. 4-24b, a two-dimensional coupler, a 50 : 50 beam splitter (BS) and two photodiodes (PD) are integrated.

The first consideration for the homodyne detector is the ability to balance the two input signals and giving average $i_1 - i_2 = 0$, which is measured by introducing a sine noise on the LO beam deliberately and measuring the output with and without the balance subtraction. Here, the LO beam is modulated with 7 MHz sine wave. The output spectrum is shown in Fig. 4-25a. The two curves show the balanced and unbalanced results, which is achieved by switching one of the PDs in on- or off-state respectively. The RMS has 25 dB difference in these two states, which indicated the homodyne

Figure 4-24: (a) Schematic basic working principle of homodyne detector. (b) Microscopic photo of a chip-based homodyne detector.

detector is capable of reducing the classical noise in LO beam by 25 dB. The value is also known as the common mode rejection ratio of the homodyne detector.

Another requirement for the homodyne detector is the ability to detect the shot noise of LO light. The transimpedance amplifier would convert the current signal from PD to a voltage signal easier to collect. Since the shot noise signal amplitude is very low down to nanoampere level, a high gain amplifier with more than 10^4 V/A transimpedance gain is required to amplify the signal and provide a readout higher than the amplifier induced electronic noise. Then the second stage voltage amplifier

Figure 4-25: (a) PD signal balance test when LO is modulated at 7 MHz and no signal is presented. (b) Transimpedance amplifier design for homodyne detector.

amplifies the signal together with the electronic background noise to get a readable output. Fig. 4-25b shows the schematic design of the transimpedance amplifier. The operational amplifier (op-amp) chip is selected to be low noise and high-bandwidth. The first stage uses an ADA4817 op-amp chip with 3 dB bandwidth of 1.05 GHz and low noise of $4 \text{ nV}/\sqrt{\text{Hz}}$. Moreover, the second stage uses an OPA847 op-amp chip with

Figure 4-26: Total noise and fitted shot noise with different LO power level at the (a) 1 MHz band and (b) 3 MHz band.

gain-bandwidth product of 3.9 GHz and low noise of $0.85 \text{ nV}/\sqrt{\text{Hz}}$. The first stage is designed to have a 10^4 V/A transimpedance gain. Then the AC component passing through a second stage amplifier with 60 V/V gain and provides the output.

The total noise of the homodyne detector is measured as a function of LO power at 1 MHz and 3 MHz bands, as shown in Fig. 4-26a and Fig. 4-26b, respectively. Since

the shot noise amplitude has a linear relationship with the LO power. The measured total noise is increasing with the LO power. The total noise can be decomposed into two parts, which is the static electronic noise and the linear increasing shot noise. The yellow region marked the static electronic noise when no light is present, And the fitted shot noise in red has a linear relationship with LO power. When LO power is higher than 11 dBm, the shot noise is at least 5 dB higher than electronic background noise, which offers us 5 dB signal to noise ratio for shot noise measurement.

4.3 Fabrication Process

The fabrication process of the active components is an extension for the passive processes. As shown in Fig. 4-27, The main differences are the newly introduced doping process and metal layer processes.

Before introducing the fabrication process in detail, the doping parameters are analyzed first. Since doping is the process for implanting ions into the substrate material by accelerating ions to high energy up to MeV level, the result of doping is affected by four major parameters, which are the dopant dose, ion energy, screen oxide thickness and thermal annealing process. There are two dopants involved in our process, which are phosphorous for *n*-type doping and boron for *p*-type doping. The doping process is simulated by Silvaco TCAD software with a one-dimensional simulator.

Dopant dose is the number of ions per unit area. It is intuitive to understand that the final dopant concentration would be higher when dopant dose is higher. In this simulation, the ion energy is kept constant and the thermal annealing process is applied. Fig. 4-28 shows the dopant concentration distribution at 1×10^{13} , 1×10^{14} , 1×10^{15} and 4×10^{15} ions/cm², where 4×10^{15} ions/cm² is the maximum dose available with our equipment. The target wafer is SOI wafer with 220 nm crystal silicon device layer and

Figure 4-27: Fabrication process for the silicon photonics active components.

Figure 4-28: Dopant concentration distribution at different dopant dose (ions/ cm^2) for (a) phosphorous and (b) boron.

70 nm screen oxide on top. The region for silicon device layer is marked with a gray dashed line. The top surface of silicon device layer is defined as the zero position. The screen oxide is on the negative side of the silicon region and the bottom oxide is on the positive side. The ion beam comes from the negative side. The result shows a roughly linear relationship between the final concentration inside the silicon layer and the applied dopant dose. The highest possible doping concentration for phosphorus and

boron is $3 \times 10^{19} \text{ cm}^{-3}$ and $1 \times 10^{20} \text{ cm}^{-3}$, respectively. Although the structure of devices may also affect the concentration distribution, Fig. 4-28 is good enough to provide an approximation for the dopant dose selection. It is also worth noticing the peak concentration appears in the screen oxide layer, which indicates most of the dopants are blocked by screen oxide. The doping parameters here still have space for optimization.

Another important parameter is the ion energy. The ion energy together with the ion mass could determine how deep the ion could travel into the wafer. The simulation here uses $4 \times 10^{15} \text{ ions/cm}^2$ dose and different ion energies. Since phosphorus has high mass than boron, higher energy is expected for phosphorus to achieve a similar doping result with boron. Thus, 20 to 80 keV is calculated for phosphorus and 10 to 40 keV is used for boron. As shown in Fig. 4-29, the higher the ion energy, the deeper the ions can travel. At 20 keV for phosphorus and 10 keV for boron, most of the ions are blocked at the screen oxide layer. Less than 1% of the dopants could travel to the silicon layer. Moreover, at 80 keV for phosphorus and 40 keV for boron, the peak concentration is inside the silicon layer. The result also shows that for a lighter ion like boron, the peak position is deeper than phosphorus even with half the energy. Thus, the doping parameters must be optimized separately depends on the dopant properties.

There are two major functions of screen oxide. One is to protect the silicon layer from low energy ion debris and surface damage. The other is to reduce the channeling effect and provide a more controllable doping profile. The effect of screen oxide is shown in Fig. 4-30. The most natural effect is that it absorbs some energy from the coming ions and reduced the penetration depth. Since our wafer has a 70 nm hard mask oxide on top of the silicon layer, it will be used directly as the screen oxide in the doping process. Comparing to a commonly used thickness of 10 to 20 nm for screen oxide, the

Figure 4-29: Dopant concentration distribution at different ion energy for (a) phosphorous and (b) boron.

doping concentration decrease to 20% for phosphorus and 30% for boron. The decreasing is acceptable in our case, so no further process is introduced to remove the hard mask oxide.

Thermal annealing process is required because the high energy ions could destroy the crystal structure of silicon. Thermal annealing could reconstruct the crystal

Figure 4-30: Dopant concentration distribution with different screen oxide thickness for (a) phosphorous and (b) boron.

structure and activate the dopants. To maintain the doping profile, a short anneal time with very high temperature is advantageous, which is called the rapid thermal annealing (RTA). In our process, the RTA time is 5s at 1050 °C. Simulation is performed for four doping doses of phosphorous, which is 1×10^{13} , 1×10^{14} , 1×10^{15} and 4×10^{15} ions/ cm^2 . As shown in Fig. 4-31, the most significant changing occurs at the temperature ramp-up region. After the RTA and ramp down, the distribution won't change much. Another

Figure 4-31: Dopant concentration distribution before and after the thermal annealing at different dopant dose.

phenomenon is that the final distribution tends to be uniform for a higher dose, or from another point of view the dopants could move more freely at a higher dose. Thus, if the final doping profile is critical, a careful control of the RTA process is required.

Back to the fabrication process, the first two steps pattern the whole device layer with waveguide, grating and rib structures. Then step 3 is the doping process. The doping processes use $1\text{-}\mu\text{m}$ thickness photoresist as the mask. For modulator, carrier injection modulator we designed only requires a p^{++} and an n^{++} doping. Boron with about 4×10^{15} ions/ cm^2 dose is used for p^{++} doping, while n^{++} doping uses phosphorus also with about 4×10^{15} ions/ cm^2 dose. For detector, the silicon side requires a p^+ and a p^{++} doping. The p^{++} parameter is the same with the modulator. The p^+ uses boron with about 1×10^{14} ions/ cm^2 dose. Finally, after the three doping steps, the whole wafer is sent for the RTA process for 5s at 1050°C .

In step 4, the hard mask or screen oxide is removed by DHF. Then 120 nm thickness of USG oxide is deposited on top of the whole wafer, which is used as the passivation layer for Ge growth. After opening a window on the photoresist, A combination of dry and wet etching is used to open the area for Ge growth while maintaining the surface quality of silicon.

Step 5 is the step for Ge growth and doping. First, a silicon-germanium buffer layer with about 20 nm thickness is deposited, followed by a Ge seed layer with about 30 nm thickness. Then the Ge is epitaxially grown to a thickness of 500 nm. The top doping for Ge detector uses phosphorus with about 1×10^{15} ions/cm² dose. Finally, the wafer is annealed at 500 °C for 300s. Fig. 4-32a shows the SEM photo of the grown Ge on top of silicon. The Ge structure is false-colored in orange, which has a frustum shape as stated in the last section.

In step 6, the first cladding oxide layer is deposited. Similar to the passive process, 200 nm of USG oxide plus 400 nm HDP-CVD oxide is deposited. Then, the via hole for electrode connection is etched through the whole oxide cladding layer.

Then the first layer of metal is patterned in step 7. Two layers of metal are used due to the restriction for metal contamination control in CMOS foundry. After any metal process, the wafer is not allowed going back to the non-metal tool and the silicon device layer is not allowed to expose again. The metal used here is a 750-nm thickness aluminium (Al) deposited by physical vapor deposition (PVD). An additional layer of 50-nm tantalum nitride (TaN) is used as the adhesion layer between Al and silicon or oxide. Al etching is performed on the same machine with Al₂O₃ by RIE. One consideration for Al processes is that Al is very reactive under extreme pH conditions. However, the developer of photoresist often contains alkaline, and some side products,

(a)

(a)

Figure 4-32: (a) SEM photo of the epitaxial germanium grown on top of silicon layer. (b) SEM photo of metal layer after etching.

Figure 4-33: SEM photo of first metal layer on the photodiode.

such as aluminium chloride, could produce acid during the Al RIE etching process. Thus, it is essential to take some precaution when etching Al. In our process, a 30-nm thickness silicon nitride (SiN) is used as the protection layer on Al to avoid direct contact with photoresist developer. Also, process priority is given to such wafer to minimizing the overall processing time. After the Al etching, the photoresist is removed first, and the wafer is cleaned using ultrasonic. The SiN protection layer is left as the etching stop layer if another via hole is required on top. The Al etching result is shown in Fig. 4-32b. The Al layer is sandwiched between two layers of TaN. Slight erosion of Al on edge can be seen clearly. In Fig. 4-33, the SEM photo shows the first metal on the photodiode structure, in which the dent caused by via hole can be seen clearly.

Titanium nitride (TiN) heater pattern is in step 8. Titanium nitride is used as heater due to its good electrical conductivity. Before depositing the TiN layer, another

(a)

(b)

Figure 4-34: (a) SEM photo of the second via hole opened on Al. (b) SEM photo of a heater structure patterned by the second layer of metal.

Figure 4-35: TEM photo shown the cross section of (a) modulator and (b) photodiode.

cladding oxide layer with a 1.5 μm thickness is first deposited to cover the first metal layer. Chemical-mechanical planarization (CMP) is then performed to planarize the surface. Next, 120 nm thick TiN layer is deposited by PVD. Similar to the Al patterning process, the surface of TiN is also covered with a 30-nm SiN as protection layer and etching stop.

Then in step 9, the second layer of 2- μm thickness Al is patterned. First, 500 nm of cladding oxide is deposited to cover the TiN heater. CMP is performed again to planarize the surface. Then via hole is etched on the cladding oxide. This time, the etching of oxide is stopped on the SiN layer. After the oxide etching, the SiN is removed to expose the surface of Al and TiN, as shown in Fig. 4-34a. Finally, the second metal layer is patterned using the same process as stated in step 7. Fig. 4-34b shows an SEM photo of a heater structure patterned with the second layer of Al.

Finally, in step 10, the wafer is prepared for dicing using the similar process for passive components. The trenches are patterned for sidewall protection and the whole

chip surface is covered with Al_2O_3 . The window is patterned on top of the metal pad for electrical connection.

After finishing the whole process, some chips are cut by the focused ion beam (FIB) for the cross-section transmission electron microscope (TEM) check. Two photos for the modulator and photodiode are shown in Fig. 4-35.

4.4 Summary

This chapter focused on the design, simulation, fabrication and experiments of various silicon photonic active components. In section 4.1 three major types of the modulator are covered, namely thermo-optic modulator, carrier injection or depletion modulator and electro-optic modulator. The result shows that at a same applied voltage, the phase shift for the thermo-optic modulator is the highest. With careful design, for a 100- μm length modulator, 2π phase shift could achieve under 2V applied voltage, which is impossible for the other two types. Thus, if compact size is the main requirement, the thermo-optic modulator is the one to choose. Consider the modulation frequency, both carrier injection type and electro-optic type could achieve above GHz level, which is much higher compare to the thermo-optic modulator with kHz level speed. Moreover, the electro-optic modulator based on AlN has some additional bonus. Not only it has high second-order nonlinearity, but also it can achieve ultra-low power modulation, while sacrificing some fabrication cost. Thus, the analysis in this section would give a guideline for choosing the different types of modulators. In section 4.2 the chip-based photodiode based on Ge is designed and tested. The experimental result shows the responsivity for the Ge PD is about 0.73 A/W at 1550 nm wavelength, the operation bandwidth is more than 2.4 GHz and the dark current is down to nanoampere level. The result proved the function of chip-based Ge PD for

quantum information applications. Subsequently, a balanced homodyne detector based on Ge PD is designed, which is one of the key contributions of this thesis and one of the core elements for continuous variable quantum information. The testing shows the common mode rejection ratio is more than 25 dB and the working bandwidth is 1-10 MHz. Finally, in Section 4.3, the fabrication process for all the active components are summarized in detail.

Chapter 5 Chip-based Continuous Variable Quantum Key Distribution

In this chapter, the chip for continuous variable quantum key distribution (CV-QKD) is designed, fabricated, and experimentally demonstrated for the first time. All the essential components including germanium photodetectors are integrated onto a single chip using nano-silicon photonic fabrication process, which could significantly reduce the size and price of current QKD systems. In addition, quantum optics theory and information theory related to CV-QKD are provided, including the essential post-processing protocols. Then the chip based QKD system is demonstrated theoretically and experimentally. The experimental results showed a secure key rate of 0.83 MHz at 0 km and 2.3 kHz at 100 km transmission distance under the collective attack assumption.

5.1 Information reconciliation and privacy amplification

For a QKD system, quantum mechanics ensures that no one could eavesdrop any information without being noticed. The post-processing, including error correction and privacy amplification, ensures the security of information when an eavesdropper is detected. Fig. 5-1 schematically shows the information possessed by each party during a post-processing process. Assume Alice (A) are trying to distribute a secure key series to Bob (B). The amount of information possessed by Alice is 1. Eve (E) is trying to eavesdrop the information in the center. After a key distribution process, the key they received before any post-processing is called the sifted key. Now, Bob could receive parts of the information due to the channel loss and the disturbance from Eve, which is marked in yellow column. Eve could also get some information from the process, which is marked in blue. One thing to notice is that the information possessed by Bob here should be more than Eve to get a reliable secure key, which is one of the limitations for QKD using classical post-processing techniques. The information difference between Bob and Eve is marked by δ_0 . Moreover, the goal for QKD is to share an identical key between Alice and Bob, while Eve gets no information at all. Thus, post-processing is required to achieve this goal. During error correction phase, Bob and Alice would announce some information via a public channel to make their key identical, but Eve is able to receive the same amount of information. After this phase, Bob's information would become 1 while Eve's information would increase by a slightly higher amount than Bob's increment. The information difference becomes $\delta_1 < \delta_0$. Next, to erase everything known to Eve, privacy amplification is performed by publicly announce some more information and selectively discard part of the information. After this step, Alice and Bob would share a new shorter series of secure key, while Eve's knowledge on the key becomes 0.

Figure 5-1: Information possessed by each party during post processing processes, including the error correction and privacy amplification.

5.1.1 Sliced error correction

Sliced error correction is a reconciliation protocol for CV-QKD, which could convert the continuous variable keys into binary bits [178-180]. The continuous variable key series X sent by Alice follows a Gaussian probability with $X \sim N(0, \Sigma^2)$, while the continuous variable key series X' received by Bob would have some additional shot noise with $X' = X + \varepsilon$, where $\varepsilon \sim N(0, \sigma^2)$. For a forward reconciliation process, Bob would try to guess the key X sent by Alice based on the key X' he received. For a reverse reconciliation process, Alice would try to guess the key X' received by Bob based on the key X she sent. Although the security level is different in these two

cases, the basic working principle is the same. Here we will introduce reverse reconciliation protocol.

Bob would use a slice function $S(x')$ to map one of his key x' into an m bits binary string $(S_1(x'), S_2(x'), \dots, S_m(x'))$. The domain for Bob's key would be divided into $t = 2^m$ intervals by $\tau_0, \dots, \tau_t$, where $\tau_0 = -\infty$ and $\tau_t = \infty$. Each interval a with $(x': \tau_a \leq x' < \tau_{a+1})$ is mapped to a m bits binary string $binary(a)$. This mapping is the basic working principle of slice function $S(x')$. By figuring out which interval a does x' lie in, the resulting binary string $(S_1(x'), S_2(x'), \dots, S_m(x'))$ would be $binary(a)$. Fig. 5-2a show an example of the slice function with $m = 3$. If Bob has a continuous variable key x' between τ_4 and τ_5 , the resulting binary string $S(x)$ would be $binary(4) = (1, 0, 0)$.

Alice on the other side, would try to guess Bob's binary string $S(x')$ based on her key x , using the slice estimator $\hat{S}$. The output from slice estimator is a binary string given by

$$\left(\hat{S}_1(x), \hat{S}_2(x, \hat{S}_1(x)), \dots, \hat{S}_m(x, \hat{S}_1(x), \dots, \hat{S}_{m-1}(x)) \right). \quad (5-1)$$

The digits are calculated one by one with a recursion. The estimation for next digits is based on not only the key x Alice owns but also the estimation of all the previous digits.

The slice estimator for digit i is defined as

$$\hat{S}_i(x, b) = \arg \min_{\hat{s}} \Pr(S_i(X') = \hat{s} | S_{1 \dots i-1}(X') = b, X = x), \hat{s} \in \{0, 1\}. \quad (5-2)$$

She then calculates the two probabilities P_0 and P_1 with $\hat{s} = 0$ and $\hat{s} = 1$, respectively. If P_0 is larger, the estimation for digit i is 0. Similarly, if P_1 is larger, the estimation for digit i is 1. As an intuitive example, the calculation of slice estimator $\hat{S}_3$ for digit 3 when $m = 3$ is shown in Fig. 5-2b. Assume Alice already knows the last 2 digits of Bob's

Figure 5-2: (a) Working principle of the slice function $S(x)$. (b) Working principle of the slice estimator $\hat{S}(x)$.

binary bits $(S_1(x'), S_2(x'))$ as $(1, 0)$, she then calculates the conditional probability P_0 and P_1 as

$$\left\{ \begin{array}{l} P_0 = \int_{\tau_0}^{\tau_8} \int_{\tau_2}^{\tau_3} p(x, x') dx dx' , \\ P_1 = \int_{\tau_0}^{\tau_8} \int_{\tau_6}^{\tau_7} p(x, x') dx dx' . \end{array} \right. \quad (5-3a)$$

$$\left\{ \begin{array}{l} P_0 = \int_{\tau_0}^{\tau_8} \int_{\tau_2}^{\tau_3} p(x, x') dx dx' , \\ P_1 = \int_{\tau_0}^{\tau_8} \int_{\tau_6}^{\tau_7} p(x, x') dx dx' . \end{array} \right. \quad (5-37b)$$

Table 5-1: The interval boundary values set for slice error correction.

τ_{16}	$\tau_{17}, -\tau_{15}$	$\tau_{18}, -\tau_{14}$	$\tau_{19}, -\tau_{13}$	$\tau_{20}, -\tau_{12}$	$\tau_{21}, -\tau_{11}$	$\tau_{22}, -\tau_{10}$	$\tau_{23}, -\tau_9$
0	0.14	0.28	0.421	0.567	0.725	0.885	1.05
$\tau_{24}, -\tau_8$	$\tau_{25}, -\tau_7$	$\tau_{26}, -\tau_6$	$\tau_{27}, -\tau_5$	$\tau_{28}, -\tau_4$	$\tau_{29}, -\tau_3$	$\tau_{30}, -\tau_2$	$\tau_{31}, -\tau_1$
1.22	1.39	1.56	1.73	1.9	2.15	2.5	3.1

The probabilities collapse to the shadow areas in Fig. 5-2b when the two lower digits are known. The probability of giving the next digit as 0 is higher than the probability of giving 1. Thus, the slice estimator returns a result 0.

Since the X and X' are both gaussian variables, the conditional probability in an analytic form can be given as

$$P(a \leq X' \leq b, X = x') = \frac{1}{2\pi\Sigma\sigma} \int_a^b \exp\left(-\frac{x'^2}{2\Sigma^2} - \frac{(x - x')^2}{2\sigma^2}\right) dx' . \quad (5-4)$$

The interval defined by $\tau_0, \dots, \tau_i$ could largely affect the reconciliation efficiency. Optimization is performed on the value of intervals. The maximum net amount of information is around 2 for $m = 5$ and $\text{SNR} = 15$. The optimized interval set shown in Table 5-1 could provide practical net amount of information for about 1.973.

The sliced error correction is tested using two Python software mimicking Alice and Bob, respectively. Alice first generates a series of 5000 Gaussian random variables with variance $\Sigma^2 = 1$. Then she adds a Gaussian noise with $\sigma^2 = 1/15$ on it and sends to Bob via a simulated quantum channel. After Bob receives the Gaussian key, they apply the slice error correction protocol. Bob disclose all 2 lower bits from his slice function to Alice, and Alice uses them to do the slice estimation for the 3 higher bits. Part of the results are shown in Table 5-2. The disclosed bits during error correction is mark in red. Finally, after the whole error correction process, the shared binary keys with Alice and Bob are compared. The result shows a total 943 bits of error in 25000 bits of binary key.

Table 5-2: Sliced error correction testing result

Alice Gaussian key	Alice binary key	Bob Gaussian key	Bob binary key
-1.243506	010 10	-1.057794	010 10
0.408446	100 10	0.528386	100 10
-1.828195	010 00	-1.563723	010 00
0.498048	100 10	0.956080	100 10
-1.881240	010 00	-1.804208	010 00
...	...	...	...

The error rate is 3.77%. Among the 25000 bits of binary key, the practical net amount of secure information is $1.973 \times 5000 = 9865$ bits.

5.1.2 Binary error correction

As shown from the testing result in the last section, the binary keys generated from sliced error correction still contains some bit errors. To get an identical key series, a binary error correction must be introduced to correct the errors, while minimizing the publicly disclosed information. The Cascade error correction protocol is one of the efficient protocols specially designed for this job [181]. Cascade uses parity check to find the error present while limiting the information disclosed to only a few parity bits. The detail will be discussed below.

Alice and Bob first rearrange their key strings using a same random permutation, which would evenly distribute the errors presented. Then, they divide their new key strings into small sub-blocks B_j with same length w . The parities of every sub-blocks are exchanged and compared between Alice and Bob. If the parity of a sub-block is the same for both party, it means there is an even number of errors in this sub-block. Similarly, if the parity is different, there is an odd number of errors. In the case when

odd number of error is detected, bisection searching is used to find the exact position of error. Alice and Bob compare the parity of half the sub-block. If the parity is different, they keep search this half with bisection. Otherwise, they search the other half. The searching goes until they located the single bit which is wrong. Bob could correct it by simply invert the bit.

The process could simply repeat multiple times to get a reliable correction, which is the protocol named BBBSS (named after its author Bennett, Bessette, Brassard, Salvail and Smolin). Cascade is an improved version of BBBSS with higher efficiency. It uses four iterations. Different from the simple BBBSS protocol, Cascade tracks all the previous calculated sub-block. Starting from the second iteration, when an error bit is corrected, all the previous calculated sub-block contains this error bits would have a parity flip, which would introduce some new parity difference between Alice and Bob. Subsequently, they correct the new detected error using the same bisection technique. This process is repeated until all known parity difference is solved, which ends the iteration step. After four iterations, the probability for correcting all the possible errors is reasonably large.

The block size w plays a very important role in this protocol. It is chosen such that minimum parity bits are revealed and each sub-block contains no more than one error with reasonably high probability. In the original version of Cascade, the block size for each iteration is optimized to be $w_1 = 0.73/e$ and $w_i = 2w_{i-1}$, where e is the estimated error rate. The number of parity bits r disclosed is given by

$$r \approx l(1.1 + e)h(e), \quad (5-5)$$

where l is the total data size and $h(e)$ is the entropy of a binary random variable defined as

$$h(e) = -e \log e - (1 - e) \log(1 - e) . \quad (5-6)$$

For the error rate of $e = 0.0377$ from the sliced error correction, the calculated $h(e) = 0.2316$ and $r/l = 0.2635$, which means the disclosed bits in this process is about $26.35\% \times 25000 = 6588$ bits. The remaining secure key is $9865 - 6588 = 3277$ bits.

5.1.3 Privacy amplification

At this point, Alice and Bob already get an identical key string, but it is not secure enough. Eve got to know part of the information. Thus, Alice and Bob need to squeeze the information known to Eve based on their information advantage over Eve, which can be understood as discarding the part of information known to Eve. This process is known as privacy amplification.

One of the common ways to do this is applying a hash function. Hash function is the function $f(A) = B$ that maps data of arbitrary size A to a data of fixed size B . There are a few properties for hash function that is very important in cryptography. (1) It is very easy to calculate B with known f and A . (2) It is very hard to calculate A with known f and B . (3) It is also very hard to calculate B with incomplete information on f and A . The three properties ensure when Alice and Bob apply the same hash function on their data, the output would be very hard to predict by Eve.

Toeplitz matrices is one of the simple way to generate a hash function available for this job. The Toeplitz matrices is defined as $M_{(i,j)} = a_{i-j}$, where a_i is a series of random value with $a_i \in \{0,1\}$, $M_{(i,j)}$ is the (i,j) element of matrix M . In an intuitive way, the Toeplitz matrices can be expressed as [182]

$$M = \begin{pmatrix} a_0 & a_{-1} & a_{-2} & \cdots & a_{-j+1} \\ a_1 & a_0 & a_{-1} & & \\ a_2 & a_1 & & & \vdots \\ \vdots & & & \ddots & \\ a_{i-1} & & \cdots & & a_{i-j} \end{pmatrix}. \quad (5-7)$$

For a privacy amplification process, Alice and Bob needs to convert their identical key with length n to a secure key with length l , which could achieve by multiply a Toeplitz matrix with a size of $n \times l$. In this way, the additional random bits required for generating Toeplitz matrix is $(n + l - 1)$ bits. In the previous example, the secure keys are 3277 bits among a raw key of 25000 bits long. The Toeplitz matrix used would have a size of 25000×3277 . The resulting 3277 bits are secure against eavesdrop.

5.2 Chip-based CV-QKD

In this section, a single chip solution of CV-QKD system with GG02 coherent-state protocol is demonstrated. Compared to the single photon QKD, the use of coherent state for information encoding makes the fully integrated chip possible. The transmitter chip can generate Gaussian-modulated coherent state signal and the receiver chip can do homodyne detection on the transmitted states. The successful demonstration may lead to a longer and more secure transmission. It poses high potential in the hybrid quantum-classical communication network using fiber and free-space optics.

5.2.1 GG02 protocol

As shown in Fig. 5-3, a silicon photonic chip is designed as the transmitter (Alice). 1550-nm continuous wave (CW) laser die can be bonded to the chip through the die bonding process. Then, light is coupled into the waveguide with a grating

Figure 5-3: Schematic of the CV-QKD system. Two chips are used as the transmitter and receiver.

coupler. The first modulator is used as an attenuator in this experiment and is reserved for generating pulses in our future works. A 1:99 directional coupler splits the input laser into two paths, the weaker one as signal path and the stronger one as the local oscillator (LO). The signal path is modulated with an amplitude modulator (AM) and a phase modulator (PM) to generate a series of coherent state $|x_A + ip_A\rangle$, where x_A and p_A are random numbers with a Gaussian distribution. The mean is zero and variance is $V_A N_0$, where V_A is the modulation variance with the unit of shot-noise-unit (SNU), and N_0 is the shot noise variance. The information is encoded on the 1-10 MHz sideband frequencies of the signal light without the requirement of a pulsed laser [183, 184]. In this way, the low-frequency noise ($<1\text{MHz}$) in the system are bypassed. A filter is then required to demodulate the information from one of the sideband frequencies. To keep the relative phase between the signal path and LO path after transmission, the modulated signal and LO are multiplexed as two orthogonal polarization states with the

two-dimensional grating coupler. After the transmission line with transmittance T , the receiver (Bob) first compensates the polarization drift with a polarization controller. Then another two-dimensional grating coupler is used as the demultiplexer for signal and LO. Here, because the information is encoded on the sideband frequencies, which is the AC component of signal light, a 35-dB extinction ratio is good enough to separate signal and LO light. There is no need to keep an ultra-high (60-80 dB) intensity difference between signal and LO like the protocol shown in the previous works [37, 38]. Finally, Bob randomly chooses to measure quadrature x or p with the homodyne detector and filters out the required frequency. Alice and Bob can then share a set of correlated Gaussian data. The reverse reconciliation is then performed and followed by error correction and privacy amplification. Alice and Bob can share a set of secure keys after that. The secure keys can be used to encrypt secret messages with unconditional security. Therefore, two chips as Alice (transmitter) and Bob (receiver) can be designed and fabricated. The transmitter chip encoded information in x and p quadrature of the signal light. Then the receiver chip measures one of the quadratures.

5.2.2 Transmitter chip

As shown in Fig. 5-4a, the transmitter is fabricated on 220-nm silicon-on-insulator (SOI) platform. Three consecutive lithography steps and reactive-ion etching processes formed the silicon waveguide, grating coupler and rib structures. The waveguide has a width of 450 nm. It was then followed by boron and phosphorous doping steps, with an impurities concentration of approximately 10^{20} cm^{-3} . After covering the silicon structures with 2- μm thickness silicon dioxide (SiO_2), the via hole structures are etched on the SiO_2 layer for electrical connection. Then aluminum of 750- μm thickness is deposited and patterned as the conduction layer, which acts as the

(a)

(b)

Figure 5-4: (a) Microscopic photo of the transmitter chip. (b) Photo of the wire bonded chip for testing.

electrical connections for active control. The metal layer is false-colored in yellow and the waveguide layer is colored in red. The modulator here is based on the plasma dispersion effect and working at carrier-injection mode. The amplitude modulator is designed as a Mach-Zehnder Interferometer (MZI) structure, where the two arms are modulated by carrier-injection phase modulator. The region near the waveguide is heavily doped. When an electric field is added to the doped region, large amount of free

carrier can be injected into the waveguide. The refractive index of the waveguide is then affected by the concentration of free carrier, which can further change the phase or amplitude of light after MZI structure. The chip is connected to a printed circuit board (PCB) board uses wire bonding technique. The PCB board is used as an electronic interface for testing, as shown in Fig. 5-4b.

As shown in last chapter, the system is fully capable of operating at 1-10 MHz band. To modulate the signal precisely, the cross-modulation of the AM and PM is measured to confirm that the two modulators cannot affect each other. In the testing, the source is a 1550-nm laser with 12 dBm power from a tunable laser (Santec TSL-510). After a polarization controller, the laser is coupled to the transmitter chip. A sine wave of 5-MHz frequency is added on AM and 6-MHz is added on PM using arbitrary waveform generator (HP 33120A). The output signal spectrum from the homodyne detector is monitored on an oscilloscope (Tektronix MDO4104B-3), while the LO phase is scanned by another PM. The result in Fig. 5-5a shows the power root-mean-square (RMS) at 5 MHz and 6 MHz, which corresponded to the AM and PM modulation signal. Difference between these two signals is shown in Fig. 5-5b. The difference between the maximum and minimum in Fig. 5-5b is more than 40 dB, which means the AM signal and PM signal has nearly no crosstalk between each other and can precisely reflect the modulation signal added. The maximum and minimum points are also used to choose the measured quadrature. The difference is measured before and after each measurement to confirm that there is no significant signal drift. Fig. 5-5c shows the spectrum from 4.5 to 6.5 MHz at 1.1 V and 2 V phase modulation power respectively, which correspond to the maximum and minimum point marked with dash lines in Fig. 5-5a and Fig. 5-5b. After modulation, 2D grating coupler is used to

Figure 5-5: (a) Cross-modulation measurements of AM and PM. (b) Difference between AM and PM signals (c) Measured frequency spectrum at maximum and minimum cross-modulation difference

multiplex signal and LO light in polarization. The coupling loss for grating coupler structure is about 4 dB.

5.2.3 Receiver design

The receiver is a homodyne detector with a PM to select the measuring quadrature. The Ge photodiode introduced in Section 4.2.2 is perfect for this job. For the first step, a fiber coupled off-chip indium gallium arsenide (InGaAs) photodiode is used to characterize the transmitter chip. The off-chip receiver is assembled using all fiber components. As shown in Fig. 5-6a A fiber-based polarization beam splitter (Thorlabs PBC1550SM-FC) replaces the 2D grating coupler. Then the two paths are connected to a polarization maintenance fiber 50:50 beam splitter (Thorlabs PN1550R5F2). The two output ports are finally connected to the fiber coupled photodiode (Lightsensing LSIPD-LD50).

Figure 5-6: (a) Schematic of the off-chip homodyne detector configuration. (b) PD signal balance test when LO is modulated at 5 MHz and no signal is presented.

Figure 5-7: (a) The noise spectrum at different input power. (b) Total noise and fitted shot noise with different LO power level at the 3 MHz band.

For the common mode rejection ratio measurement, Fig. 5-6b shows the balance testing result when LO beam is modulated with 5 MHz sine wave. The two curves show the balanced and unbalanced results, which is achieved by switching one of the PDs in on- or off-state respectively. For the shot noise measurement, since the off-chip detector introduce much less noise. The first stage transimpedance amplifier is changed to have

a 10^5 V/A transimpedance gain. The output noise spectrum at different input LO power is shown in Fig. 5-7a. With the increase of input LO power, the noise level also increases. The noise at 3 MHz band is selected and shown in Fig. 5-7b. The black curve is the total measured noise. The yellow region is the electronic background noise level. The fitted shot noise in red curve has a good linear relationship with LO power. When the LO power is higher than -5 dBm, the shot noise is at least 10 dB higher than electronic background noise, which offers us 10 dB signal to noise ratio for shot noise measurement. The RMS has more than 40 dB difference in these two states, which indicates the homodyne detector is capable of reducing the classical noise in LO beam for more than 40 dB.

5.2.4 Experimental of key distribution

In the key distribution testing, the source is a 1550-nm laser with 12 dBm power from a tunable laser (Santec TSL-510). After a polarization controller, the laser is coupled to the photonic chip. AM and PM are then performed by adding two white noise signals from arbitrary waveform generator (HP 33120A). The white noise frequency can reach up to 10 MHz. For the current proof-of-principle test, the LO phase tuning is also conducted on the same chip. A Peltier device together with a temperature controller (Thorlab TED200C) is used to stabilize the temperature of the whole chip, which will reduce the noise from environmental heat fluctuation and heat crosstalk on the chip. The output from the homodyne detector is monitored in both time and frequency domain on an oscilloscope (Tectronix MDO4104B-3). Then the data are analyzed off-line using Matlab.

The noises and losses in the system are first carefully measured to calculate the secure key rate. Fig. 5-8 shows the different noise spectrums in the system. The

Figure 5-8: Noise power spectrum of the oscilloscope, electronic noise, shot noise and when the signal is modulated at 10-shot-noise-unit (SNU)

oscilloscope noise is at about -120 dBm level, and the electronic noise of the detector is about 30 dB higher. The shot noise is 5-10 dB higher than the electronic background noise, which corresponds to a 5-10 dB signal to noise ratio at the 1-10MHz band. The result also indicates the operation bandwidth for our off-chip homodyne detector is at the 1-10MHz band. White noise from function generator is used to modulate the signal randomly as required by the GG02 protocol. With modulation on, the measured output is 10 dB higher than the shot noise, which is referred as a 10 SNU modulation.

Then the distribution of Gaussian key is tested. The quadrature selection is achieved by tuning the phase of LO and maximizing the cross-modulation peak-to-peak difference as stated in Section 5.2.2. The output signal, the modulation signal for x and p quadrature modulation are all recorded. The data is collected on the oscilloscope for 4 ms with a sampling frequency of 25 MHz. The output signal at Bob side is synchronized with Alice's modulation signal by measuring the cross-correlation between the two signals. Fig. 5-9a (and Fig. 5-9b) shows the normalized cross-

Figure 5-9 Cross-correlation result of Bob's measurement result and Alice's modulation on (a) corresponding quadrature (b) different quadrature. The inset shows the correlated gaussian key in two different situations.

correlation measurement result between the x quadrature (and the p quadrature) modulation signal and the homodyne detector output while the homodyne detector is measuring the x quadrature. The peak value of the cross-correlation has about 10-times difference. All the signal is then synchronized based on the cross-correlation result and

then pass through a digital bandpass filter of 2.8 to 3.2 MHz. Then the filtered signals are down-sampled to 3 MHz to generate a set of correlated gaussian key. The gaussian keys are shown in the inset of Fig. 5-9a and Fig. 5-9b, with Bob's key as x -coordinate and Alice's key as y -coordinate. The result also proved the correlation between Bob's key and only one of Alice's key that has the same measured quadrature.

5.2.5 Secure key rate analysis

The CV-QKD system can be characterized mainly by the modulation variance $V_{mod} = V - I$, channel transmission T and the excess noise ε [185, 186]. By considering the source of noise, the total noise can be categorized into two parts, which can be expressed as $\chi_{tot} = \chi_{ch} + \chi_{det}/T$. The channel added noise $\chi_{ch} = 1/T - 1 + \varepsilon$ consists of the channel loss and excess noise, which is all the noise added before Bob. The homodyne detector added noise is expressed as $\chi_{det} = (1 + \nu_{el})/\eta - 1$, where ν_{el} is the electronic noise of Bob's equipment and η is homodyne detection efficiency. The noise source is clarified so that another assumption can be made, which is the so-called "Loose Security Assumption" or "Trusted-device Scenario," which means an eavesdropper cannot have access to the noise χ_{det} from the instrument in Bob's lab.

The mutual information between Alice and Bob I_{AB} is given by

$$I_{AB} = \frac{1}{2} \log_2 \frac{V_B}{V_{B|A}} = \frac{1}{2} \log_2 \frac{V + \chi_{tot}}{1 + \chi_{tot}}, \quad (5-8)$$

where $V_B = \eta T(V + \chi_{tot})$ is the variance obtained by Bob and $V_{B|A} = \eta T(1 + \chi_{tot})$ is the conditional variance of Bob's measurement regarding Alice's sending. Then Eve's information in the reverse reconciliation protocol is given by

$$I_{BE} = \frac{1}{2} \log_2 \frac{V_B}{V_{B|E}} = \frac{1}{2} \log_2 \frac{\eta T(V + \chi_{tot})}{\eta \left(\frac{1}{T(1/V + \chi_{ch})} + \chi_{det} \right)}. \quad (5-9)$$

The Holevo information between Eve and Bob is given by $\chi_{BE} = S_E - S_{E|B}$, where S_E is the von Neumann entropy for Eve's measurement and can be calculated from the symplectic eigenvalue $\lambda_{1,2}$ of the covariance matrix

$$\gamma_{AB} = \begin{pmatrix} V \cdot I_2 & \sqrt{T(V^2 - 1)} \cdot \sigma_z \\ \sqrt{T(V^2 - 1)} \cdot \sigma_z & T(V + \chi_{ch}) \cdot I_2 \end{pmatrix} = \begin{pmatrix} a \cdot I_2 & c \cdot \sigma_z \\ c \cdot \sigma_z & b \cdot I_2 \end{pmatrix} \quad (5-10)$$

where I_2 is the two-dimensional identity matrix and σ_z is the Pauli-z matrix. The symplectic eigenvalue can be calculated by

$$\lambda_{1,2} = \frac{1}{2} \left(\sqrt{(a+b)^2 - 4c^2} \pm (b-a) \right). \quad (5-11)$$

The entropy $S_{E|B}$ is calculated from the same covariance matrix after a projection caused by Bob's homodyne detection. The new symplectic eigenvalue $\lambda_{3,4}$ is given by

$$\lambda_{3,4} = \sqrt{\frac{1}{2} \left(\alpha \pm \sqrt{\alpha^2 - 4\beta} \right)}, \quad (5-12)$$

with

$$\begin{cases} \alpha = \frac{a(ab - c^2) + b + (a^2 + b^2 - 2c^2)\chi_{det}}{b + \chi_{det}}, \end{cases} \quad (5-13a)$$

$$\begin{cases} \beta = \frac{a(ab - c^2) + (ab - c^2)^2 \chi_{det}}{b + \chi_{det}}. \end{cases} \quad (5-47b)$$

The Holevo information is given by

$$\chi_{BE} = g\left(\frac{\lambda_1 - 1}{2}\right) + g\left(\frac{\lambda_2 - 1}{2}\right) - g\left(\frac{\lambda_3 - 1}{2}\right) - g\left(\frac{\lambda_4 - 1}{2}\right), \quad (5-14)$$

where $g(x) = (x + 1) \log_2(x + 1) - x \log_2 x$.

The secure key rate is calculated under the assumption of individual attack and collective attack [185, 186]. The coherent attack is the most potent attack, but it was

proven not more efficient than collective attacks for Gaussian modulated CV-QKD protocol [187, 188]. Therefore, the secure key rate calculation based on collective attack assumption could provide unconditional security. The asymptotic secure key fraction (bits/symbol) against individual attack and collective attack, which are referred to as Shannon key rate and Holevo key rate, is $r_{individual}^{asymptotic} \geq I_{AB} - I_{BE}$ and $r_{collective}^{asymptotic} \geq I_{AB} - \chi_{BE}$, respectively. I_{AB} and I_{BE} are the mutual information shared between Alice & Bob and between Bob & Eve respectively. χ_{BE} is the Holevo information between Eve & Bob. The secure key rate K is given by $K = SR \cdot r$, where SR is the symbol rate in the unit of symbols/s.

Above analysis has not considered the effect of all the post-processing. A practical data reconciliation method requires more bits to be exchanged than the theoretical limit. Also, some bits are used for system calibration and parameter estimation, which also cannot be used in final key extraction. Thus, a more practical secure key rate is given by $K_{individual,practical}^{asymptotic} = SR(1 - FER)(1 - \nu)(\beta I_{AB} - I_{BE})$ and $K_{collective,practical}^{asymptotic} = SR(1 - FER)(1 - \nu)(\beta I_{AB} - \chi_{BE})$, where SR is the symbol rate, FER is the frame-error-rate, ν is the fraction of bits that cannot be used to distill the final key and β is the reconciliation efficiency.

As shown in Fig. 5-10, the effect of modulation variance V_{mod} , channel transmittance T and the excess noise ε to the final secret fraction is calculated. The calculation is under the assumption of collective attack and a reconciliation efficiency of 0.95 is considered. The secret fraction in the unit of bits/symbol indicates how much secret keys can be extracted from one Gaussian symbol. In Fig. 5-10a, the secret fraction as a function of excess noise from 0.07 to 0.25 and transmittance from 0 to 1

Figure 5-10: Secure key fraction as a function of transmittance and (a) excess noise (b) modulation variance. The inset is a zoom-in view of the 0 to 0.2 region.

is shown. The modulation variance is $V_{mod} = 10$ in this calculation. When the secret fraction is less than 0, it means no secure key can be extracted from the key distribution process. The yellow shade marked such region. The result shows that lower the excess noise, higher the secure key fraction. When excess noise $\varepsilon = 0.25$, the minimum transmittance acceptable is about 0.8, which reaches the limitation for our fiber-chip

coupler. Thus, the excess noise must be lower than 0.25 for our photonic chip to function normally. Fig. 5-10b shows the secret fraction at a modulation variance of 2 to 15 and transmittance from 0 to 1. The total excess noise is set to $\varepsilon = 0.1$ in this calculation. At the low-loss and short-distance scenario (Transmittance > 0.1), the secret fraction increases when the modulation variance increases. For the modulation variance at 4, the limitation for transmittance is 0.12. Higher modulation variance could significantly increase the secure key fraction at low loss scenario. However, at the high-loss and long-distance region (Transmittance < 0.1), there is an optimum modulation variance available, as shown in the inset of Fig. 5-10b. The maximum secure key fraction and minimum acceptable transmittance occurs at a modulation variance of about $V = 4$. The corresponding maximum transmission distance is about 100 km.

5.2.6 Experimental secure key rate characterization

To evaluate the secure key rate of the current chip-based CV-QKD system, all the losses and noises is carefully measured. The total losses consist of the losses on the transmission line and Bob's equipment, while the losses on the Alice side cannot affect the final security key. The losses on the transmission line is set as a variable T with $1 > T > 0$. The total loss of Bob's equipment has 0.7 dB contribution on the signal path. The 0.7 dB loss is contributed by the 50:50 fiber coupler, the polarization controller and some fiber connectors. The homodyne detection efficiency is measured to be $\eta = 0.61$. In addition, the total excess noise $\varepsilon = 0.094$ shot noise unit (SNU) at a modulation variance of $V_{mod} = 8.2$ SNU and $T = 1$. Electrical noise $v_{el} = 0.064$ SNU. The parameters are also summarized in Table 5-3. With all the data available, the secure key rate of the current CV-QKD system, including the Alice chip and the off-chip

Table 5-3 Summary of parameters used in the secure key rate calculation.

Parameters	Value	Experimental/ Theoretical
Detector quantum efficiency	0.72	Experimental
Homodyne detection efficiency	0.61	Experimental
Electrical noise (SNU)	0.064	Experimental
Total excessive noise (SNU)	0.094	Experimental
Modulation variance (SNU)	8.23	Experimental
Operation band (MHz)	2.5	Experimental
Loss at receiver (dB)	0.7	Experimental
Fiber loss (dB/km)	0.14	Theoretical
Reconciliation efficiency	0.95 & 0.98	Theoretical

detector, can be estimated. As shown in Fig. 5-11, the Shannon raw key rate and Holevo raw key rate are given as the dashed line. Considering a more practical situation, the reconciliation efficiency $\beta = 0.95$ and 0.98 is chosen [42, 185, 189], which represents the commonly used case and the state-of-the-art case. The corresponding Shannon

Figure 5-11: The secure key rate under individual attack and collective attack.

The reconciliation efficiency of 0.95 and 0.98 is also considered.

effective key rate and Holevo effective key rate are shown as the solid line in Fig. 5-11. Based on the collective attack assumption, a transmission distance of about 120 km (65km) fiber length is achieved at reconciliation efficiency $\beta = 0.98$ ($\beta = 0.95$). Higher reconciliation efficiency could significantly increase the maximum transmission distance. Thus, with the 3 MHz symbol rate, the effective Holevo secure key rate at $\beta = 0.98$ can reach 0.83 MHz at 0 km and 2.3 kHz at 100 km of simulated fiber length.

5.3 Summary

To summarize, we present an integrated chip of GG02 CVQKD for the first time. All components, including beam splitters, modulators, multiplexers and homodyne detectors are integrated onto both the transmitter and receiver chips. The performance of each component is carefully characterized. In this proof-of-principle experiment, the key distribution occurred between the transmitter chip and an off-chip receiver. The experimental results show a secure key rate of 0.83 MHz at 0 km and 2.3 kHz at 100 km transmission distance under the collective attack assumption, which is comparable to the traditional bulk optics or fiber system. Although the experiment is performed with an off-chip detector, the experimental result proved that the chip-based detector provides similar performance, as shown in Table 5-4. The on-chip detector is also fully capable of performing a QKD experiment. However, the main limitation is the required LO power. In our future work, a shot noise limited laser source with higher power will be used for the chip characterization. The future demonstration will focus on implement a complete system including the chip-based laser source using III-V laser die bonding technique. The real-time control, synchronization and data processing could achieve by using a field-programmable gate array (FPGA) control card. Previous work of chip-based DV-QKD and optical transceiver using Si platform proved the

Table 5-4 Specification comparison of on-chip and off-chip homodyne detector used in this work.

	On-Chip Detector	Off-Chip Detector
CMRR (dB)	25	40
Quantum efficiency	0.59	0.72
Electrical noise (SNU)	0.0693	0.0635
Bandwidth (MHz)	10	10
Coupling loss (dB)	5	0.7
LO power (dBm)	11	-3

feasibility for our chip to implement more CV- or DV-QKD protocols and even integrate with the classical communication apparatus. Fully packaging of such a system may move one more step towards a miniaturized and powerful device. Together with the advantage of robustness and low cost, this work can push forward the practical usage of a hybrid quantum-classical communication chip for the advanced communication network.

Chapter 6 Conclusions

6.1 Conclusions

We present an integrated chip of GG02 CVQKD for the first time. All components, including beam splitters, modulators, multiplexers and homodyne detectors are integrated onto both the transmitter and receiver chips. Theoretical modeling, component design and simulation, chip fabrication and experimental studies were presented. The performance of each component is carefully characterized. In this proof-of-principle experiment, the key distribution occurred between the transmitter chip and an off-chip receiver. The major conclusions are listed below.

- 1) The fiber-chip coupler, including the taper coupler and grating coupler were designed and experimentally tested. For the taper coupler, the simulation showed an 87% coupling efficiency with a 160 nm waveguide and a 47% coupling efficiency with a 200 nm waveguide. For the grating coupler, the coupling efficiency was 38%. A special two-dimensional grating coupler was designed as a polarization beam splitter/combiner. The extinction ratio was found to be 35 dB.
- 2) The directional coupler and MMI devices were designed for a 50 : 50 beam splitter and a 1 : 99 beam splitter, which were then used to construct the MZI and ring resonator structures. The fabricated 1 : 99 MMI was measured to have an extinction ratio of 17 dB. The MZI based on 50 : 50 directional coupler had a maximum extinction ratio of 23 dB. The ring resonator had maximum Q-factor of 6×10^4 and an extinction ratio of 18 dB.

- 3) Two important polarization manipulation components, including the polarization beam splitter and polarization rotator were designed and tested. The two-stage cascade polarization beam splitter had a maximum 44 dB extinction ratio. The polarization rotator was designed for rotating between TE and TM mode. The extinction ratio was measured to be 25 dB maximum.
- 4) Three types of modulators were designed and studied in detail, including the thermo-optic modulator, carrier injection or depletion modulator and electro-optic modulator. The result showed that at the same applied voltage, the phase shift for the thermo-optic modulator was the highest. With careful design, for a 100- μm length modulator, a phase shift of 2π could be achieved under 2V applied voltage, which was impossible for the other two types. Thus, if compact size is the main requirement, the thermo-optic modulator is the modulator of choice. Considering the modulation frequency, both the carrier injection type and the electro-optic type could achieve above the GHz level, which was much higher than the thermo-optic modulator with kHz-level speed. Moreover, the electro-optic modulator based on AlN had some additional benefits. Not only did it have high second-order nonlinearity, but it could also achieve ultra-low power modulation with some additional fabrication cost. Thus, a guideline was provided for choosing the best type of modulators for a given application.
- 5) The on-chip germanium photodiode was designed and tested. The test results showed the responsivity for the Ge PD to be about 0.73 A/W, with an operation bandwidth of more than 2.4 GHz and dark current down to the nanoampere level. The homodyne detector based on this Ge PD showed a common mode rejection ratio of more than 25 dB and a working bandwidth of 1-10 MHz.

- 6) The CV-QKD Gaussian-modulated coherent state protocol, including the data reconciliation, error correction, and privacy amplification protocols were studied. The first chip for CV-QKD was designed, fabricated and tested. All the noise sources and losses in the system were measured and analyzed in detail. The results were used for the secure key rate analysis of the system. A proof-of-principle chip-based CV-QKD test was conducted. The experimental results showed a secure key rate of 0.83 MHz at 0 km and 2.3 kHz at 100 km transmission distance under the collective attack assumption.

The successful demonstration of such a single chip-based QKD offers new possibilities for low-cost, scalable and portable quantum networks. All these advantages lead to a scalable and more secure network, which offers high potential in the hybrid quantum-classical communication network using fiber and free-space optics.

6.2 Recommendations

Recommendations for future demonstration are summarized as follows:

- 1) Further system-level optimization of all the components used in the QKD system is possible. A delay line could be introduced to do time-division multiplexing to further reduce the cross-talk between the LO light and signal light. Some recent developments of the self-referenced CV-QKD protocol also suggest that the LO could be generated locally on Bob's side, which would significantly improve the security of the CV-QKD system.
- 2) A complete system could be implemented including the chip-based laser source using the III-V laser die bonding technique. The real-time control, synchronization, data reconciliation, error correction, privacy amplification and

data processing could be achieved by using a field-programmable gate array (FPGA) control card.

- 3) Previous work on chip-based DV-QKD and optical transceivers using the Si platform proved the feasibility for our chip to implement more CV- or DV-QKD protocols and even integrate with the classical communication apparatus. Fully encapsulating and packaging such a system may move one more step towards a miniaturized and powerful device. Together with the advantages of robustness and low cost, this work can push forward the practical use of a hybrid quantum-classical communication chip for advanced communication networking.

Author's Publications

Journal paper

- [1] Feng, X.*, **Zhang, G.***, Chin, L.K., Liu, A.Q. and Liedberg, B., 2017. Highly sensitive, label-free detection of 2, 4-dichlorophenoxyacetic acid using an optofluidic chip. *ACS sensors*, 2(7), pp.955-960. (*contribution equally)
- [2] **Zhang, G.**, Haw, J.Y., Cai, H., Xu, F., Assad, S.M., Fitzsimons, J.F., Zhou, X.Q., Yu, S., Wu, J., Ser, W., Kwek, L.C., and Liu, A.Q. A Single Integrated Chip for Continuous-variable Quantum Key Distribution. *Nature Photonics*. (Submitted)
- [3] **Zhang, G.**, Haw, J.Y., Cai, H., Xu, F., Assad, S.M., Fitzsimons, J.F., Zhou, X.Q., Yu, S., Wu, J., Ser, W., Kwek, L.C., and Liu, A.Q. High speed modulator using low loss aluminium nitride. *Applied physics letters*. (Preparing)

Conference paper

- [1] **Zhang, G.**, Cao, L., Haw, J.Y., Jia, X.J., Su, X.L., Kwek, L.C., Fitzsimons, J.F., Lam, P.K., Zhou, X.Q., Gao, W.B., Gong, J.B. and Liu, A.Q., 2018, May. Integrated Chip for Continuous-variable Quantum Key Distribution using Silicon Photonic Fabrication. In *CLEO: QELS_Fundamental Science* (pp. FTu3G-2). Optical Society of America. (Oral)
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Appendix A1 Optical Waveguide

The waveguide is the structure that guides waves, such as electromagnetic wave and sound. The simplest optical waveguide structure contains a core with a refractive index of n_1 and thickness $2a$, as shown in Fig. 1-1a. The cladding material surrounding the core with a refractive index of n_0 . While n_1 is higher than n_0 , light can be trapped and transmitted in the core with a minimum loss by total internal reflection. There are two major restraints for the light wave guided inside. One is the total internal reflection condition. The incident condition for the waveguide can be derived as [1, 2],

$$\theta_{in} \leq \sin^{-1} \sqrt{n_1^2 - n_0^2} \equiv \theta_{\max} \equiv NA, \quad (1-1)$$

where θ_{in} is the incident angle of the light beam, $\theta_{\max}$ is the maximum incident angle, which is also well known as the numerical aperture (NA).

Besides the total internal reflection limitation for incident angle θ_{in} , the propagation angle ϕ must satisfy the mode formation condition also. As shown in Fig. 1-1b, consider a plain wave traveling inside the waveguide core with a wavenumber of kn_1 . The light wave is traveling along the solid line and the dashed line marked the phase front of the light wave. Two different light ray AD and BC are at the same phase front, so the phase difference between the two light rays must be $2m\pi$, where m is an integer. To calculate the mode formation condition, the phase shift Φ at the total internal reflection interface should be considered also, which is given by [1]

$$\Phi = -2 \tan^{-1} \frac{\sqrt{n_1^2 \cos^2 \phi - n_0^2}}{n_1 \sin \phi} = -2 \tan^{-1} \sqrt{\frac{2\Delta}{\sin^2 \phi} - 1}, \quad (1-2)$$

Figure 1-1: (a) Total internal reflection condition for waveguide. (b) Mode formation condition for waveguide.

where the relative refractive index difference Δ between n_0 and n_1 is defined as

$$\Delta = \frac{n_1^2 - n_0^2}{2n_1^2} . \quad (1-3)$$

The phase shift at the total internal reflection interface is also known as the Goos-Hänchen shift. Moreover, the optical path length l_1 between B and C and l_2 between A and D is calculated as

$$\begin{cases} l_1 = 2a \left(\frac{1}{\sin \phi} - 2 \sin \phi \right) , & (1-4a) \\ l_2 = \frac{2a}{\sin \phi} . & (1-4b) \end{cases}$$

Considering the Goos-Hänchen shift at point A and D , the mode formation condition is described as

$$(kn_1l_2 + 2\Phi) - kn_1l_1 = 2m\pi, \quad (1-5)$$

where $k = 2\pi/\lambda$ is the wavenumber of light, and m is an integer. By substituting Eq. (1-3) to (1-4) into Eq. (1-5), the mode formation condition can be expressed as

$$\tan\left(kn_1a \sin \phi - \frac{m\pi}{2}\right) = \sqrt{\frac{2\Delta}{\sin^2 \phi} - 1}, \quad (1-6)$$

where propagation angle ϕ is the only variable when the waveguide structure is fixed. Since m is an integer, solving the equation can only produce a series of discrete ϕ value. The discrete solutions of light field satisfying Eq. (1-7) are called mode. When $m = 0$, the propagation angle ϕ reaches the minimum, which is called the fundamental mode. When $m > 0$, the propagation angles ϕ are larger. Those are called higher order mode. As shown in Fig. 1-2a and Fig. 1-2b, the positive phase front is shown in solid line and the negative phase front is shown in dashed line. Due to the interference, the light will form a standing wave in both the x - and z -axis. In x -axis, the fundamental mode has one constructive interference point in the center of the waveguide and two destructive interference points at the edge. Thus, the standing wave for fundamental mode forms a Gaussian-like shape in the x -axis. The first order mode has two constructive interference points and three destructive interference point in the x -axis, which forms a sine-like shape standing wave in the x -axis. In z -axis, the standing wave has a period of $\lambda/(n_1 \cos \phi)$, where λ is the wavelength of the light wave. Thus, the wave can be treated as transmitting inside a bulk material with refractive index of $n_1 \cos \phi$, which is defined as the effective refractive index n_{eff} of waveguide. By introducing the propagation constant along x and z direction as

Figure 1-2: Schematic for (a) fundamental mode and (b) first order mode formation.

$$\begin{cases} \beta = kn_1 \cos \phi, \\ \kappa = kn_1 \sin \phi. \end{cases} \quad (1-7a)$$

$$(1-7b)$$

Then, the effective refractive index becomes

$$n_{eff} = \frac{\beta}{k}, \text{ with } n_0 \leq n_{eff} \leq n_1. \quad (1-8)$$

For a rectangle shape waveguide structure, transverse electric (TE) wave and transverse magnetic (TM) wave will experience different waveguide thickness so that the mode will be different. The subscript number represents the order of mode. As an example, TE₀ and TM₀ are the fundamental TE and TM mode, TE₁ and TM₁ is the first order TE and TM mode. In this research, higher order mode will cause trouble for mode coupling with single mode fiber. Undesired mode coupling will also induce difficulties

Figure 1-3: The effective refractive index as a function of waveguide width for the fundamental, first order and second order modes.

in the calculation. As a result, single mode waveguide with only fundamental TE and TM mode is designed. Considering the available experimental setup and fabrication equipment, 1550 nm is chosen as the target wavelength. Silicon is used as the core material and silicon dioxide is used as the cladding material. The thickness of the waveguide is determined by the silicon on insulator (SOI) wafer used. The commonly used one has a silicon layer of 220 nm. The width is the parameter which needs special design. The refractive index for silicon and silicon dioxide at 1550 nm is 3.48 and 1.44, respectively. Lumerical MODE Solutions software is used to calculate the waveguide propagation mode. The relationship between effective refractive index and waveguide width is shown in Fig. 1-3. When the width is larger, the effective refractive index is increasing for all the modes supported. Typically, the higher the effective refractive index, better the mode confinement inside the waveguide, which also means a lower transmission loss. For an extreme case when $n_{eff} = n_{SiO_2}$, there is no power transmitted inside the waveguide. As a result, the waveguide can support a certain mode when the

effective refractive index $n_{eff} = n_{SiO_2}$. To restrict a single mode transmission and obtain better mode confinement in the same time, the waveguide width is chosen as 450 nm, where TE₁ mode just appears.

Appendix A2 Quantum Optics Introduction

Quantum optics is the field that using quantum mechanics theory to investigate the phenomena involving light and light-matter interaction. In this section, some basic information related to continuous variable QKD are provided, including the definition of field quadrature and the measurement uncertainty of field quadrature.

A2.1 Light waves as classical harmonic oscillator

We begin by explaining the connection between light and classical harmonic oscillator. For a classical harmonic oscillator with mass m and angular frequency ω , the displacement x and momentum p_x are related by [3, 4]

$$\begin{cases} p_x = m\dot{x} , \\ m\ddot{x} = \dot{p}_x = -m\omega^2 x . \end{cases} \quad \begin{matrix} (2-1a) \\ (2-1b) \end{matrix}$$

The solution for x and p are

$$\begin{cases} x(t) = x_0 \sin \omega t , \\ p(t) = p_0 \cos \omega t , \end{cases} \quad \begin{matrix} (2-2a) \\ (2-2b) \end{matrix}$$

with

$$p_0 = m\omega x_0 . \quad (2-3)$$

The energy of the simple harmonic oscillator (SHO) is

$$E_{SHO} = \frac{p_x^2}{2m} + \frac{1}{2}m\omega^2 x^2 , \quad (2-4)$$

where the first term represents the kinetic energy and the second term represents the potential energy of the particle.

To represent light wave as a harmonic oscillator, the equivalent position and momentum for a light wave must satisfy Eq. (2-1) and (2-4). Consider a monochromatic light transmitting in the z -axis and linearly polarize in the x -axis. The electric field $\mathcal{E}_x$ can be expressed as

$$\mathcal{E}_x(z, t) = \mathcal{E}_0 \sin kz \sin \omega t , \quad (2-5)$$

where $\mathcal{E}_0$ is the amplitude, $k = 2\pi/\lambda$ is the wavenumber of light and ω is the angular frequency. Since the electric field is in the x -axis, the magnetic field will be in the y -axis. Based on the Maxwell equation, the magnetic field B_y and the electric field is related by

$$-\frac{\partial B_y}{\partial z} = \epsilon_0 \mu_0 \frac{\partial \mathcal{E}_x}{\partial t} , \quad (2-6)$$

where ϵ_0 is the vacuum permittivity and μ_0 is the vacuum permeability. The resulting magnetic field can be expressed as

$$B_y(z, t) = B_0 \cos kz \cos \omega t , \quad (2-7)$$

where

$$B_0 = \frac{\mathcal{E}_0}{c} . \quad (2-8)$$

Then the energy of the wave can be calculated by integrating the energy density U , which is given by

$$U = \frac{1}{2} \left(\epsilon_0 \mathcal{E}^2 + \frac{1}{\mu_0} B^2 \right) . \quad (2-9)$$

The total energy E is

$$E = \frac{V}{4} \left(\epsilon_0 \mathcal{E}_0^2 \sin^2 \omega t + \frac{B_0^2}{\mu_0} \cos^2 \omega t \right) . \quad (2-10)$$

Then two new parameters q and p are introduced as

$$\begin{cases} q(t) = \left(\frac{\epsilon_0 V}{2\omega^2} \right)^{1/2} \mathcal{E}_0 \sin \omega t, \end{cases} \quad (2-11a)$$

$$\begin{cases} p(t) = \left(\frac{V}{2\mu_0} \right)^{1/2} B_0 \cos \omega t \equiv \left(\frac{\epsilon_0 V}{2} \right)^{1/2} \mathcal{E}_0 \cos \omega t. \end{cases} \quad (2-11b)$$

It is clear to see that p and q satisfying the following relationship,

$$\begin{cases} p = \dot{q}, \end{cases} \quad (2-12a)$$

$$\begin{cases} \dot{p} = -\omega^2 q. \end{cases} \quad (2-12b)$$

Eq. (2-12) has exactly the same form with Eq. (2-1) by assuming

$$\begin{cases} q(t) = \sqrt{m}x(t), \end{cases} \quad (2-13a)$$

$$\begin{cases} p(t) = \left(\frac{1}{\sqrt{m}} \right) p_x(t). \end{cases} \quad (2-13b)$$

Thus, $p(t)$ and $q(t)$ are called the generalized position and momentum coordinates. On substituting Eq. (2-11) into Eq. (2-10), the total energy can be expressed as

$$E = \frac{1}{2} (p^2 + \omega^2 q^2), \quad (2-14)$$

which is also in the same form as the energy of a classical harmonic oscillator. For a classical harmonic oscillator, the energy oscillates back and forth between the kinetic and potential energy. For an electromagnetic wave, the energy oscillates back and forth between the electric field and magnetic field. The results give an analogy between the classical harmonic oscillator and the electromagnetic wave. Thus, the electromagnetic wave can be analyzed at quantum mechanics level.

A2.2 Field Quadrature

Then the definition of field quadrature is provided. Consider again for a linear polarized monochromatic light with phase ϕ , the electric field can be written as [3]

$$\begin{aligned}\mathcal{E}_x(z, t) &= \mathcal{E}_0 \sin kz \sin(\omega t + \phi) \\ &= \mathcal{E}_0 \sin kz (\cos \phi \sin \omega t + \sin \phi \cos \omega t) \\ &= \mathcal{E}_1 \sin \omega t + \mathcal{E}_2 \cos \omega t ,\end{aligned}\tag{2-15}$$

where $\mathcal{E}_1 = \mathcal{E}_0 \sin kz \cos \phi$ and $\mathcal{E}_2 = \mathcal{E}_0 \sin kz \sin \phi$ are called the field quadrature.

The electric field amplitude at point z can be expressed in a complex form with

$$\mathcal{E}(z) = \mathcal{E}_0(z) e^{i\phi} = \mathcal{E}_1(z) + i\mathcal{E}_2(z) ,\tag{2-16}$$

where $\mathcal{E}_0(z) = \mathcal{E}_0 \sin kz$. In this complex form, the field amplitude can be represented as a vector in the complex plane, with a vector length of $\mathcal{E}_0(z)$ and angle of ϕ . The real part and the imaginary part corresponded to the projection on x - and y -axis, respectively. To simplify future calculations, the field quadratures are then replaced by the dimensionless field quadratures $X_1(t)$ and $X_2(t)$, which can be expressed as

$$\left\{ \begin{aligned} X_1(t) &= \left(\frac{\epsilon_0 V}{4\hbar\omega} \right)^{1/2} \mathcal{E}_0 \sin \omega t , \end{aligned} \right.\tag{2-17a}$$

$$\left\{ \begin{aligned} X_2(t) &= \left(\frac{\epsilon_0 V}{4\hbar\omega} \right)^{1/2} \mathcal{E}_0 \cos \omega t . \end{aligned} \right.\tag{2-17b}$$

Then the electric field in time and space can represent with the dimensionless field quadratures as

$$\mathcal{E}_x(z, t) = \left(\frac{4\hbar\omega}{\epsilon_0 V} \right)^{1/2} \sin kz (\cos \phi X_1(t) + \sin \phi X_2(t)) .\tag{2-18}$$

Comparing Eq. (2-11) with (2-17), the field quadrature of light is directly connected to the position and momentum of the simple harmonic oscillator with

$$\begin{cases} X_1(t) = \left(\frac{\omega}{2\hbar}\right)^{1/2} q(t) , \\ X_2(t) = \left(\frac{1}{2\hbar\omega}\right)^{1/2} p(t) . \end{cases} \quad (2-19a)$$

$$(2-19b)$$

The connection can then be used to analyze electromagnetic wave with the knowledge from quantum mechanics.

To measure the field quadrature, a homodyne detector is used. As shown in Section 4.2.2, homodyne detector subtracts the signal from two balanced PDs and gives an output current $i = i_1 - i_2$. Consider the input signal as E_S and LO as $E_{LO} \exp(i\phi_{LO})$, the output current can be expressed as

$$i \propto 2E_{LO}(E_S^X \cos \phi_{LO} + E_S^P \sin \phi_{LO}) , \quad (2-20)$$

where E_{LO} is the amplitude of LO beam and ϕ_{LO} is its phase relative to the signal beam. E_S^X and E_S^P is the amplitude of the signal light projecting to x and p quadrature, respectively. When $\phi_{LO} = 0$, the output is only proportional to E_S^X , and when $\phi_{LO} = \pi/2$, the output is proportional to E_S^P . Thus, the x or p quadrature of the signal light can be measured by tuning the phase of LO light.

A2.3 Light waves as quantum harmonic oscillator

As mentioned in the last chapter, the knowledge on quantum harmonic oscillator can be used to analyze the quantized electromagnetic wave. There are two important relationships involved. One is the quantized energy level in the form of [3]

$$E_n = \left(n + \frac{1}{2}\right) \hbar\omega , \quad (2-21)$$

where n is a non-negative integer. The equation shows that for a quantum harmonic oscillator, the total energies available are a series of discrete values. The minimum

possible energy is $\frac{1}{2}\hbar\omega$. For an electromagnetic wave, it can be interpreted as the total energy for n -photons $n\hbar\omega$ plus the vacuum energy $\frac{1}{2}\hbar\omega$. The other relationship is the Heisenberg uncertainty principle for x and p , which can be expressed as

$$\Delta x \Delta p_x \geq \frac{\hbar}{2} . \quad (2-22)$$

Consider the relationship shown in Eq. (2-13) and (2-19), the total uncertainty for quadratures X_1 and X_2 can be given by

$$\Delta X_1 \Delta X_2 = \left(\frac{\omega}{2\hbar}\right)^{\frac{1}{2}} \left(\frac{\Delta x}{\sqrt{m}}\right) \left(\frac{1}{2\hbar\omega}\right)^{\frac{1}{2}} \left(\sqrt{m} \Delta p_x\right) = \frac{1}{2\hbar} \Delta x \Delta p_x . \quad (2-23)$$

By combining Eq. (2-22) and (2-23), the uncertainty for quadratures X_1 and X_2 becomes

$$\Delta X_1 \Delta X_2 \geq \frac{1}{4} . \quad (2-24)$$

For a classical light, the two quadratures are random. The minimum uncertainty for X_1 and X_2 appears when

$$\Delta X_1 = \Delta X_2 = \frac{1}{2} . \quad (2-25)$$

This is the case for a linear polarized monochromatic light with totally no classical noise in it. In quantum optics, this is named as the coherent state $|\alpha\rangle$, where α is a complex number defined as

$$\alpha = X_1 + iX_2 . \quad (2-26)$$

Due to the measurement uncertainty of both quadratures, there is a physical limitation for classical noise reduction techniques. The minimum achievable noise level by classical noise reduction techniques is called the shot noise, which can be treated as a boundary between classical and quantum light. A perfect coherent light has exactly shot noise level noise. The light with sub-shot-noise level noise in any quadrature is

called a squeezed state, which has no classical equivalent. In our CV-QKD experiment, the protocol based on coherent state is used. The light source required is just a low noise laser without any special “quantum” treatment.

Appendix A3 Quantum Information Theory

In this section, firstly, some basic concept of information theory will be covered, including the definition of entropy, which will be used to analyze the security of QKD system.

A3.1 Shannon entropy

Shannon entropy is named after Claude Shannon, which measures the average amount of information stored in a series of data. Like the entropy in thermodynamics, Shannon entropy also measures the disorder of the data series. Higher Shannon entropy means the data series is more disordered. Thus, more information is required to describe the data series. In another word, more information can be encoded in the data series. For a random data series X with possible values x , the probability to get each possible values x is p_x . The Shannon entropy $H(X)$ is defined as [4]

$$H(X) \equiv - \sum_x p_x \log_2 p_x , \quad (3-1)$$

with the unit of digits. Here we also define $0 \log 0 = 0$. The maximum Shannon entropy is 1, which means the information represented by one digit is one digit, and there is no data redundancy. For a data series X contains only (0, 1), the maximum Shannon entropy of 1 bit occurs when both the probabilities for getting 0 and 1 are 1/2. When one of the probability p_x is 0, there's no randomness in the data series, so the Shannon entropy becomes 0.

For a continuous random variable X with probability density function $p(x)$, Shannon entropy can also be defined in a similar way as

$$H(X) = - \int_{-\infty}^{\infty} p(x) \log p(x) dx . \quad (3-2)$$

The joint entropy measures the disorder of a pair of random variables (X, Y) , which is defined as

$$H(X, Y) \equiv - \sum_{x,y} p(x, y) \log p(x, y) . \quad (3-3)$$

For the same random variable pair (X, Y) , if the measurement is only performed on Y and $H(Y)$ bits of information is acquired, the conditional entropy is defined as the rest of information on the pair (X, Y) , which is given by

$$H(X | Y) \equiv H(X, Y) - H(Y) . \quad (3-4)$$

The mutual information measures the mutual part of information between two random variables (X, Y) . It is defined as

$$H(X : Y) \equiv H(X, Y) - H(X | Y) . \quad (3-5)$$

Or

$$H(X : Y) \equiv H(X) + H(Y) - H(X, Y) . \quad (3-6)$$

A3.2 Von Neumann entropy

The Shannon entropy measures the information carried in the classical probability distribution. For a quantum state, the information can be calculated similarly, which is called the Von Neumann entropy. For a quantum state ρ , the Von Neumann entropy is given by [4]

$$S(\rho) = -\text{tr}(\rho \log_2 \rho) . \quad (3-7)$$

It can also be expressed with the eigenvalue of ρ . Consider λ_x as the eigenvalues of ρ , the Von Neumann entropy can be re-arranged as

$$S(\rho) = - \sum_x \lambda_x \log_2 \lambda_x . \quad (3-8)$$

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