Towards Integrated QPSK Transceiver on Zero-Change CMOS Foundry Process

by

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Submitted to the Department of Electrical Engineering and Computer Science
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Abstract

In recent years, the demand for Internet bandwidth increases while the unit price of bandwidth decreases. To keep up with the trend, more cost-effective optical telecommunication links are required. Due to the limited amount of optical fibers and wavelength range, increasing spectral efficiency is a desirable approach. Advance modulation scheme, such as quadrature phase-shift keying (QPSK), is a possible solution. Another limitation is energy consumption. Power density within telecommunication data centers are regulated by law in consideration of ambient temperature and noise level. Therefore, optical transceivers with higher energy efficiency is desired. Traditional transceivers use III-V chips for photonic components and CMOS chips for electronic circuits. Monolithic integration of photonic and electronic components helps removing energy consumption for inter-chip communication and hence increase overall energy efficiency.

In this thesis, a QPSK transceiver in zero-change CMOS process is proposed. Research is focused on three photonic components: 90 degree hybrids, poly-silicon photodetectors and QPSK modulators. Hybrids are used to mix QPSK-modulated signal and local oscillators with four equally-spaced phase delays. Multi-mode interferometers (MMI) are designed for this purpose. Best devices provides intensity imbalance around 1 dB and phase error around 10°. For poly-silicon photodetectors, sub-bandgap defect states are used for electron-hole pair generation. A ring-resonant structure is used to enhance absorption and reduce footprint. Best devices have quantum efficiency around 3.5% and dark current less than 60 nA under 25V reverse bias. The 3 dB frequency is around 1.2 GHz. Finally, double-coupled ring resonators with feed through waveguides are used for QPSK modulators. Carrier injection changes the resonance condition and provides phase shift. With the development of these devices, the path towards a monolithic QPSK transceiver in CMOS becomes clear.

Thesis Supervisor: Rajeev Ram
Title: Professor of Electrical Engineering
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Chapter 1

Introduction

The demand of telecommunication bandwidth has increased dramatically over the years. As shown in Fig. 1-1(a), Internet traffic in North America has been increasing exponentially over time. In the meanwhile, the unit price of Internet transit, defined as “the business relationship whereby one ISP provides access to all destinations in its routing table” and measured by $/Mbps, has been decreasing exponentially. These trends require the telecommunication community to provide solutions with exponentially better cost efficiency over time.

Due to the high price of laying new optical fibers, the industry has focused on increasing the bandwidth into individual fibers. In long haul telecommunication system where Erbium doped fiber amplifier (EDFA) is needed, the wavelength range for optical carriers are limited to 1530 to 1565 nm, also known as C-band. Under this circumstances, spectral efficiency, defined as bits/s/Hz, determines the capacity of an optical fiber. Researchers are building optical links with more colors, higher modulation rates and advanced modulation schemes. However, as the communication bandwidth increases, the energy consumption increases as well, since the basic building block, transistors’ power consumption is proportional to $CV^2f$, where $C$ is the capacitance, $V$ is the voltage across the transistor and $f$ stands for the frequency. However, power density in a central office or data center is limited. A typical IT equipment rack with a footprint of 0.622 square meter is allowed to consume 5 kW of power. This limit is primarily set by safe ambient temperature and noise level.
Figure 1-1: Increasing demand and decreasing price for Internet connection caused by cooling fans\cite{20}.

Widely deployed telecommunication system use wavelength division multiplexing on-off key modulation (WDM-OOK), which is illustrated in Fig. 1-2. As shown in Fig. 1-2(a) in time domain, the carrier is turned on and off to represent 1 and 0. This corresponds to two points in the constellation diagram shown in Fig. 1-2(b). OOK does not distinguish between in-phase and quadrature channels.

Figure 1-2: Illustration of OOK modulation scheme

For higher spectral efficiency, quadrature phase shift key (QPSK) is introduced
and implemented. For even higher spectral efficiency, higher order quadrature amplitude modulation (QAM) is in research as well. In this chapter, the background of QPSK modulation scheme will be illustrated in detail first. Afterwards, the incentive for photonic integration and presentation of state-of-the-art will be given, followed by the background knowledge of common photonic integration platforms. Among these platforms, advantages and concerns about CMOS photonic will be elaborated. Additionally, a literature review will be given regarding three important components of QPSK optical links: 90 degree hybrids, telecommunication wavelength photodetectors and QPSK modulators in CMOS compatible environment. Theoretical background will be given for these devices as well. Finally, an idea for an integrated QPSK modulator will be presented.

1.1 QPSK Modulation Scheme

To further increase bandwidth efficiency, advanced modulation schemes, such as DP-QPSK(Dual Polarization Quadrature Phase Shift Key), may be introduced. Theoretical discussion about QPSK started in the early 70s\cite{23} and demonstration of optical QPSK modulators can be found in 1990\cite{24}. Aside from expanding the communication data rate, QPSK enables electronic signal processing for distortion handling, such as dispersion compensation. In QPSK modulation, the signal is separately phase modulated into two orthogonal channels of the same electrical field. A simple constellation is shown in Fig. 1-3(a). It can be observed that as phase is modulated between two opposite points in one channel, the other channel will not be affected. Therefore, a single beam of light with a single wavelength is capable of carrying two distinct signal streams at the same time under QPSK modulation scheme.

At the receiver side shown in Fig. 1-3(b), the modulated signal is mixed with a coherent local oscillator by a 90 degree hybrid. Ideally, the output of the 90 degree hybrid are 4 vector summation of signal and local oscillator fields, where the 4 relative phase between them is 90 degree away from each other. The signal and local oscillator
can be represented by Eqn. (1.1)

\[ E_s(t) = \sqrt{P_s(t)}e^{j\phi_s(t)}e^{j\omega_s t} \]

\[ E_{LO}(t) = \sqrt{P_{LO}(t)}e^{j\phi_{LO}(t)}e^{j\omega_{LO} t} \quad (1.1) \]

Assume the signal and local oscillator are mixed with an ideal 90 degree hybrid and shine into 4 ideal photodetectors with identical responsivity. The two outputs from I channel are shown in Eqn. (1.2)

\[ I_1(t) = \frac{R}{2} \{ P_s(t) + P_{LO}(t) + 2\sqrt{P_s(t)P_{LO}(t)} \sin [(\omega_s - \omega_{LO})t + \phi_s(t) - \phi_{LO}(t)] \} \]

\[ I_2(t) = \frac{R}{2} \{ P_s(t) + P_{LO}(t) - 2\sqrt{P_s(t)P_{LO}(t)} \sin [(\omega_s - \omega_{LO})t + \phi_s(t) - \phi_{LO}(t)] \} \quad (1.2) \]

If the two outputs are connected in a balanced configuration, the difference of two photocurrents will be measured. The common DC portion \((P_s(t) + P_{LO}(t))\) is then
removed and the output from the I channel is shown in Eqn. 1.3

\[ \Delta I(t) = I_1(t) - I_2(t) = 2R\sqrt{P_s(t)P_{LO}(t)} \sin[(\omega_s - \omega_{LO})t + \phi_s(t) - \phi_{LO}(t)] \] (1.3)

Under ideal QPSK modulation, the power of signal and local oscillator should be time invariant. Additionally, the frequency of the local oscillator is locked onto the signal frequency, using a DSP and phase-tracking algorithm. Therefore, the output field reduces to Eqn. 1.4. It can be observed that by modulating the signal by a \( \pi \) phase shift, the output signal will change intensity and sign.

\[ \Delta I(t) = 2R\sqrt{P_sP_{LO}} \sin[\phi_s(t) - \phi_{LO}] \] (1.4)

By applying the same derivation to quadrature channel signal, the output current is shown in Eqn. 1.5. It can be observed that the output here is orthogonal to the I channel. Therefore, phase modulation of \( \pi \) will not cross talk with each other.

\[ \Delta I(t) = 2R\sqrt{P_sP_{LO}} \cos[\phi_s(t) - \phi_{LO}] \] (1.5)

In a balanced detection scheme shown in Fig. 1-3(c) ideally, the system should give a zero output when illuminated equally. However, in reality, a few defects can lead to a non-zero output, such as imbalance in power splitting of 90 degree hybrid and difference in two photodetectors. In order to characterize the performance, the idea of common mode rejection ratio (CMRR) is introduced[1] and defined in Eqn. 1.6. \(|I_1|\) and \(|I_2|\) are the output current under single illumination upon the two detectors respectively.

\[ CMRR = \frac{|\Delta I|}{|I_1| + |I_2|} \] (1.6)
## 1.2 Incentive for Photonic Integration

The requirement for future optical communication link points to high level of integration, which has advantages in energy efficiency, lower cost, performance and reliability.

First of all, integration helps reduce both electrical and optical loss between components and hence increases energy efficiency. For electronic circuits, communication between the chip and the outside world is through contact pads. The contact pad is typically much larger than the typical transistor gate length of tens of nm. The pad size for a high speed I/O in 130 nm CMOS has a width of around 50 $\mu$m. A large area corresponds to large capacitance. Since the energy stored in a capacitor is proportional to $CV^2$, it takes significant amount of energy for signal I/O between chips. A typical high speed electric link on the order of few Gbps cost 20 pJ/bit, which is unacceptable for future Terabit optical link. Literature shows for CMOS transimpedance amplifier (TIA), driving output pads cost approximately 90% of the total energy consumption. By integrating different electrical circuit functions, such as analog-digital converter (ADC), digital signal processor (DSP) and TIA, a significant portion of energy can be saved. Additionally, less energy consumption result in less power consumption in cooling as well.

On the optical side, integration reduces energy cost as well. Since an optical fiber has a circular mode with diameter of 9 $\mu$m and on chip waveguide is typically rectangular and much smaller, there is a large mismatch between fiber mode and on chip waveguide mode. Therefore, coupling between chip and fiber has always been a challenge. Typically, a loss of at least 1 dB is present in each coupling. By integrating multiple optic devices in the same chip, the coupling loss can be eliminated. Additionally, it is beneficial to integrate optics and electronics in the same chip, since the link between photodiode and TIA is susceptible to ambient electromagnetic noise, which deteriorates the signal noise ratio. In summary, it is energy efficient to integrate multiple electronics and optics devices to the same chip.

Secondly, integrated optics can be cost efficient. Fiber to chip packaging is chal-
lenge since the mode size and refractive index of fiber and on chip waveguide is very different. It takes careful design to minimize the coupling loss. Additionally, for a fiber array coupling into a photonic chip, alignment poses even more challenge. To connect electronics chips to PCB, costly wire bonding or flip-chip bonding is needed. The packaging cost can contribute up to 90% of a opto-electronic device\textsuperscript{[28]}. By integrating multiple devices together, the packaging cost can be greatly reduced. A linecard is a collection of electric and optic devices that interfaces with optic network. As shown in Fig. 1-4, in the ideal case, a linecard, which is assembled from individual laser chip, transceiver (combination of transmitter and receiver) and other electrical components, can be all integrated into CMOS process and fabricated from 12 inch Si wafer. This integration reduces the packaging cost as well as fabrication cost. Moreover, reduction of packaging, joints and coupling increases the system reliability as well\textsuperscript{[29]}.

![Value chain for CMOS integration](image)

Figure 1-4: Value chain for CMOS integration\textsuperscript{[2]}
1.3 State-of-the-art Photonic Integration Circuits for Telecommunication

The telecommunication industry is approaching the Terabit per chip era. The state-of-the-art photonic integration circuits (PIC) is mainly in traditional InP platform. Higher and higher order of integration is observed.

For WDM-OOK system, Infinera has demonstrated 40x40 Gbps capacity in a single InP PIC in 2006[6]. The block diagram is shown in Fig. 1-5. An electro-absorption modulator is employed. An array waveguide grating (AWG) is used to combine 40 colors.

![Figure 1-5: 40x40G WDM transmitter diagram](image)

For QPSK system, Infinera has developed single chip 10x40G, 5x100G and 10x100G links over the years in 2005, 2010 and 2011 respectively[6]. Block diagrams of 10x40G transmitter and receiver are shown in Fig. 1-6 and Fig. 1-8 respectively. Illustration of a single Mach-Zehnder modulator is shown in Fig. 1-7. The input laser is divided into two streams and modulated individually. The two streams will be polarization multiplexed into transverse electric (TE) and transverse (TM) mode later. The fields in the two polarizations are orthogonal to each other. Therefore, they are capable of carrying separate information. Each polarization is phase modulated for both I and Q channels. On the receiver side, an AWG is used for de-multiplexing. 2x4 hybrid are used for mixing with local oscillators.

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Figure 1-6: 10x40G QPSK transmitter

Figure 1-7: Mach-Zehnder QPSK modulator

Figure 1-8: 10x40G QPSK Receiver
Photos of fabricated PIC and packages for 5x100G QPSK transmitter and receiver are shown in Fig. 1-9(a) and Fig. 1-9(b) respectively.

(a) 500G Transmitter PIC
(b) 500G Receiver PIC

Figure 1-9: Photos of 500G transmitter and receiver PIC

Literature in 2011 demonstrates of 10x112 Gbps for both transmitter and receiver PIC from Infinera, which enables terabits QPSK optical link.

1.4 Integrated Photonic Platform

Multiple approaches have been investigated and developed towards photonic integration. They include pure InP platform, Silicon III-V hybrid platform, Germanium-on-silicon platform and CMOS photonics. The InP platform is a mature III-V process for photonics and it is the most advanced photonic platform for Infinera so far. Optic components, such as detectors, modulators and lasers with great performance are available in this platform. However, electronics on InP is very limited. Vitesse, the leading semiconductor foundry focused on IC on InP platform, has integrated transimpedance amplifiers and limiting post amplifiers in their product line. Most complicated circuits, such as DSP, which is essential to QPSK link, only exist on
CMOS platform. The three other approaches are CMOS compatible. Therefore, ultimate integration of optics and electronic on the same chip is possible. A comparison of these platforms are shown in Table 1.1

<table>
<thead>
<tr>
<th></th>
<th>Pure InP</th>
<th>III-V Hybrid</th>
<th>Ge-on-Si</th>
<th>Zero-change CMOS</th>
</tr>
</thead>
<tbody>
<tr>
<td>Fab challenge</td>
<td>Small wafer size</td>
<td>wafer bonding</td>
<td>high strain</td>
<td>construct dielectric environment</td>
</tr>
<tr>
<td>Fab cost</td>
<td>wafer cost</td>
<td>III-V wafer bonding</td>
<td>extra mask for Ge cost</td>
<td>minimal marginal cost</td>
</tr>
<tr>
<td>Scalability</td>
<td>limited by small wafer size</td>
<td>limited by small III-V wafer size</td>
<td>The same as CMOS</td>
<td>The same as CMOS</td>
</tr>
<tr>
<td>Absorption material</td>
<td>InP compatible III-V</td>
<td>III-V</td>
<td>Ge</td>
<td>SiGe or Poly-Si</td>
</tr>
<tr>
<td>Wavelength range</td>
<td>full C band</td>
<td>full C band</td>
<td>Up to 1550nm</td>
<td>Up to 1300nm for SiGe</td>
</tr>
<tr>
<td>PD responsivity</td>
<td>1.1A/W</td>
<td>1.1A/W</td>
<td>1.1A/W</td>
<td>0.31A/W</td>
</tr>
<tr>
<td>PD dark current</td>
<td>100pA</td>
<td>100mA at -2V</td>
<td>1µA</td>
<td>30mA</td>
</tr>
<tr>
<td>PD speed</td>
<td>86GHz</td>
<td>467MHz at -4V</td>
<td>32GHz</td>
<td>3GHz</td>
</tr>
</tbody>
</table>

Table 1.1: Integration Platforms for QPSK Receiver

1.4.1 Silicon III-V Hybrid Approach

Compared with Group IV semiconductors (Silicon, Germanium), III-V material systems have superior optical properties, since most III-V materials have direct bandgap, where both Silicon and Germanium have indirect bandgap. On the other hand, the Silicon based material system is dominant in CMOS process. Therefore, high precision lithography can be leveraged to fabricated sophisticated optical structure with minimum marginal cost. Additionally, high index contrast and low loss enable highly compact passive devices. Therefore, an obvious approach to leverage the advantage of both sides is to combine them together.

Due to large lattice mismatch between Silicon and III-V materials, direct growth is impossible. Therefore, a wafer bonding technique must be employed. The predominant process in literature is shown in Fig. 1-10(a). Wafer bonding relies on van der Waals bonds or hydrogen bonds. Essential steps includes growth of a thin layer (5nm) of highly strained native oxide is grown on both SOI and III-V wafers, surface activation by hydroxyl groups, vaporization of trapped liquid, spontaneous mating and thermal annealing. The gas by-product is detrimental to bonding strength. Therefore, vertical outgassing channels (VOC) are introduced. The Si III-V hybrid platform is useful for various optical devices. Electrically pumped lasers[39, 41, 42, 43], amplifiers[44, 45] and photodetectors[32, 36, 45] have been demonstrated. The cross-section of a typical hybrid laser is shown in Fig. 1-10(b). It can be observed that
Figure 1-10: Illustration of wafer bonding and hybrid laser device
the III-V material is bonded on top of the SOI substrate. The optical mode is jointly confined by the buried oxide and air gap around the Silicon waveguide. The top portion of the waveguide overlaps with the gain portion, which are multiple quantum wells (MQW). The confinement factor in MQW can be easily controlled by adjusting the width of the Silicon waveguide. Maximum power of around 14mW and differential efficiency of 12.7% is reported for this device.

1.4.2 Germanium on Silicon Approach

Germanium has been an attractive alternative approach to build active devices on Silicon substrate due to its lower bandgap, higher electron and hole mobility and other favorable properties. Rapid progress has been reported about Ge-on-Si photodetectors\textsuperscript{37}, modulators\textsuperscript{46} and electrically pumped laser\textsuperscript{47}.

The first challenge of the Ge-on-Si platform is the epitaxial growth, since Ge and Si has a lattice mismatch of 4.2%. This will result in high surface roughness and high density of threading dislocations. One possible approach to address this problem is to use SiGe buffer layer. The first report on this method appeared in 1984\textsuperscript{48}. There are other approaches without using SiGe buffer layer. For instance\textsuperscript{37}, a two step growth technique is reported. The first step is grow a thin Ge buffer layer of 30-60nm directly on Si substrate under low temperature (320-360°C). The islanding of Ge is suppressed under such low temperature. When the buffer layer is thicker than 30nm, the lattice mismatch does not affect the Ge any more.

The second major challenge is the fact that Ge is an indirect bandgap material under relaxed condition. As shown in Fig. 1-11(a), the L valley has lower energy (0.664eV) than Γ valley (0.8eV). This problem can be addressed by strain engineering. Under biaxial tensile strain, both the L and Γ valley drop in energy. However, the Γ valley drops faster. The relationship between bandgaps and tensile strain is shown in Fig. 1-11(b). It can be observed that at roughly 2% tensile strain, Ge becomes direct bandgap material. Usually, thin Ge epitaxial layers on Si is compressive strained, since Ge has a larger than lattice constant than Si. However, since the critical thickness is very small, a Ge layer thicker than 200nm is almost completely
relaxed above 600°C. Since Ge has a larger thermal expansion coefficient than Si, tensile strain forms in Ge when cooled down. A tensile strain of 0.25% can be obtained by this method. Additionally, there are reports about using GeSn buffer layer to reach a tensile strain of 0.68% [49].

Since the currently achieved tensile strength is not sufficient to convert Ge to direct bandgap material, heavy doping is introduced to fill the L valley. Literature [50] reports with a 0.25% tensile strain and n-type doping level of $7.6 \times 10^{19}$ cm$^{-3}$, direct bandgap transition is observed and it is possible to achieve a gain of 400 cm$^{-1}$. The downside is the significant free carrier absorption associated with heavy doping.

![Diagram](image)

(a) Impact of strain on band structure [37]

(b) Bandgap versus strain [37]

Figure 1-11: Strain engineering of Ge-on-Si platform

Another concern with Ge-on-Si platform is the pure thermal stability of pure Ge. Under foundry CMOS process, the high temperature will let Ge form SiGe alloy and
hence distort the structure. Usually, separate low temperature process is needed for Ge process, which add to cost and lower yield.

1.4.3 Silicon Photonics on CMOS Foundry Process

A new approach for photonic integration, implementing both passive and active photonic devices on zero-change commercial CMOS foundry platform, has emerged\[51\] for chip-to-chip interconnect purpose. Some unique feature of the CMOS process can be leveraged to benefit photonic devices. First of all, the massive parallel fabrication capability can lower the cost per device effectively. Moreover, strong index contrast between Silicon substrate and oxide or nitride enables higher performance and more compact devices, hence denser integration. Additionally, very high precision (1nm) and expensive lithography of CMOS process can be implemented in optical devices with minimal marginal cost. Therefore, sophisticated micro-structures can be implemented in large numbers, including ring filter banks and compact multimode interferometer.

The first challenge of building photonic devices on a CMOS foundry platform is to create a suitable dielectric environment for optical confinement. In a bulk CMOS process, the poly-Silicon transistor gate has a significant index difference from the back-end dielectric, hence can be patterned to form waveguides. However, as shown in Fig. 1-12(a), there is only a shallow trench isolation (STI) layer between the poly-Silicon gate and Silicon substrate. The STI typically has a thickness around 400nm, which is not sufficient to provide adequate optical isolation. As a result, an as-fabricated waveguide has a propagation loss on the order of 1000 dB/cm\[7\]. A possible approach is to use post-processing to chemically etch the Silicon substrate beneath the optical waveguides, which is shown in Fig. 1-12(b). This method does not require foundry process customization and has minimal impact on the performance of electronics on the same chip. A propagation loss of around 50 dB/cm has been reported\[7\].

Photonic devices can be implemented in the SOI CMOS platform as well. An illustration of SOI platform is shown in Fig. 1-13. As shown in the figure, both body-Si and gate poly-Si can be used as waveguide or combined as strip waveguide.
(a) Bulk CMOS chip with polysilicon waveguide  
(b) Localized substrate removal

Figure 1-12: Illustration of local substrate removing

Figure 1-13: Cross section of SOI platform

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However, since the process is optimized for transistor performance, the buried oxide layer typically has thickness of 20nm to minimize thermal impedance. Such a thin oxide layer does not provide sufficient isolation from the Silicon substrate and post-processing is required. Apart from localized substrate removal, a faster approach for SOI is to remove all substrate and transfer the stack on optical adhesive, which has an index around 1.5. Previous work\[8\] shows deviation of performance of integrated transistors due to substrate transfer is less than 5%. The propagation loss for single crystal Silicon waveguide is around 3 dB/cm at telecommunication wavelength, which is comparable to the best result in literature. The poly-Si waveguide has a propagation loss of around 50 dB/cm, which can be leveraged for detection and modulation purpose.

![SEM of fabricated poly-Si ring in 32 nm bulk CMOS process][9]

Figure 1-14: SEM of fabricated poly-Si ring in 32 nm bulk CMOS process\[9\]

The planar precision of fabricated optical device in CMOS process is good. A scanning electron micrograph (SEM) of micro-ring on 32 nm bulk-CMOS process is shown in Fig. 1-14. It can be observed that the waveguide width in horizontal and vertical directions has a mismatch on the order of 0.1%. In a zoomed out photo, it can be observed that no visible distortion can be observed on the ring structure.

Various key photonic devices have been demonstrated on both bulk and SOI CMOS platform. The first report of integrated optics devices in bulk CMOS process has been reported in 2008\[52\]. Vertical grating couplers have been demonstrated to enable fiber coupling in and out of the chip. Basic devices, such as ring resonator filters, modulators and Mach-Zehnder interferometer have been reported. A second-
order ring resonator is reported [53] has 30 dB extinction ratio and 3 dB bandwidth of 130 GHz. Additionally, a 4 channel filter bank has been demonstrated with 240 GHz channel spacing by ring radius step of 5 nm. A complete integrated optical receiver has been reported in 45 nm SOI platform [26]. Additionally, an 8 channel ring filter bank and a transmitter with 10 dB on-off extinction at 5 GHz has been reported [54].

1.4.4 Proposal and Consideration of Using CMOS Photonics

The diagram of one coherent receiver PIC on InP of Infinera (adapted from [6]) is shown in Fig. 1-15. It can be observed that the local oscillator (LO), polarization beam splitter (PBS), optical hybrid and photo diodes are integrated on a InP chip. The ADC and DSP can potentially be integrated on a single CMOS chip. The two chips are connected by 8 high speed electronic links. In this work, the proposed structure uses CMOS platform. The block diagram is shown in Fig. 1-16. As shown in the figure, the PBS, optical hybrid, photodiodes will all be integrated with the same CMOS chip where ADC and DSP sits. Therefore, for each PM-QPSK receiver, eight high speed electrical links are eliminated. However, since it is very difficult to build laser on silicon, the local oscillator needs to be coupled from outside source. Therefore, an additional fiber coupling is required. In summary, by integrating PBS, hybrid and photodiodes with ADC and DSP, a trade between 8 electrical links for one fiber coupling site is presented.

CMOS photonics has its advantages. First of all, CMOS photonics provides the ultimate integration possibility. As show in the proposed QPSK receiver structure, CMOS photonics is capable of monolithically integrating the entire receiver. Higher order integration results in the benefit of higher energy efficiency, lower packaging cost, higher performance and reliability.

Secondly, the DSP and ADC occupied much larger real estate on chip compared with the optics devices. Fujitsu 100G receiver on 65 nm CMOS has a die size of 15 mm x 15 mm [55]. Therefore, integration of optics side by side with the electronics add only small marginal cost to the whole receiver. Moreover, the precise lithography of 1 nm accuracy of CMOS can be leveraged to fabricate sophisticated optics structures,
Figure 1-15: Integrated coherent receiver diagram of Infinera[6]
Figure 1-16: Proposed integrated coherent receiver diagram on CMOS platform such as ring resonator banks.

However, CMOS photonics has its downside as well. First of all, as more and more devices are integrated into the same chip, the receiver becomes more complicated. As a result, the yield of DSP is tied with optical devices and might be compromised. Therefore, optics devices with large tolerance of fabrication variation, such as refractive index, waveguide thickness, need to be designed.

Secondly, most current DSP chips are made in bulk CMOS process, which is cheaper than SOI CMOS process. However, in the current literature, waveguides in zero-change bulk CMOS have a loss of around 50 dB/cm\textsuperscript{[56]}, which make it very difficult to make high performance optical devices. On the contrary, a low waveguide loss of 3 dB/cm has been demonstrated in SOI CMOS process. As a result, the DSP circuits need to be migrated to more expensive SOI process in order to be integrated together with optics devices. This may weaken the price advantage from less packaging of the proposed work.
Thirdly, traditional photodiodes which are typically built on a III-V platform and rely on direct bandgap transition generally have much better performance than all-silicon photodiodes. Since silicon is indirect bandgap and transparent in telecommunication wavelength, developing a photodiode with adequate responsivity, dark current and bandwidth remains a challenge.

On the transmitter side, traditionally, nested Mach-Zehnder modulators are used for QPSK modulation, which occupy large real estate. Therefore, the cost associated with larger footprint further diminish the price advantage of CMOS photonics. Additionally, all silicon modulators with telecom grade is a design challenge as well. In long haul optical links, each EDFA adds at least 3 dB of noise figure. Therefore, a high extinction ratio is expected for telecom modulators.

Additionally, for DWDM-OOK system, a larger number of wavelength channels are needed than for QPSK system for the same data rate. The current state-of-the-art Infinera system has 40 wavelengths with 50 GHz channel spacing. Two interleaving AWGs are used as wavelength filters in this system, which has large footprint. Much smaller filters can be realized by ring resonator filters. However, resonance structures intensify the field density in cavity, which could trigger non-linear effect, such as two photon absorption (TPA). These non-linear effect could cause distortion to the signal.

In summary, by using CMOS photonics, higher order of integration can be realized. Various advantages, such as lower packaging cost, higher energy efficiency can be achieved. However, several issues need to be addressed before CMOS-based telecommunication optics link can reach similar performance as current state-of-the-art. Compact and robust 90 degree hybrid, QPSK modulator and high performance all silicon detectors need to be designed. This thesis is focused on building a QPSK receiver on CMOS platform. The main challenges are developing high performance, fabrication error tolerant 90 degree hybrids and photodetectors at 1550 nm. Additionally, some preliminary design of QPSK modulator will be presented. System simulation is also needed to figure out the requirement for individual devices.
1.5 90 Degree Hybrid

In integrated optical communications, in order to enable coherent mixing of signal and local oscillator, it is essential to have a 90 degree hybrid device. There are a few ways to fulfill the requirement in integrated optics. Most of them fall into two categories: multimode interferometer (MMI) and cascaded directional couplers. There are other approaches, such as array waveguide grating (AWG). A simple comparison these approaches is shown in Table. 1.5.

The following sections will first describe physics of cascaded directional couplers and MMIs. Afterwards, a literature review and a comparison between these two approaches will be presented.

### 1.5.1 Physics of Cascaded Directional Couplers

The behavior of a single directional coupler can be described with coupling mode theory. A simple four-port directional coupler is shown in Fig. A-2, which is basically two parallel identical waveguides. Assuming a beam with normalized amplitude is injected into to waveguide 1 and no input from waveguide 2. The output can be represented by Eqn. A.14. The energy is oscillating between two waveguides. Additionally, since $|a_1(z)|^2 + |a_2(z)|^2 = |a_1(0)|^2$, total energy is conserved. At a specific length so that $\kappa z = \frac{\pi}{4}$, the optic power is equally split between two outputs. The
propagation matrix for such a directional coupler is shown in Eqn. A.15

\[ a_1(z) = -a_1(0) \cos(\kappa z)e^{-j\beta z} \]
\[ a_2(z) = ja_1(0) \sin(\kappa z)e^{-j\beta z} \] (1.7)

\[
\begin{vmatrix}
  a_1(L) \\
  a_2(L)
\end{vmatrix} = \frac{\sqrt{2}}{2} \begin{vmatrix}
  -1 & j \\
  j & -1
\end{vmatrix} \begin{vmatrix}
  a_1(0) \\
  a_2(0)
\end{vmatrix} \] (1.8)

A schematics of cascaded directional couplers based 90 degree hybrid is shown in Fig. 1-18. Two stages of directional couplers are cross connected with 4 waveguide with equal length. A 90 degree phase shifter is inserted in one of the waveguide.

By applying the propagation matrix in Eqn. A.15, the result in each stage is shown
in the table below. All intensity information is ignored for simplicity.

\[ a_1 = -E_1 \quad b_1 = -E_1 \quad c_1 = E_1 + jE_2 \]
\[ a_2 = jE_1 \quad b_2 = E_2 \quad c_2 = -jE_1 - E_2 \]
\[ a_3 = jE_2 \quad b_3 = jE_1 \quad c_3 = -jE_1 - jE_2 \]
\[ a_4 = -E_2 \quad b_4 = -E_2 \quad c_4 = -E_1 + E_2 \]

Therefore:

\[ ||c_1|^2 - |c_2|^2| = |2\sqrt{P_1 P_2}\cos(\phi_1 - \phi_2)| \]
\[ ||c_3|^2 - |c_4|^2| = |2\sqrt{P_1 P_2}\sin(\phi_1 - \phi_2)| \] (1.9)

As shown in Eqn. 1.9, the output matches the definition of detected QPSK modulated signal specified in Eqn. 1.4 and Eqn. 1.5.

### 1.5.2 Derivation of Self-imaging by Mode Interference

An MMI relies on self imaging to realize power splitting and phase shifting. For a 90 degree hybrid, a 2x4 MMI can be used. For a waveguide supporting a large number of modes, the input light can produce an n-fold image of itself at designated length \[\text{\textsuperscript{III}}\], which is called self-imaging. A two-dimensional representation of a typical MMI is shown in Fig. 1-19. In a high index contrast waveguide, for mode number \(\nu\), the propagation constant can be approximately described by Eqn. B.1. Therefore, the difference of propagation constants between fundamental mode and each excited mode can be described. The beat length is defined in Eqn. B.2 and B.3. Here, \(\beta_0\) and \(\beta_\nu\) represent the propagation constants for fundamental and \(\nu_{th}\) order mode respectively. \(W_e\) is the effective width of the MMI region, which can be approximated to the physical width of the MMI in the case of high index contrast. The length \(L_\pi\) is defined as the beat length between the two lowest order modes. The other parameters have their conventional meanings.
\[ \beta_\nu = k_0 n_r - \frac{(\nu + 1)^2 \pi \lambda_0}{4 n_r W_e^2} \tag{1.10} \]
\[ L_\pi = \frac{\pi}{\beta_0 - \beta_1} \tag{1.11} \]
\[ \beta_0 - \beta_\nu = \frac{\nu(\nu + 2)\pi}{3L_\pi} \tag{1.12} \]

The general multi-fold image appears at intermediate length between 0 and \( L_\pi \). The position and phase of each image are analytically described in [12] using Fourier analysis. As shown in Fig. [B-1] the real MMI width is defined in region [0,W]. A virtual MMI width is added in the region [-W,0] and the anti-symmetric image of the input field is assumed in this region. Therefore, the total input field(real and virtual) can be represented by the superposition of anti-symmetric modes of the extended MMI region. At a minimal distance of \( \frac{3 L_\pi}{N} \), \( N \) images with equal intensity will be formed. If they are numbered by \( q = 0, 1, \cdots, N - 1 \), the positions \( x_q \) and phases \( \phi_q \) can be described by Eqn. [B.8]. As shown in Fig. [B-2] a total of 2\( N \) images will be produced in the extended MMI region, when half of them are inverted. Therefore, in the real MMI region, there will be \( N \) images.
Figure 1-20: Input field decomposed into superposition of anti-symmetric modes of extended width\textsuperscript{12}

Figure 1-21: Output field positions\textsuperscript{12}
\[ x_q = (2q - N) \frac{W}{N} \]
\[ \phi_q = q(N - q) \frac{\pi}{N} \] (1.13)

A 4×4 coupler is shown in Fig. B-4. Two inputs are introduced into the MMI from port 4 and 2 respectively. The output delays are shown in Table B. It can be observed that the 4 phases are 90 degrees away from each other. Therefore, the 4×4 MMI is capable for serving as a 90 degree hybrid, which is essential in QPSK communication link.

\[ \phi_{41} = \pi \quad \phi_{21} = \frac{3}{4} \pi \quad \phi_{21} - \phi_{41} = -\frac{1}{4} \pi \]
\[ \phi_{42} = -\frac{3}{4} \pi \quad \phi_{22} = \pi \quad \phi_{22} - \phi_{42} = \frac{5}{4} \pi \]
\[ \phi_{43} = \frac{3}{4} \pi \quad \phi_{23} = \pi \quad \phi_{23} - \phi_{43} = \frac{1}{4} \pi \]
\[ \phi_{44} = \pi \quad \phi_{24} = \frac{7}{4} \pi \quad \phi_{24} - \phi_{44} = \frac{3}{4} \pi \]

1.5.3 Current Research in Integrated MMI Design

The multimode interferometer (MMI) has been a widely used as a 90 degree hybrid because of its simple geometry and small footprint. Additionally, the MMI is fully passive, which is appealing in optical telecommunication where power budget is essential.
Table 1.3: Comparison of a few MMI design

<table>
<thead>
<tr>
<th></th>
<th>Typical</th>
<th>Low index contrast</th>
<th>Non-straight</th>
</tr>
</thead>
<tbody>
<tr>
<td>CMRR (dB)</td>
<td>15</td>
<td>20</td>
<td>20</td>
</tr>
<tr>
<td>Phase accuracy(°)</td>
<td>3</td>
<td>5</td>
<td>5</td>
</tr>
<tr>
<td>Footprint(μm × μm)</td>
<td>21.6×978</td>
<td>7.7×115.5</td>
<td>9.02×227</td>
</tr>
</tbody>
</table>

The first integration-compatible MMI design is implemented in GaInAsP/InP platform[57]. It achieved a splitting ratio of 97:99:104:99, insertion of less than 1dB and phase deviation of less than 3 degree. The interferometry region is compact and the footprint is 21.6×978μm.

There are reports on SOI based MMI devices in recently year. In Zimmerman’s paper[58], a relative large(length of 2400μm) device is designed. It achieved a phase error less than 5% and power imbalance less than 0.5dB over the entire C band. In Halir’s paper[61], a shallower etched SOI region is used for interferometric region to reduce index contrast and alleviate phase error. The design is compact(7.7×115.5μm). It achieved CMRR of less than 20dB and phase error better than 5 degrees on a wavelength range from 1510nm to 1560nm.

There is also unconventional MMI design, where the multimode interferometric region is no longer a straight waveguide. A linear taper is introduced to exploit restricted excitation [64]. Under restricted excitation, the light is injected into designated position of a MMI, so that some of the mode will not be excited. As a result, beat length and device footprint is shorter. It achieved power imbalance less than 1dB, CMRR of 20dB and phase deviation less than 5 degree for the C-band spectral range.

1.5.4 Comparison between MMI and Cascaded Directional Couplers

Both MMI and cascaded directional couplers are widely used to serve as 90 degree hybrids and both them of them have their strength and weakness. MMI has a much
simpler geometry and smaller footprint. Therefore, it is easier to fabricate and is capable of higher integration density. Additionally, the MMI is a fully passive device, which is very attractive, since the link energy efficiency is critical. However, since the performance of the MMI relies on interference, it is inherently wavelength sensitive. On the other hand, cascaded directional couplers are relatively complicated system and requires much more real estate on chip. Additionally, the two stages of directional couplers are connected with waveguides and the phase difference between I and Q channels are entirely dependent on the relative phase in these waveguides. Therefore, the performance is very sensitive to intra-chip fabrication defects and temperature variation. In reality, thermal tuning is necessary for compensating all the phase deviations, which make it an active device. However, at the expense of energy and space, cascaded directional couplers typically provides better power splitting and phase accuracy. Additionally, by using adiabatic directional coupler, it can be relatively broadband.

### 1.6 CMOS compatible infra-red photodetector

#### 1.6.1 Current Literature

Aside from 90 degree hybrids, photodetectors are critical components of QPSK optical telecommunication links as well. Photodetectors in CMOS or bulk Silicon process can be a major challenge, since the bandgap of Silicon is around 1.1 eV, which corresponds to a wavelength of 1.13$\mu$m. Conventional one photon bandgap transition is impossible in telecommunication wavelength. Alternative approaches include SiGe, pure Ge, two photon absorption (TPA) and defect state. A summary of different approaches is shown in table below.

<table>
<thead>
<tr>
<th></th>
<th>SiGe$^{[70]}$</th>
<th>Ge$^{[37]}$</th>
<th>TPA$^{[71]}$</th>
<th>Defect State$^{[38]}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Responsivity</td>
<td>0.4A/W at 1.3$\mu$m</td>
<td>1.1A/W</td>
<td>0.125A/W</td>
<td>0.21A/W</td>
</tr>
<tr>
<td>Dark current</td>
<td>27pA/$\mu$m$^2$</td>
<td>1$\mu$A</td>
<td>15pA</td>
<td>30nA</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>1.5GHz</td>
<td>32GHz</td>
<td>0.1GHz</td>
<td>3G</td>
</tr>
</tbody>
</table>
Germanium has a lower bandgap of 0.8 eV, hence is highly absorptive below 1550 nm. Since a SiGe alloy is available in CMOS process for strain engineering purpose, it is possible to build photodetectors in telecommunication wavelength with SiGe. Literature shows works of this approach [72, 73, 70] and with SiGeC material system [74].

Aside from SiGe, pure Ge can be shifted to direct bandgag by strain engineering and achieve good band to band transition rate. However, a high dark current density is usually associated with Ge photodetectors, since there is no high-quality passivation layer for Ge [75]. But it can be mitigated by decreasing the size of the device, given the signal can be nicely coupled into the device. Additionally, high lattice mismatch limits the thickness of epitaxial growth of Ge on Si. While this helps to reduce carrier transit time, the absorption and responsivity are affected. Additionally, device capacitance is increasing with small junction thickness, which degrades bandwidth of the photodiode. Making waveguides perpendicular to the junction address this problem by enabling independent optimization of responsivity and bandwidth. Significant effort can be found in the literature on Ge-on-Si photodetector and some of them have performance approaching III-V devices [76, 77, 78, 79, 80, 81].

Additionally, there are also efforts in exploiting two photon absorption (TPA) for below bandgap absorption [82, 83, 84, 71]. However, they typically requires high optical power and have low quantum efficiency. Therefore, this is not a desirable approach.

Another way to solve this is to use intra-band defect states, which requires large density of grain boundary, or two-photon absorption, which requires a large light intensity. These requirement points to resonant designs with poly-Si. A popular choice is ring resonators, which has been well investigated. Ring resonator has a small footprint. Rings with radii of 10 \( \mu m \) on CMOS platform has been demonstrated [9]. Additionally, Q-factor on the order of a million has been reported [85]. The basic theory of ring resonator will be introduced in the appendix.

The basic principles of photodiodes will be described in the following section.
1.6.2 Basic Physics of Photodiodes

A photodiode converts light into electric potential. PIN junctions are common structures for photodiodes in semiconductor industry. A simple schematic of a PIN diode is shown in Fig. 1-23. The two ends of the diode are P and N doped respectively and the section in between is intrinsic. Their widths are represented by $L_p$, $L_N$ and $W_i$ respectively. When operating as a photodiode, the PIN diode is usually reverse biased. As shown in Fig. 1-24, when a photon with energy higher than the bandgap is absorbed by the diode, an electron is excited into the conduction band, leaving a hole in valence band. The electron-hole pair is then separated by the electric field. The electron drifts to the N-doped end, while the hole drifts to the P-doped end.

![Schematic of a simple PIN diode](image)

Figure 1-23: Schematic of a simple PIN diode

To describe the performance of a photodiode, three most important figures of merit are used: quantum efficiency, dark current and bandwidth. Quantum efficiency is the number of electron-hole pairs generated for every incident photon. It is often defined by Eqn. (1.14) where $I_{ph}$ is the photocurrent, $h\nu$ is the photon energy of the incident light, $q$ is the charge of an electron, $P_{opt}$ is the incident light power. Sometimes responsivity is used instead of quantum efficiency. As shown in Eqn. (1.15), it is defined as the ratio of photocurrent over incident light power.
Figure 1-24: Band diagram of a simple PIN diode

\[ \eta = \frac{I_{ph} h\nu}{q P_{opt}} \]  

(1.14)

\[ R = \frac{I_{ph}}{P_{opt}} \approx \frac{\eta \lambda (\mu m)}{1.24} \]  

(1.15)

The dark current is the electric current when no light is incident on the detector. In optical communication, light-generated current is the desired signal. The noise in the dark current contributes to the overall sensitivity of the receiver. Therefore, the dark current should be minimized for better sensitivity. The source for dark current can be Auger generation and defect-assisted generation. From detailed balance, these generation are equal to recombination and can be estimated with minority carrier concentration and lifetime. The equation is shown in Eqn. 1.16. Here \( A \) is cross-sectional area of the diode, \( \Delta n_{po} \) and \( \Delta p_{qa} \) are excess minority carrier concentration, \( \tau_p \) and \( \tau_n \) are minority carrier life time, \( G_0 \) is the generation rate in intrinsic region.

\[ I_{dark} = qA \left( \frac{\Delta n_{po}}{\tau_p} L_p + \frac{\Delta p_{qa}}{\tau_n} L_n + G_0 W_i \right) \]  

(1.16)
The bandwidth of a photodiode describes how fast the diode can operate under a modulated optical signal. It is usually limited by the transit time across the depletion region and RC limit of the diode. The 3dB roll-off frequency can be calculated with Eqn. 1.17. Here $R$ and $C$ are resistance and capacitance of the diode respectively, $\tau_{tr}$ is the transit time.

$$f_{3dB} = \frac{1}{2\pi \sqrt{(RC)^2 + (\tau_{tr}/2.636)^2}}$$

For all-silicon photodetector at telecommunication wavelength, traditional band to band absorption is impossible since photon energy is smaller then the band gap. Instead, poly-silicon is used and defect states are responsible for electron-hole pair generation. However, large density of defect states could lead to large recombination rate and hence reduces quantum efficiency. Additionally, small carrier mobility in poly-silicon could potentially limit the bandwidth due to large transit time.

1.6.3 Challenge and Goal

Current CMOS compatible photodiodes all have their own weakness. The SiGe detectors only work for wavelengths shorter than 1300 nm[70]. Since the most common telecommunication wavelength is 1550 nm, the application of SiGe detectors are limited. Germanium detectors have comparable performance to mature III-V photodetectors[37]. However, it takes separate steps to handle Germanium on CMOS, which increases the complexity of the whole process. The TPA detectors have limited performance[71]. Defect based detectors have adequate performance, but the example shown in literature[38] is fabricated in an emulated DRAM process, instead of a CMOS process.

According to the specification from Ciena [86], a photodetector needs to be able to detect input power of -30 dBm. If the dark current is around 100 nA, the responsivity should be more than 0.1 A/W to obtain a 3 dB difference between light and dark. Additionally, a bandwidth of 20 GHz is desired. A poly-silicon infra-red photodetector
needs to be designed to fulfil these requirement, under the constraint of zero-change CMOS process.

A few challenges exist for designing such a photodiode in CMOS process. First of all, there is a large uncertainty of the loss of poly-silicon waveguide. Previous work in our group[87] shows a variation from 50 to 130 dB/cm at 1220 nm. This makes it difficult to achieve critical coupling. The ring resonators should be designed so that they are robust to loss variation.

Additionally, in zero-change CMOS process, the top surface of the poly-silicon layer is not polished. As a result, surface roughness may contribute to the loss. It is difficult to determine the relative strength of surface scattering and material absorption. This could potentially limit the quantum efficiency of the photodetectors.

Another challenge is to create PN junctions for carrier extraction. Since the heavily doped regions have very high loss due to free carrier absorption, they should be away from the waveguides. However, long distance leads to long transit time and slower detectors. Additionally, more carriers will recombine before reaching heavily doped regions and hence lower the quantum efficiency. The photodetectors need to balance between this trade-off.

The design of the poly-silicon photodetectors will address these challenges. The design and measurement results will be presented in Chapter 3.

\section*{1.7 Integrated QPSK Modulator}

Integrated QPSK modulators are critical components of optical QPSK link. A QPSK receiver is usually constructed by two phase modulators with 90 degree phase shift between them. Most integrated phase modulators are based on Mach-Zehnder modulators (MZM) or ring resonators. These two approaches will be presented in the following sections.
1.7.1 Mach-Zedner Phase Modulator

Mach-Zedner interferometer can be used for both intensity and phase modulation. The refractive index can be controlled by applying electric field. As shown in Fig. 1-25, the two arms are fed with RF signals of opposite signs and hence creates phase shift of opposite signs. This is called push-pull operation. As a result, the combined optical signal can be represented in Eqn. 1.18. The amplitude of the modulated signal is proportional to \( \cos(\theta_s(t)) \), where \( \theta_s(t) \) is the phase shift caused by \( V_s(t) \). If the RF signals are DC biased at zero amplitude point, an additional AC RF signal will produce light with \( \pi \) phase difference and equal intensity.

![Figure 1-25: Basic Mach-Zedner Modulator](image)
\[ E_{\text{out}} = E^{j(\omega t + \theta_s(t))} + E^{j(\omega t - \theta_s(t))} \]
\[ = 2E^{j\omega t \cos(\theta_s(t))} \quad (1.18) \]

Optical phase modulators by MZM are first implemented in LiNbO$_3$ and III-V semiconductors. In recent years, there are implementation in silicon as well. The refractive index of silicon waveguide is controlled by carrier density. However, due to weak electro-refractive effect in silicon, a long device is required. A recent publication\[89\] demonstrated a MZM phase modulator on silicon. With a $V_\pi$ of 3.1V, a phase shifter of 6 mm is used. Another problem related with long device is high insertion loss. In this paper, an insertion loss of 9 dB is reported. Micro-ring based devices have much smaller footprint and potentially less insertion loss.

### 1.7.2 Micro-ring Based Phase Modulator

Most of the current research in micro-ring based phase modulators uses resonance shift\[90\]\[13\]91\]92. An illustration is shown in Fig. 1-26. In an over-coupled ring resonator, the output experiences an intensity dip across the resonance wavelength, as well as a $2\pi$ phase shift. The resonance wavelength can be modulated, so that a certain wavelength experiences alternating phases but same intensity. A QPSK modulator can be built by connecting two phase modulators with 90 degree phase difference. An example is shown in Fig. 1-27.

However, phase modulation by this scheme introduces large insertion loss inevitably. For the structure shown in Fig. 1-27, a ring insertion loss of 7.5 dB is reported\[13\]. Additionally, precise control is required for accurate phase response, since the phase shift is sensitive to the resonance wavelength.

To address the problems above, an approach based on double-coupled ring resonator was introduced\[14\]. As shown in Fig. 1-28, the output power alternates between through and drop ports, depending on the resonance condition. Meanwhile, a $\pi$ phase shift exits between the two outputs. The paper proposed to feed the through
Figure 1-26: Illustration of single ring phase modulation.\textsuperscript{[13]}

Figure 1-27: Microring-based QPSK modulator.\textsuperscript{[13]}
port output into the drop port, so that it may serve as a phase modulator.

![Ring Frequency Response Diagram]

Figure 1-28: Spectral amplitude and phase response of double-coupled ring.

However, there are some problems in this paper. First of all, instead of effective index, this paper uses group index to calculate resonant wavelength. This will result in wrong phase response and should be corrected. Secondly, the analytical solution for wavelength response in this paper is not accurate. This will result in different ways of modulation. Finally, this modulation scheme needs to be adapted into zero-change CMOS process. The detailed derivation and design will be presented in Chapter 4.

### 1.8 Conclusion

This Chapter reviews the motivation and background for integrated photonic QPSK receiver in CMOS process. Future optic communication link requires higher spectral efficiency, and lower cost. Integration provides a promising solution, due to its better performance, less energy consumption and less packaging cost. Among the current
integration platforms, CMOS photonics appears to be a more attractive choice. Its ability to integrate electronics and optics in the same chip provides higher order of integration. As a result, lower energy loss and package cost is expected. However, several challenges remain. Due to fact that Silicon is transparent in telecommunication wavelength, it is not trivial to design a photodetector. Additionally, all the devices have to be tolerant to thickness variation of CMOS foundry process.

This thesis will describe the effort in designing a QPSK transceiver on CMOS platform. CMOS compatible 90 degree hybrid, photodetector and QPSK modulator design will be presented. Chapter 2 will focus on the design and measurement of 90 degree hybrid. Two main approaches are investigated: multimode interferometer (MMI) and cascaded directional couplers (CDC). Chapter 3 will focus on the design and characterization of a 1550 nm defect based ring resonant photodetector. Design and simulation for integrated QPSK modulators based on double coupled ring resonators will be presented in chapter 4. Finally, chapter 5 will summarize this thesis and mention possible future development.
Chapter 2

90 Degree Hybrid

The 90 degree hybrid is a critical component in a coherent communication system. It mixes the signal and local oscillator with 4 different phase delays 90 degree from each other. In this Chapter, two different designs are discussed: multimode interferometer (MMI) and cascaded directional couplers with a focus on MMI. The chapter starts with the design of MMI in a foundry process. Afterwards, numerical simulation results for optimization of MMI are presented. Figure of merits (FOMs) are defined for intensity and phase balance respectively. Additionally, a test structure for MMI is presented. Finally, the characterization of fabricated MMI are presented and discussed. In the end of this chapter, a simple discussion of cascaded directional couplers is presented.

2.1 Multimode Interferometer

2.1.1 MMI Design

A multimode interferometer (MMI) utilized self-imaging to produce multi-fold images of the input field with different phase delays. By carefully designing the geometry, a MMI can serve as a 90 degree hybrid. In this thesis, MMIs are designed to be implemented in Micron emulated DRAM process and IBM 12SOI process.

The dielectric environment for passive devices in Micron emulated DRAM process
Figure 2-1: Design of the MMI in Micron emulated DRAM process
is shown in Fig. 3-1(a). The gate poly-silicon is used as waveguide core. The core is surrounded by a thin conformal layer of Si$_3$N$_4$. The cladding is formed by SiO$_2$. In this process, the thickness of poly-Silicon core is 225 nm.

The bird’s eye view of the MMI design is shown in Fig. 3-1(c). It consists of a wide multimode region with 4 single mode ports attached to each end. The single mode waveguide was designed to barely support only the first order mode, so that bending loss is minimized. In this case, the width is determined to be 440 nm. The mode profile of a single mode waveguide is shown in 3-1(b).

The width of the MMI was set as 3.8 $\mu$m to be wide enough to support 10 lateral modes. The required length for four-fold image is expected at a length of $3\pi/4(\beta_0 - \beta_1)$. In this equation, $\beta_0$ and $\beta_1$ represent the propagation constants of fundamental mode and first order mode of multimode region respectively. The length of the MMI is then estimated to be 27.6 $\mu$m. Two mode profiles showing the multi-fold images of input from port 1 and port 3 are presented in Fig. 3-1(d) and 3-1(e) respectively.

As shown in Fig. 3-1(c), $a$ is the distance between edge of MMI region and the outer waveguides, $b$ is the spacing between outer and inner waveguide and $c$ is the spacing between two inner waveguides. The input and output ports are positioned so that $c$ is roughly twice of $a$. This relationship is described in [12]. Additionally, the distance between adjacent ports should be large enough to avoid cross talk. From numerical simulation, a gap of 0.4 $\mu$m is sufficient to keep coupling negligible. After all the constraint, the range of $a$ and $b$ can be determined.

### 2.1.2 Numerical Simulation Results

To fine tune the structure for optimum performance, numerical simulation is required. Commercial software FIMMWAVE and FIMMPROP were used to scan all geometries similar to the theoretical optimum structures.

First of all, the value of $a$, $b$ and $c$ are tuned. Under the constraint of fixed MMI region width and waveguide width, only two variables are independent. Therefore, a scan was needed on a two dimensional array of $a$ and $b$. Afterwards, for each set of $a$
and $b$, a scan of MMI region length was performed around the theoretical optimum point.

For each simulation, a complete scattering matrix of the 4x4 MMI is generated by FIMMPROP. However, only part of the information is relevant. Due to the symmetry of the MMI region, only two input ports are needed. Additionally, reflection from the input ports are not important here. Therefore, for each simulation, 8 elements of the scattering matrix were recorded. They correspond to input from port 1 and 3 on one side and output from all ports on the other side. In this thesis, they are defined as $S_{11}$, $S_{12}$, $S_{13}$, $S_{14}$, $S_{31}$, $S_{32}$, $S_{33}$, $S_{34}$ respectively for convenience. The refractive indices for poly-Si, nitride and silicon dioxide are set as 3.53, 1.9 and 1.445 respectively for the numerical simulation in this thesis.

In order to quantitatively evaluate the performance of the 90 degree hybrid, figures of merit (FOM) were created for intensity and phase balance respectively. They are defined in Eqn. 2.1 and Eqn. 2.2 respectively. Here $\sigma_{int}$ is the standard deviation of the population ($(|S_{11}| + |S_{31}|)^2$, $(|S_{12}| + |S_{32}|)^2$, $(|S_{13}| + |S_{33}|)^2$, $(|S_{14}| + |S_{34}|)^2$). Typical performance in literature [58] is intensity standard deviation of 1 dB and phase standard deviation of 5°. This corresponds to $FOM_{int} = 18$ and $FOM_{ph} = 2.87$ respectively.

$$FOM_{int} = \frac{\sum_{j=1,2,3,4} |S_{ij}|^2}{\sigma_{int}}$$ (2.1)

$$FOM_{ph} = \frac{1}{\sum_{i=1,2,3,4} |\cos(ph_i - ph_j)|}$$ (2.2)

where:

$$ph_i = \text{arg}(S_{1i}) - \text{arg}(S_{3i})$$

For Micron D1L designs, the global maximum $FOM_{int}$ is 230 in a structure with
Figure 2-2: FOM from scan of possible MMI structures in Micron emulated DRAM process
The global maximum $FOM_{ph}$ is 25 in a structure with $a = 0.36\,\mu m$, $b = 0.4\,\mu m$ and length of $25\,\mu m$. Part of the simulation result is shown in Fig. 2-2.

For a configuration with $a = 0.255\,\mu m$ and $b = 0.43\,\mu m$, the curves of FOMs versus MMI region length are presented in Fig. 2-2(a) and 2-2(b). It can be observed that FOM for intensity peaks at 27.3 $\mu m$. The normalized power at this point are 1, 0.9039, 0.8902 and 1.0192 for four ports respectively. The FOM for phase peaks at 26.9 $\mu m$. The normalized phase at this point are 0, -89.3, 85.7 and 171 degrees respectively.

Additionally, the FOMs at MMI region length of 27.6 $\mu m$ of a full range of $a$ and $b$ combinations are presented in Fig. 2-2(c) and 2-2(d). It can be observed that the structures with higher FOMs in both intensity and phase reside in a region where $a \approx 0.25\,\mu m$ and $b \approx 0.52\,\mu m$. Under this condition, the separation of the inner two waveguides can be determined to be 0.5 $\mu m$, which is twice of $a$ as predicted by theory.

In addition to Micron emulated DRAM process, MMIs are designed and implemented on IBM 12SOI process as well. In this process, crystal body Silicon is used as the waveguide core instead of poly-Silicon. The MMI region width is set to be 5 $\mu m$. A set of design curves are shown in Fig. 2-3. Fig. 2-3(a) and Fig. 2-3(b) show the relationship between figure of merits and length of the multimode region for $a = 0.24\,\mu m$ and $b = 0.58\,\mu m$. Fig. 2-3(c) and Fig. 2-3(d) present how the FOMs change with positioning of input and output ports at fix MMI region length of 34.3 $\mu m$.

Self-imaging is the result of mode beating. From [11], in the ideal situation, the difference between propagation constants of $\nu_{th}$ order mode and fundamental mode is proportional to $\nu(\nu + 2)$. Two approximations are required for this to be true. First of all, the multimode waveguide has a very high index contrast. As a result, the modes are closely confined within the core and can be approximated by metallic boundary condition. Therefore, the wave vector in transverse direction can be written as $k_{xu} = (\nu + 1)\pi/W$, where $W$ is the effective width of fundamental mode.

The second approximation is, the wave vectors in transverse direction is much
Figure 2-3: FOM from scan of possible MMI structures in IBM 12SOI process
smaller than the propagation constants. Under this assumption, the propagation constant can be related to wave vectors in transverse direction by first order Taylor expansion, which is shown in Eqn. 2.4. Therefore, since $k_{xν}$ increases with the order of modes $ν$, this approximation becomes less accurate for higher order modes.

\[
β_ν = \sqrt{k_0^2n_r^2 - k_{xν}^2}
\]

(2.3)

\[
≈ k_0n_r - \frac{k_{xν}^2}{2k_0n_r}
\]

(2.4)

In Fig. 2-4(a) and Fig. 2-4(b), the difference of propagation constants between higher order modes and the fundamental mode are plotted for D1L and EOS18 chips respectively. Analytical results from approximation are compared against the result from numerical simulation. It can be observed that approximation from second order Taylor expansion gives more accurate results than first order approximation.

This raises a trade-off in choosing MMI width for 90 degree hybrid. A wider MMI support larger number of modes and provide better resemblance of the input mode. However, higher order modes have less accurate beating length and hence deteriorate self-imaging. Therefore, the optimum width of MMI need to carefully investigated.

Additionally, since the D1S and EOS18 chips show very similar curves, it is reasonable to expect that they have similar performance.
In a foundry process, the thickness of gate silicon or body silicon has some variation. Therefore, the performance of the MMI will be affected. The FOMs for a MMI with thickness of ±10 nm from the nominal 225 nm are shown in Fig. 2-5(a) and Fig. 2-5(b). It can be observed that both the intensity and phase FOMs peaks move to longer length with increasing thickness.

From Eqn. 2.4, the difference of propagation constants between fundamental mode and higher order modes are inversely proportional to the effective index $n_r$ of the center ridge. As the thickness of poly-silicon core increases, $n_r$ increases. Therefore, beat length between modes increases. This explains observation in Fig. 2-5(a) and Fig. 2-5(b).

As shown in Table. 2.1.2, the normalized change in MMI length and $n_r$ is very close to each other. This is consistent with the theory mentioned above.

<table>
<thead>
<tr>
<th>Thickness (nm)</th>
<th>215</th>
<th>225</th>
<th>235</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\beta_0$ (1/µm)</td>
<td>11.72537</td>
<td>11.83222</td>
<td>11.97846</td>
</tr>
<tr>
<td>$\beta_1$ (1/µm)</td>
<td>11.63921</td>
<td>11.74682</td>
<td>11.89409</td>
</tr>
<tr>
<td>MMI Length (µm)</td>
<td>27.3467</td>
<td>27.5901</td>
<td>27.9269</td>
</tr>
<tr>
<td>Normalize</td>
<td>0.9912</td>
<td>1</td>
<td>1.0122</td>
</tr>
<tr>
<td>$n_r$</td>
<td>2.898818</td>
<td>2.930868</td>
<td>2.96095</td>
</tr>
<tr>
<td>Normalize</td>
<td>0.9891</td>
<td>1</td>
<td>1.0103</td>
</tr>
</tbody>
</table>
Figure 2-6: Delay line structure for measurement
2.1.3 Testing Structures

Since the phase information can not be easily detected, a delay line structure was
designed to characterize the performance of 4x4 MMI on both intensity and phase
balance. As shown in Fig. 2-6, a 2x2 MMI is connected to the input of the test
structure. The two outputs are connected to port 1 and 3 of the 4x4 MMI respectively.
One of the two connecting waveguides are intentionally elongated to introduce a phase
difference between two arms. Therefore, as the wavelength of the input light is tuned
over a wavelength range, the effective index of the waveguides is tuned and hence the
phase difference.

Within a small wavelength range, the scattering matrix of a MMI does not change
significantly. The simulated performance of a D1L MMI with $a = 0.255\mu m$, $b =
0.43\mu m$ and length of 26.9$\mu m$ is shown in Table. 2.1 at three difference wavelengths. It
can be observed that both the intensity and phase behavior has a variation below 5%.
Therefore, it is safe to assume a constant scattering matrix within a wavelength sweep
of 4 nm. Under this assumption, tuning the wavelength is equivalent to tuning the
relative phase between two input ports. If both MMI are ideal, the phase relationship
can be described by Fig. 2-6(a). The magnitude of the four outputs are shown in Eqn.
2.5. Here $\Delta \phi$ represents the phase difference generated by the delay line. It can be
observed that the four outputs should have equal magnitude and change sinusoidally
with the phase difference. Additionally, they have a $\pi/2$ phase spacing from each
other.

\[
P_1 \propto |E|^2(1 + \cos(\phi + \Delta \phi)) \\
P_1 \propto |E|^2(1 + \cos(\phi + \Delta \phi + \pi/2)) \\
P_1 \propto |E|^2(1 + \cos(\phi + \Delta \phi + 3\pi/2)) \\
P_1 \propto |E|^2(1 + \cos(\phi + \Delta \phi + \pi))
\]  

During a measurement, a wavelength scan was performed for each input-output
<table>
<thead>
<tr>
<th>wavelength (nm)</th>
<th>1548</th>
<th>1550</th>
<th>1552</th>
</tr>
</thead>
<tbody>
<tr>
<td>normalized power 1</td>
<td>1</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>normalized power 2</td>
<td>1.13</td>
<td>1.10</td>
<td>1.08</td>
</tr>
<tr>
<td>normalized power 3</td>
<td>0.852</td>
<td>0.846</td>
<td>0.857</td>
</tr>
<tr>
<td>normalized power 4</td>
<td>1.23</td>
<td>1.21</td>
<td>1.19</td>
</tr>
<tr>
<td>FOM_int</td>
<td>25.8</td>
<td>26.8</td>
<td>29.4</td>
</tr>
<tr>
<td>normalized phase 1</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>normalized phase 2</td>
<td>-90.3</td>
<td>-89.9</td>
<td>-89.4</td>
</tr>
<tr>
<td>normalized phase 3</td>
<td>83.6</td>
<td>84.7</td>
<td>85.6</td>
</tr>
<tr>
<td>normalized phase 4</td>
<td>169.5</td>
<td>170.5</td>
<td>171.4</td>
</tr>
<tr>
<td>FOM_ph</td>
<td>2.75</td>
<td>2.98</td>
<td>3.11</td>
</tr>
</tbody>
</table>

Table 2.1: Comparison of MMI performance under different wavelength port combination. By comparing the light intensity out of four ports, the FOMs for both intensity and phase can be calculated.

The blueprint for the device was draw in Virtuoso Cadence. A screen shot of the Cadence layout of the delay line test structure is shown in Fig. 2-6(b). Since, Cadence only accept rectangles, any structure other than simple rectangles need to be constructed by a continuum of small rectangles. For structures requiring large number of rectangles, Cadence SKILL code is used to automate the drawing process, so that it is not overwhelmingly tedious. Additionally, even structures with simple geometry, such as a section of waveguide, usually consist of multiple mask layers with different dimensions. Therefore, they can also benefit from programmed drawing.

Programs were written for basic building blocks, such as bent waveguide and rings. When drawing these building blocks, instead of manually drawing rectangles at different mask layers, users only need to input required parameters to generated the corresponding structures. These programs are called parametrized-cells (p-cells).

In the delay line test structures, the grating couplers, tapered waveguides and bent waveguides are all pre-written p-cells. They are connected to create the complete delay line test structures. In this thesis, p-cells written by previous student Jason Orcutt were used.
Figure 2-7: Photos of D1S (a) wafer, (b) reticle, (c) array of test structures and (d) individual delay line
Figure 2-8: Comparison of measured and simulated results for the same MMI
2.1.4 Measurement Results and Discussion

Two batches of MMI devices were fabricated in the Micron process and named D1S and D1L respectively. In D1S, only length variation was included. Photos of D1S wafer, reticle, array of test structures and individual delay line are shown in Fig. 2-7.

Result for one device in D1S chip with length of 26.7\(\mu\)m, \(a = 0.25\mu\)m and \(b = 0.47\mu\)m is presented in Fig. 2-8. The measured results are compared against the simulated results for the same structure and good agreement can be observed. The FOMs can be calculated accordingly. The measured FOMs for intensity and phase are 24.9 and 2.03 respectively, while simulated FOMs are 28.7 and 2.85.

The FOMs for devices with length from 26.4\(\mu\)m to 27.2\(\mu\)m are shown in Fig. 2-9. Measured FOMs are plotted along with the simulated FOMs for the same structure. It can be observed that simulation and measurement results roughly agree with each other.

![Graph](image.png)

(a) FOM of intensity versus length

(b) FOM of phase versus length

Figure 2-9: Comparison of FOM from simulation and measurement

Devices on D1L chip were further optimized. The positions of four input and output waveguides were tuned for better performance. However, the devices show even worse result than D1S chip. Excessive reflection from the vertical grating couplers might be responsible for this. A Febry-Perot cavity is formed between each pair of vertical grating couplers. Its own wavelength response interferes with that of the MMI. To verify this suspicion, wavelength sweeps were performed on four waveguides,
Figure 2-10: TEM image of Micron D1S waveguide
with vertical grating couplers on two ends of each waveguide. The transmission pattern is shown in Fig. 2-11. As shown in the figure, the transmission power fluctuates sinusoidally with respect to wavelength. The peak to peak magnitude is around 4dB, which is sufficient to disturb MMI performance. It can be observed that waveguide 1 and 4 have a free spectrum range (FSR) of 62 GHz, while waveguide 2 and 3 have a FSR of 70 GHz. Since the waveguides have length of 630 $\mu$m and 566 $\mu$m and group index of 4.74, their FSR can be calculated by $c/2n_g L$. The estimated FSR are 57 GHz and 63 GHz, which are within 10% deviation from the measured values. Therefore, the reflection off the grating couplers is very likely to be the reason for the unsatisfactory results.

Figure 2-11: Wavelength response of FP cavity formed by vertical grating couplers

In summary, 4x4 MMIs are designed as 90 degree hybrid. A numerical scan was performed to reveal the relationship between the device geometry and performance. Some of the better design are fabricated in an emulated DRAM process and characterized. Performance close to the simulated optimum values was observed.
2.2 Cascaded Directional Couplers

Aside from MMI, cascaded directional couplers can be used as 90 degree hybrid as well. A schematic of the design is shown in Fig. 2-12(a). It can be observed that two stages of directional couplers are cross-linked by four waveguides. These waveguides were designed to have equal length, so that chip-wise problems, such as thickness inaccuracy of dielectric layers and fluctuating ambient temperature, have minimum impact on relative phase delay between different waveguides. The crossing of two central waveguides are designed to be vertical and perpendicular respectively to comply with design rule. Thermal tuning pads are placed near the waveguides to control the relative phase delay by heating the waveguides.

This design was implemented in IBM 12SOI process. A photo of the fabricated device is shown in Fig. 2-12(b). However, due to a drawing error, one section in the input waveguide is disconnected. Therefore, no measurement data has been collected so far.

2.3 Conclusion

In this chapter, the design and characterisation of 90 degree hybrid on CMOS platform is presented. The discussion is focused on MMI. Designs are made to be implemented in both Micron emulated DRAM process and IBM 12SOI process. During design process, numerical simulation is used for fine-tuning the structures. In order to quantify the performance of the 90 degree hybrids, figure of merits are defined for intensity and phase balance respectively. To enable measurement of phase information, delay line structures are designed. The devices on Micron D1S process is characterized and the FOMs are calculated accordingly. The measurement results are compared against simulation results and good match can be observed. After the discussion of MMI, the cascaded directional couplers are briefly mentioned.

In the proposed integrated QPSK receiver, the output of 90 degree hybrids would be coupled to photodetectors on the same CMOS chip. The design, implementation
Figure 2-12: Illustration of cascaded directional couplers
and characterization of the such CMOS photodetectors will be presented in the next chapter.
Chapter 3

Poly-Silicon Ring Resonant Photodetector

Photodetectors are essential components of optical communication links. Integration of photodiodes into CMOS chips enables monolithic optical receivers. However, since the bandgap of silicon is higher than the photon energy at telecommunication wavelength, it is impossible to design photodetectors using conventional band to band absorption. A possible solution is using defect states in poly-silicon to absorb sub-bandgap photons. A ring resonator structure may be introduced to increase absorption and hence reduce the device size.

In this chapter, the design of a ring resonant poly-silicon photodetector is presented first. Afterwards, measurement results for dark current, photocurrent, and bandwidth are shown and discussed respectively. Finally, theoretical models are discussed in attempt to explain the measurement results.

3.1 Design

3.1.1 Schematics

The photodetector is designed to be implemented in IBM 45nm SOI process. A schematics is shown in Fig. 3.1. As shown in the schematics, a single-ring-single-bus
structure is used. Crystal silicon is used as waveguides. It has thickness of 80 nm and width of 600 nm. A wide poly-silicon cap is introduced within in ring region and they form ridge waveguides. The thickness of poly-Si is 65 nm. It can be observed from Fig. 3-1 that the mode is well confined.

The center part of poly-silicon is the intrinsic region of a p-i-n diode. The outer and inner rims of the poly-silicon are P+ and N+ doped respectively. Since the heavy doped region is very lossy due to free carrier absorption, the intrinsic region should be wide enough to keep optical mode away from the doped region. However, the intrinsic region should not be too wide, since it will potentially limit the bandwidth of the photodetector due to long transit time. Optical mode simulation is done by commercial software FIMMWAVE. The intrinsic region is chosen to be 1.6 $\mu m$ in this device. The mode profile of a cross-section inside the ring is shown in Fig. 3-1. Metal contacts are built on top of the doped region to allow electrical access to the N+ and P+ regions.

A tapering of poly-silicon cap is designed for smooth coupling and reduce re-
flection and scattering loss. FDTD (Finite Difference Time Domain) simulation by commercial software OmniSim shows the insertion loss of one taper is around 0.45dB.

The radius of the ring is set at 12 $\mu m$, so that the bending loss is small. FDTD simulation shows the round trip through power is 0.9645, which correspond to 20.8 dB/cm. This number is close to the lower bound of anticipated waveguide loss.

Additionally, the optical input and output are through vertical grating couplers.

### 3.1.2 Gap versus Coupling

The coupling between ring and bus waveguides needs to match the round trip loss in the ring, so that critical coupling can be achieved. The coupling coefficients are obtained by FDTD simulation through commercial software OmniSim. The refractive indices for poly-silicon, crystal silicon and silicon dioxide are 4.53, 4.48 and 1.44.

According to the coupled mode theory, the coupling coefficients are proportional to the mode overlap. For two adjacent rectangular waveguides, the cladding modes decay exponentially with the distance away from waveguide core. Therefore, if we assume the distance between two adjacent waveguides are $d$, the two overlap modes can be normalized to $e^{-x}$ and $e^{x-d}$. Therefore, the mode overlap can be calculate by

$$\frac{1}{d} \int_{0}^{d} e^{-x}e^{x-d}dx = e^{-d}.$$  

Therefore, the coupling strength decreases exponentially with the distance.

The simulated relationship between coupling versus gap size is shown in Fig. 3-2. It can be observed that the log-scaled coupled power drops linearly with as gap size increases. This is consistent with the theoretical prediction. The coupling coefficients of intermediate gap sizes can be interpolated from the simulated curve.

Previous work in our group[87] suggests that there is a large uncertainty of the waveguide loss. For a waveguide with narrow poly-silicon ridge sitting on wide crystal silicon slab, measured loss ranges from 1 dB/cm to 60dB/cm for 1550 nm and 50 dB/cm to 130 dB/cm for 1220 nm. Additionally, low waveguide loss of 3dB/cm is observed before for crystal silicon waveguide. Surface roughness of poly-silicon could be one of the possible reasons for high loss.

In this thesis, the range of waveguide loss is considered between 20 to 150 dB/cm.
Figure 3-2: Power fractional coupling versus gap sizes

Figure 3-3: Simulated maximum extinction versus waveguide loss for different gaps between ring and bus waveguides
As a result, multiple devices with different gap sizes were needed, so that at least one device can reach critical coupling. The gap sizes are chosen in a way that at least one device provides an extinction of 20 dB. The calculated gap sizes are 100, 130, 160, 190, 220, 260, 300 and 340 nm. Once the corresponding coupling strength are interpolated from curve in Fig. 3-2, the relationship between extinction ratio and waveguide loss in the ring can be calculated by Eqn. 1.18. The curve is presented in Fig. 3-3.

3.1.3 Layout

The design is drawn in Cadence Virtuoso. A screen shot is shown in Fig. 3-4. Three rectangular metal contact pads are used to access the P and N doped regions. They are arranged in GSG (Ground-Signal-Ground) pattern with 100 \( \mu m \) spacing to allow high frequency connection with a GSG probe from Cascade Microtech.
3.2 Characterisation

Photos of fabricated devices are shown in Fig. 3-5. Subfigures (a), (c) and (e) show the front side of a whole reticle, a block of devices and a single photodetector. Subfigures (b), (d), (f) show the back side version. Due to the top metal layers of SOI process, the underlying dielectric structures are not visible from front side. On the back side, the silicon substrate is removed and whole circuit is transferred onto a transparent handle layer. Therefore, the dielectric structures can be observed from the back side.

3.2.1 Dark Current

The dark current of a fabricated photodetector is shown in Fig. 3-6(a). It can be observed that the dark current rises almost exponentially with voltage on reverse bias. On forward bias, it seems the dark current increases even faster than exponential increment. Additionally, no obvious turn-on can be observed. This means the photodetector behaves more like a conductor than a diode.

Another measurement was performed to determine the relationship between dark current and temperature. The result is shown in Fig. 3-6(b). It can be observed that the dark current increases with temperature. Over the range from 25 to 50 degree Celsius, the dark current increases more than two fold.

There are a few theories that explain the I-V characteristics in poly-silicon p-i-n junction. The fact that the dark current changes significantly with temperature suggests it might be governed by thermionic emission. Other literature considered diffusion and recombination in intrinsic region the dominating factors. Numerical simulation models can be built based on solution of electron and hole continuity equation. However, this theory should give a near linear relationship in semilog I-V curve, which is not consistent with experimental results.

A recent review of tapeout files suggests the intended intrinsic region might be doped by mistake. This effectively turns the p-i-n diode into a p-n diode. If the doping degenerates, it forms an Esaki diode, while tunnelling plays an important role. As shown in Fig. 3-7(a) with high doping concentration and narrow depletion
Figure 3-5: Photos of fabricated photodetectors
region, tunnelling current dominates at low bias. At forward bias, as the bias voltage increases, the energy of free electrons in N region is raised to the bandgap of P region, which decreases tunnelling current. As forward bias increases even further, electrons start to be able to overcome the potential barrier. Therefore, current increases again, in the same way as traditional p-n junctions. As shown in Fig. 3-7(b), typical Esaki diodes have a negative differential resistance at small forward bias.

Figure 3-6: Dark current

For the photodiodes in this chapter, due to large density of defect states in poly-
silicon, the tunnelling might not be significant enough to display a negative differential resistance. However, it will explain the smaller differential resistance in lower forward bias voltage.

### 3.2.2 Passive Optical Properties

The resonance condition of the ring with different gap sizes between ring and bus waveguide are shown in Fig. 3-8. It can be observed that the photodetector with gap size of 160 nm has the highest extinction ratio of more than 25 dB. From Fig. 3-2 this gap size corresponds to a cross power coupling strength of 0.1997. Assuming a lossless coupling region, the through power coupling strength is 0.8003. Since the extinction ratio of this device is big, it is close to critical coupling. Under this assumption, the round trip through power is equal to through coupling strength. This will give a waveguide loss of 128 dB/cm. A lossy coupling region will result in a smaller through power coupling and hence a larger waveguide loss.

![Figure 3-8: Measured resonance of different photodetectors](image)

Another way to estimate the waveguide loss is by fitting simulated through port
Figure 3-9: Comparison of ring resonance between measurement and simulation power spectrum with experimental data. A fitted through port response from this value is compared with the measured response and shown in Fig. 3-9. The waveguide loss associated with this curve is 130 dB/cm, which is very close to the value calculated above.

Additionally, it can be observed that for photodetectors with gap sizes of 130 and 190 nm, the extinction ratio are still more than 10 dB, which is not bad for photodetectors. This indicates that the performance of the photodetectors are not extremely sensitive to the gap sizes.

The relationship between resonance and injected light intensity is shown in Fig. 3-10. The power shown here is the tap power, which is 1/9 of the injected light power. It can be observed that with increasing light power, the resonance shifts to longer wavelength. Additionally, at higher light intensity, the resonance peak becomes asymmetric. This is due to the heating at higher light intensity, which increases the refractive index of waveguide and hence the resonance wavelength. The tuning sensitivity is roughly 115 GHz/mW.
3.2.3 DC Photocurrent

The photocurrent of the photodetectors were measured. Photocurrent under -25V and +15V bias are shown in Fig. 3-11(a). A slight rolling off with increasing light intensity can be observed. As a result, the quantum efficiency is expected to decrease as well. The quantum efficiency at -25V bias is shown in the same figure. The highest value is around 3.5% and it drops to lower than 2% at higher input light power.

Additionally, the photocurrent is tested under different temperature. As shown in Fig. 3-11(b) photocurrent is measured under 25, 35 and 45 degree Celsius, while the tap light power is held at 0.1 mW. The three curves roughly coincide with each other. This indicates that the photocurrent of this detector is not sensitive to temperature.

3.2.4 Bandwidth

The bandwidth of the photodiode is measured with different bias voltages. The result at 25 degree Celsius is shown in Fig. 3-12(a). The 3 dB bandwidth is roughly 1.25 GHz.
Fig. 3-12(b) shows a measurement of bandwidth under 45 degree Celsius. Compared with the results in Fig. 3-12(a), it can be observed that the bandwidth decreases slightly compared with the case under 25 degree Celsius.

Fig. 3-12(c) shows a measurement of bandwidth with different light intensity. It can be observed that the bandwidth for a high light intensity is slightly smaller.

The 3dB bandwidth under RC and transit limit can be calculated by Eqn. 3.1. The transit time $\tau_{tr}$ is the time needed for carriers to drift across the space charge region. The speed of carriers are described by Eqn. 3.2. Here $E$ is the electric field, $\mu$ is the mobility of a certain carrier species, $L_{sp}$ is the length of space charge region. Carrier mobility in poly-silicon is much smaller than crystal silicon. The actual values are undetermined. From Eqn. 3.2 it can be observed that transit time is directly proportional to the bias voltage. Since the three measurements under different bias voltages give similar results. This indicates the photodetector is not transit limited.

$$f_{3dB} = \frac{1}{2\pi \sqrt{(RC)^2 + (\tau_{tr}/2.636)^2}}$$ (3.1)
Figure 3-12: Bandwidth analysis

(a) Bandwidth versus bias voltage at 25 degree Celsius

(b) Bandwidth versus bias voltage at 45 degree Celsius

(c) Bandwidth versus input light intensity
\[ v = \mu E \]  \hspace{1cm} (3.2)

\[ \tau_{tr} = \frac{L_{sp}}{v} = \frac{L_{sp}}{\mu E} = \frac{L_{sp}^2}{\mu V_{bias}} \]  \hspace{1cm} (3.3)

The RC limited bandwidth can be estimated with the knowledge of resistance and capacity of this photodiode, which is not available at this point.

### 3.3 Conclusion

In this chapter, the effort of design and characterization of a tele-comm wavelength photodiode in CMOS platform is presented. Defect states in poly-silicon is used to absorb light in sub-bandgap wavelength. A ring resonator structure is introduced to increase absorption and reduce device footprint. The devices are fabricated in IBM 45nm SOI process. The characterisation shows dark current less than 100 nA, quantum efficiency of 3.5% and bandwidth of 1.25 GHz can be achieved at 25 V reverse bias.
Chapter 4

Integrated QPSK Modulator

Inspired by a new idea of building phase modulator with double coupled ring resonator, an integrated QPSK receiver on CMOS platform is proposed. In this chapter, the derivation for wavelength response is firstly presented. The problems in the original paper are corrected. Afterwards, a way of implementing the modulators in zero-change CMOS process is proposed.

4.1 Derivation

Directional couplers are fundamental components of integrated optic circuits. Their behavior can be described by scattering matrices. A schematic of an arbitrary 4-port directional coupler is shown in Fig. 4-1. Under the assumption that there are no self-reflection or cross-reflection from the same side, the scattering matrix can be written in the form of Eqn. 4.1. The power conservation and time reversal properties of scattering matrix dictate that $S^\dagger = S^{-1}$ and $S^\dagger = S^*$. The combination of these two conditions implies that the scattering matrix is symmetric. Subsequently, the time reversal condition can be written as in Eqn. 4.2.
Figure 4-1: Schematics of an arbitrary directional coupler

\[
S = \begin{bmatrix}
0 & 0 & t_1 & \kappa_1 \\
0 & 0 & \kappa_2 & t_2 \\
t_3 & \kappa_3 & 0 & 0 \\
\kappa_4 & t_4 & 0 & 0 \\
\end{bmatrix} = \begin{bmatrix}
0 & 0 & t_1 & \kappa_1 \\
0 & 0 & \kappa_2 & t_2 \\
t_1 & \kappa_2 & 0 & 0 \\
\kappa_1 & t_2 & 0 & 0 \\
\end{bmatrix}
\] (4.1)

\[
S^*S = \begin{bmatrix}
\kappa_1 t_1^* + t_1 \kappa_1^* & t_2 \kappa_1^* + \kappa_2 t_1^* & 0 & 0 \\
t_1 \kappa_2^* + \kappa_1 t_2^* & \kappa_2 \kappa_2^* + t_2 \kappa_2^* & 0 & 0 \\
0 & 0 & \kappa_2 \kappa_2^* + t_1 \kappa_1^* & \kappa_1 \kappa_1^* + t_2 \kappa_2^* \\
0 & 0 & t_1 \kappa_1^* + \kappa_2 t_2^* & \kappa_1 \kappa_1^* + t_2 \kappa_2^* \\
\end{bmatrix} = I
\] (4.2)

It can be observed that the first quadrant of the scattering matrix is the transfer matrix of the directional coupler and it can be defined as \( T \). By solving Eqn. 4.2, Eqn. 4.3 can be obtained. If a common phase \( \phi = \frac{\theta_{t1} + \theta_{t2}}{2} \) is extracted out, the transfer matrix can be written in form of Eqn. 4.4.

\[|t_1| = |t_2|\] (4.3)

\[|\kappa_1| = |\kappa_2|\]

\[|t_1|^2 + |\kappa_1|^2 = 1\]

\[\theta_{t1} + \theta_{t2} = \theta_{\kappa_1} + \theta_{\kappa_2} \pm \pi\]
\[ T = \begin{bmatrix} t_1 & \kappa_1 \\ \kappa_2 & t_2 \end{bmatrix} = e^{j\phi} \begin{bmatrix} \kappa_1 e^{j(\theta_{t1} - \theta_{t2})/2} & \kappa_1 e^{j(\theta_{a1} - \theta_{a2} \pm \pi)/2} \\ \kappa_1 e^{j(\theta_{a2} - \theta_{a1})/2} & \kappa_1 e^{j(\theta_{t2} - \theta_{t1})/2} \end{bmatrix} = e^{j\phi} \begin{bmatrix} t & \kappa \\ -\kappa^* & t^* \end{bmatrix} \] (4.4)

In integrated photonic circuits, the coupling region between a ring and an access waveguide is usually symmetric, since the coupling region looks identical from input and output sides. This requires the transfer matrix to be symmetric. Therefore, it can be further reduced to Eqn. 4.5. Here, \( \theta_{t1} \) and \( \theta_{t2} \) are called coupling-induced phase shift (CIPS).

\[ T = \begin{bmatrix} |t_1| e^{j\theta_{t1}} & i|\kappa_1| e^{j(\theta_{a1} + \theta_{a2})/2} \\ i|\kappa_1| e^{j(\theta_{a2} + \theta_{a1})/2} & |t_1| e^{j\theta_{t2}} \end{bmatrix} \] (4.5)

### 4.2 Double-coupled Resonator

![Schematics of a double-coupled resonator](image)

An double-coupled resonator is potentially useful as a phase modulator. A schematic is shown in Fig. 4-2. Both \( l_1 \) and \( l_3 \) are half circumference of the ring. Both of the coupling regions between ring and bus waveguides can be described by the transfer matrix in Eqn. 4.5. The through port and drop port responses are shown in Eqn. 4.6 and Eqn. 4.7. In these equations, \( \alpha \) and \( \beta \) are imaginary and real components of the...
These responses are plotted in Fig. 4-3(a) and Fig. 4-3(b). It can be observed that, when on resonance, the majority of the power goes to drop port and its relative phase to input port is close to \( \pm \pi \). When the ring is off resonance, majority of the power goes to the through port and its relative phase to input port is close to zero. Therefore, if the through and drop port are combined together with two waveguides of equal length, they can be used as a phase modulator.

\[
\frac{b_2}{a_2} = |t| e^{j\phi_{t2}} \frac{1 - e^{-(\alpha_1 + j\beta_1)l_1} e^{-(\alpha_3 + j\beta_3)l_3} e^{j2\phi_{t1}}}{1 - |t|^2 e^{-(\alpha_1 + j\beta_1)l_1} e^{-(\alpha_3 + j\beta_3)l_3} e^{j2\phi_{t1}}} \tag{4.6}
\]

\[
\frac{b_4}{a_2} = -|\kappa|^2 e^{j(\phi_{t2} + N\pi)} e^{-(\alpha_1 + j\beta_1)l_1} \frac{1 - |t|^2 e^{-(\alpha_1 + j\beta_1)l_1} e^{-(\alpha_3 + j\beta_3)l_3} e^{j2\phi_{t1}}}{1 - |t|^2 e^{-(\alpha_1 + j\beta_1)l_1} e^{-(\alpha_3 + j\beta_3)l_3} e^{j2\phi_{t1}}} \tag{4.7}
\]

Assume the propagation constants are the same throughout the whole structure and \( l_1 = l_3 \). When on resonance \( (2\phi_{t1} - \beta_1 l_1 - \beta_3 l_3 = 2N\pi, \) N is an integer), Eqn. 4.7 can be further reduced to Eqn. 4.8. When off resonance \( (2\phi_{t1} - \beta_1 l_1 - \beta_3 l_3 = (2N+1)\pi, \) N is an integer), Eqn. 4.6 can be further reduced to Eqn. 4.9. By comparing the two equations, it can be observed that N need to be an even number for the correct phase modulation. Therefore, the working wavelength is further restricted to \( 2\phi_{t1} - \beta_1 l_1 - \beta_3 l_3 = 4M\pi \), while M is an integer. Additionally, in order for the through port intensity response to be close to unity, the resonator needs to be strongly over-coupled. As shown in Fig. 4-3(a) a nominal intensity of 0.95 is obtained at through port while on resonance. In this simulation, the energy coupling coefficient is 0.1, while the round trip loss is 0.0026.

\[
\frac{b_4}{a_2} = -|\kappa|^2 e^{j(\phi_{t2} + N\pi)} e^{-(\alpha_1 + j\beta_1)l_1} \frac{1 - |t|^2 e^{-(\alpha_1 + j\beta_1)l_1} e^{-(\alpha_3 + j\beta_3)l_3} e^{j2\phi_{t1}}}{1 - |t|^2 e^{-(\alpha_1 + j\beta_1)l_1} e^{-(\alpha_3 + j\beta_3)l_3} e^{j2\phi_{t1}}} \tag{4.8}
\]
Figure 4-3: (a) Intensity response of double-coupled resonator (b) Phase response of double-coupled resonator
4.3 Double-coupled Resonator with Feed Through Waveguide

Directly combing the through and drop ports of a double-coupled resonator results in an inevitable insertion loss of 3dB. To avoid this loss, a paper in 2012 [14] proposed to feed the through port to the drop port with a waveguide. A schematic is shown in Fig. 4-4.

To simplify the derivation, the shaded region is treated as a directional coupler. After deriving its parameters, the total response of the system can be derived as if it was a 1-ring-1-bus resonator. The two coupling regions are identical. The effective transfer matrix of the shaded area is described in Eqn. 4.10 and Eqn. 4.11. Here $T$ is the transfer matrix of a simple directional coupler.

\[
\frac{b_2}{a_2} = |t|e^{j \phi t} \frac{1 + e^{-(\alpha_1 l_1 + \alpha_3 l_3)}}{1 + |t|^2 e^{-(\alpha_1 l_1 + \alpha_3 l_3)}} \quad (4.9)
\]
\[ T_{\text{eff}} = T \begin{bmatrix} e^{-(\alpha_1 + j\beta_1)l_1} & 0 \\ 0 & e^{-(\alpha_2 + j\beta_2)l_2} \end{bmatrix} T = \begin{bmatrix} t_1 & \kappa_1 \\ \kappa_2 & t_2 \end{bmatrix} \] (4.10)

Where:

\[ t_1 = |t|^2 e^{j2\phi_{t1}} e^{-(\alpha_1 + j\beta_1)l_1} - |\kappa|^2 e^{j(\phi_{t1} + \phi_{t2})} e^{-(\alpha_2 + j\beta_2)l_2} \] (4.11)

\[ \kappa_1 = \kappa_2 = j|t||\kappa|e^{j(3\phi_{t1} + \phi_{t2})/2} e^{-(\alpha_1 + j\beta_1)l_1} + j|t||\kappa|e^{j(\phi_{t1} + 3\phi_{t2})/2} e^{-(\alpha_2 + j\beta_2)l_2} \]

\[ t_2 = -|\kappa|^2 e^{j(\phi_{t1} + \phi_{t2})} e^{-(\alpha_1 + j\beta_1)l_1} + |t|^2 e^{j2\phi_{t2}} e^{-(\alpha_2 + j\beta_2)l_2} \]

Afterwards, the response for the whole system can be derived. It is shown in Eqn. 4.12.

\[ \frac{b_4}{a_2} = \frac{t_2 + (\kappa_1 \kappa_2 - t_1 t_2)e^{-(\alpha_3 + j\beta_3)l_3}}{1 - t_1 e^{-(\alpha_3 + j\beta_3)l_3}} \] (4.12)

As demonstrated in previous section, a \( \pi \) phase shift exists between drop and through port response of a double-coupled resonator. Therefore, intuitively, if the index of the ring is modulated, so that the ring is alternating between on and off resonance, phase modulation can be achieved.

To simplify the analysis, responses when on and off resonance are studied. When on resonance \( (2\phi_{t1} - \beta_1 l_1 - \beta_3 l_3 = 4M\pi, M \text{ is an integer}) \), its response is reduced to Eqn. 4.13 When off resonance \( (2\phi_{t1} - \beta_1 l_1 - \beta_3 l_3 = (4M + 1)\pi, M \text{ is an integer}) \), its response is reduced to Eqn. 4.14.

\[ \frac{b_4}{a_2} = -e^{j2\phi_{t2}} e^{-(\alpha_2 + j\beta_2)l_2} e^{-(\alpha_1 l_1 + \alpha_3 l_3)} - |t|^2 + |\kappa|^2 e^{-\alpha_1 l_1} e^{-j\phi_{t1}} e^{(\alpha_2 + j\beta_2)l_2} \] (4.13)
\[
\frac{b_4}{a_2} = e^{j\phi_{l2}} - j|\kappa|^2 e^{-\alpha_{l1}} + |t|^2 e^{j\phi_{l2}} e^{-(\alpha_{l2} + j\beta_{l2})l_2} + e^{j\phi_{l2}} e^{-(\alpha_{l1} + \alpha_{l3})} e^{-(\alpha_{l2} + j\beta_{l2})l_2} \frac{1 + |t|^2 e^{-(\alpha_{l1} + \alpha_{l3})} + j|\kappa|^2 e^{-\alpha_{l3}l_2} e^{-(\alpha_{l2} + j\beta_{l2})l_2}}{1 + |t|^2 e^{-(\alpha_{l1} + \alpha_{l3})} + j|\kappa|^2 e^{-\alpha_{l3}l_2} e^{-(\alpha_{l2} + j\beta_{l2})l_2}}
\]

(4.14)

Since phase shift from the feed through waveguide is not desired, \(\phi_{l2} - \beta_{l2} = 2P\pi\), where \(P\) is an integer. The two equations above can then be further simplified as shown in Eqn. 4.15 and Eqn. 4.16. Two approximation are used here. First of all, the resonator is strongly over coupled and round trip loss is very small. Secondly, \(|\kappa|^2\) is much smaller than \(|t|^2\). It can be observed that a \(\pi\) shift exists, while intensity is roughly unity. This is consistent with the intuition that tuning the ring could result in phase modulation.

\[
\frac{b_4}{a_2} \approx -e^{2j\phi_{l2}} e^{-(\alpha_{l1} + \alpha_{l3})} - |t|^2 + |\kappa|^2 e^{-\alpha_{l1}} e^{\alpha_{l2}l_2} \frac{1 - |t|^2 e^{-(\alpha_{l1} + \alpha_{l3})} + |\kappa|^2 e^{-\alpha_{l3}l_2} e^{\alpha_{l2}l_2}}{1 - |t|^2 e^{-(\alpha_{l1} + \alpha_{l3})} + |\kappa|^2 e^{-\alpha_{l3}l_2} e^{\alpha_{l2}l_2}}
\]

(4.15)

\[
\frac{b_4}{a_2} \approx e^{j\phi_{l2}} e^{-\alpha_{l2}l_2}
\]

Simulation was done for carrier injection modulation. The relationship between carrier concentration, refractive index and free carrier absorption are extracted from Soref’s paper[16]. The figures are shown in Fig. 4-5(a) and Fig. 4-5(b) respectively. It is assumed free electrons and holes have the same concentration.

In this simulation, the ring radius is set as 12 \(\mu m\) and the length of section \(l_2\) is set equal to circumference of the ring. The ring is on resonance at 1545.39 nm, with a corresponding propagation constant of 7.1668 \(\mu m^{-1}\). The coupling strength is set as 5%. Assume the ring is on resonance when no carrier is injected. The
Figure 4-5: (a) Carrier refraction in c-Si at $\lambda=1.55\mu m$ as a function of free-carrier concentration. (b) Absorption in c-Si at $\lambda=1.55\mu m$ as a function of free-electron concentration. (c) Absorption in c-Si at $\lambda=1.55\mu m$ as a function of free-hole concentration.[16]
intensity and phase change with carrier injection into the ring are shown in Fig. 4-6(a) and Fig. 4-6(b) respectively. The overall insertion loss is about 0.5dB. However, at injection density of $5 \times 10^{17}$ cm$^{-3}$, there is a power dip of about 3.5 dB. At low injection concentration, the ring is on resonance and most of light travels through the ring. Free carrier absorption at the point is relatively small and hence the insertion loss is small. At high injection concentration, the ring is shifted off resonance and most of light travels through the feed through waveguide. Since there is no carrier injection here, the insertion loss is small as well. However, at intermediate carrier concentration, significant amount of light travels through the ring and suffered from free carrier absorption. This explains the power dip. From Fig. 4-6(b) it can be observed that a phase modulation of $\pi$ is achieved at injection density of $5 \times 10^{18}$ cm$^{-3}$.

This phase modulation is difficult to realize by carrier extraction. Doing so requires the ring to be on resonance at large carrier concentration. This will force the light to travels through highly lossy ring, which is not desirable.

Electric contacts need to be introduced for biasing. However, in IBM SOI process, there is not partial etch in silicon layer. Therefore, the electric contacts need to land directly on body silicon waveguides without disturbing the mode. Previous work from POEM project[54] uses a multimode ring instead of a single mode ring. As shown in Fig. 4-7, the ring is much wider than the single mode waveguide to allow multimode operation. It is segmented into alternating P and N doped region. The metal contacts lie at inner side of the ring. The coupling region is carefully designed to excite only the fundamental model. Therefore, the mode does not overlap with the metal contacts. Additionally, an inner doped ring is used for thermal tuning.

There are a few details need to be settled before applying this scheme. First of all, the ring size and access waveguide geometry need to be modified to fit targeted wavelengths, such as 1550nm.

Since it is difficult to measure phase of an optical signal directly, a test structure is needed to verify the phase modulation. A possible design is illustrated in Fig. 4-8. A Mach-Zehnder interferometer is formed by phase modulator in one arm and a reference waveguide in the other arm. If the phase modulator works properly, an
Figure 4-6: (a) Intensity response with carrier injection. (b) Phase response with carrier injection.
4.4 QPSK Modulator

A QPSK modulator can be created by combining two phase modulator through a 90 degree phase shifter. A specially designed directional coupler can be a robust 90 degree phase shifter. Based on the transfer matrix shown in Eqn. 4.5 if another symmetry is applied to the two pairs of input and output, the coupling induced phase
shift from two branches need to be the same. The transfer matrix is reduced to Eqn. 4.17. It can be observed that two inputs will be mixed with a 90 degree shift. If the transmission and coupling strength is set to be the same, this can be used as a 90 degree phase shifter. However, 3dB insertion loss is introduced in this process.

\[
T = e^{j\theta_1} \begin{bmatrix} |t_1| & i|\kappa_1| \\ i|\kappa_1| & |t_1| \end{bmatrix}
\] (4.17)

A schematic of such a QPSK modulator is shown in Fig. 4-9. A CW laser source is fed into the modulator and divided into two identical branches. They are phase modulated by I and Q channel signals respectively. Afterwards, the two branches are mixed by the 90 degree phase shift directional coupler. QPSK signal can be obtained at the outputs of this directional coupler.

![Figure 4-9: Schematics of a QPSK modulator](image)

**4.5 Conclusion**

In this chapter, a design of integrated phase modulator and QPSK modulator is discussed. First of all, rigorous derivation for wavelength response of double coupled resonator is presented. Based on the derivation, a $\pi$ phase shift exists between through and drop ports. Therefore, a phase modulator can be built by feeding through port into drop port. Simulation shows low insertion loss phase modulation can be achieved by carrier injection into the ring. Finally, a QPSK modulator is proposed by combining two phase modulators with a symmetric directional coupler, which introduce 90 degree phase shift.
Chapter 5

Conclusion and Future Work

This chapter will first summarise the research effort towards integrated QPSK optical link on CMOS platform. The development focus on 90 degree hybrids, 1550 nm poly-silicon photodetectors and low-energy phase modulators. Design, fabrication and characterization of these devices will be presented. Afterwards, the schematic of proposed QPSK receiver will be presented, followed by an updated schematic with new devices. In addition to the summary, possible future improvement for these devices will be mentioned in later sections.

5.1 Conclusion

5.1.1 90 Degree Hybrid

Efforts of developing 90 degree hybrids were presented in Chapter 2. The focus is MMIs, which rely on mode beating and self-imaging to produce four-fold image of the inputs. To quantify the performance of MMIs, FOMs (figures of merit) are defined to describe the intensity and phase balance respectively. Simulation are performed to reveal the relationship between MMI geometry and FOMs.

Some of the best devices are fabricated on Micron D1S, D1L and IBM EOS18 processes. Measurement results from D1S demonstrated power imbalance within 1.2 dB and phase imbalance within 10 degrees. Additionally, good match are observed.
between experimental results and numerical simulation. Simulation also shows a thickness variation of 10 nm on Micron process result in intensity error less than 1 dB and phase error less than 10 degree. Future improvement is required to meet Ciena’s specification\cite{86} (0.1 dB for intensity error and 5 degree for phase error).

For zero-change IBM process, since the body silicon layer thickness is only 80 nm, acceptable thickness variation is around 3 nm. TEM data on recent three IBM chips shows thickness error of 3.2, 1.7 and 1.3 nm respectively. Therefore, it is possible to use MMI as 90 degree hybrid in zero-change CMOS process, along with all the high performance electronics.

### 5.1.2 Poly-silicon Photodetectors

The development of a poly-silicon ring resonant photodetector was presented in Chapter 3 of this thesis. To the author’s best knowledge, this is the first telecommunication wavelength photodetector in zero-change CMOS process.

Traditional absorptive material for telecommunication wavelengths, such as III-V alloys, are not available in CMOS process. The main dielectric material, crystal silicon, is transparent in telecommunication wavelength, since its bandgap is larger than the photon energy. However, poly-silicon can be used as absorptive material, thanks to its large density of sub-bandgap defect states. To further increase absorption and reduce device size, a ring resonator might be implemented.

Poly-silicon ring resonant photodetectors are designed to be implemented in IBM EOS18 45nm process. A reverse ridge waveguide with narrow crystal silicon ridge and poly-silicon cap is used. The inner and outer rims of poly-silicon ring are P+ and N+ doped respectively to form p-i-n diode.

Experiments show the dark I-V characteristic of these detectors is closer to a photo-resistor than diode. The dark current increases exponentially with bias voltage. Since the supposedly intrinsic region might be doped unintentionally, carrier tunnelling could dominate the dark current. Under illumination of 1550 nm, quantum efficiency of 3.5% and dark current of 60 nA are observed at 25V reverse bias. Additionally, the bandwidth of these detectors is examined. The 3 dB frequency
is close to 1.2 GHz. The roll-off beyond this point is slower than the typical 20
dB/decade. Experiments under different bias voltages show that the bandwidth is
not transit limited.

5.1.3 Integrated QPSK Modulator

A novel integrated QPSK modulator was proposed in Chapter 4 of thesis. The design
of an individual phase modulator is based on double coupled ring resonators. A \( \pi \)
phase shift exists between through port when off resonance and drop port when on
resonance. To combine these two outputs, a waveguide is used to feed through port
into drop port. The resonance condition can be tuned by carrier injection into the
ring. Simulation shows with a heavily over-coupled ring, phase modulation can be
achieved with insertion loss less than 1 dB. A QPSK modulator can be constructed
by connecting two phase modulators by a symmetric directional coupler, which in-
troduces 90 degree phase shift.

Compared with traditional MZI (Mach-Zehnder interferometer) based modulators,
ring modulators are much more compact and hence potentially very energy efficient.
Previous work shows an energy consumption as low as 40 fJ/bit [54] for ring modula-
tors, while MZI based modulators typically consume a few pJ/bit. Additionally, the
parameters of the modulators are designed to fit IBM’s 45 nm CMOS process. This
enables integration of a complete QPSK modulator into a single CMOS chip.

5.1.4 Summary

In the first chapter of this thesis, a monolithically integrated QPSK receiver was
proposed. The schematic is shown in Fig. 5-1(a). As shown in the figure, 90 degree
hybrids and photodetectors need to be developed in CMOS platform. Prototypes
are designed, fabricated and demonstrated. A new version of the QPSK receiver
schematic is shown in Fig. 5-1(b). The hybrids and phododetectors are replaced with
photos of fabricated devices. Together with existing ADC and DSP electronics, it is
now possible to fabricate a complete single polarization QPSK receiver. If integrated
PBS is added, DP-QPSK receivers can be built.

Additionally, a possible way for QPSK modulator in CMOS platform emerged during the course of this project and motivated Chapter 5. If the proposed modulator can be implemented successfully in the future, a monolithically integrated QPSK transmitter may be developed. Along with expected improvement of photodetectors in quantum efficiency and bandwidth, it is possible to build an energy efficiency transceiver with a single CMOS chip.

5.2 Future Work

5.2.1 90 Degree Hybrid

MMI theory relies on approximations such as the device has very high index contrast so that it is close to waveguide with metallic boundaries. There is always a discrepancy between this assumption and reality. Additionally, it assumes the wave vectors in the transverse direction are very small compared with propagation constants, which is increasingly inaccurate with higher order modes. Due to these factors, it is fundamentally impossible to fabricate a perfect MMI as a 90 degree hybrid. Additionally, it suggests that the optimal MMI does not necessary have the theoretical best geometry. Numerical simulation is useful in searching for the best MMI structures.

In D1S devices, only length variation was included for MMI. Further tuning of MMI geometry was implemented in D1L. However, as stated in Chapter 2, no useful data was generated from D1L devices due the faulty vertical grating couplers. It is useful to fix the coupler problem in future runs and test the modified MMI.

Additionally, the cascaded directional couplers based 90 degree hybrid implemented in EOS16 is faulty. One of the waveguide was not connected properly. This problem was fixed in EOS19 tape out. The new device need to be characterized in the future.

Other than searching for the perfect 90 degree hybrid, it is useful to understand the impact of its imperfection on a QPSK optical link. However, this requires an
Figure 5-1: (a) Proposed monolithically-integrated QPSK receiver (b) QPSK receiver with demonstrated devices
in-depth understanding of the data acquisition and error compensation algorithm in QPSK communication, which is not available to the author at this point. Discussion with engineers from Ciena suggests that phase imbalance may play a bigger role in undermining the bit error rate (BER) than intensity imbalance. Besides experience from professionals, faithful link level simulation software is useful to gain some insight. The author briefly explores the commercial software OptiSystem from Optiwave. However, this software is geared toward IM-DD WDM system. It does not include a reliable component for QPSK data acquisition. Therefore, no significant result has been obtained so far. Future investigation in this direction will be useful guidelines for designing 90 degree hybrids.

5.2.2 Poly-silicon Photodetectors

Multiple approaches may be used to modify the poly-silicon photodetectors. First of all, it turns out the bending loss inside the ring is quite high. It is close to the lower bound of our estimation of material absorption. A bigger ring will solve this problem.

In the initial design, the intrinsic region width is set to be 1.6 um. The purpose of such a wide intrinsic region is to minimize mode overlap with doped poly-silicon, which introduce large free carrier absorption. However, experiment shows the loss inside the ring is around 130 dB/cm. In this case, small free carrier absorption is acceptable. Therefore, in future fabrication, it is useful to shrink the intrinsic region. As a result, it will take less external bias to achieve the same electrical field intensity within the device. Additionally, it is believed that the electron mobility in poly-silicon is less than 20 cm²V⁻¹s⁻¹, which corresponds to a transit time around 50 ps under -25 V bias. This is comparable with the recombination time. Therefore, a smaller intrinsic region reduces the transit time and hence helps with quantum efficiency and bandwidth.

Additionally, the mask error which may cause the intrinsic region to be doped should be corrected. This will reduce light loss due to free career absorption and hence increase quantum efficiency. This will also widen the depletion region and result in a more diode-like I-V characteristics. Therefore, the dark current is expected to drop.
<table>
<thead>
<tr>
<th>Radius ($\mu$m)</th>
<th>Gap (nm)</th>
<th>Intrinsic region width ($\mu$m)</th>
</tr>
</thead>
<tbody>
<tr>
<td>12</td>
<td>160</td>
<td>1.1</td>
</tr>
<tr>
<td>12</td>
<td>160</td>
<td>1.3</td>
</tr>
<tr>
<td>12</td>
<td>160</td>
<td>1.05</td>
</tr>
<tr>
<td>24</td>
<td>130</td>
<td>1.3</td>
</tr>
<tr>
<td>24</td>
<td>130</td>
<td>1.6</td>
</tr>
</tbody>
</table>

Table 5.1: Parameters for new photodetector designs

New designs were added to later tape outs. The parameters modified include ring radius, gap sizes of coupling regions and intrinsic region width. The variations are shown in Table. 5.1

### 5.2.3 Integrated QPSK Modulator

Some preliminary structures have been implemented in IBM 45nm and 32nm SOI process. A fully passive structure is shown in Fig. 5-2(c). It is a MZI (Mach-Zehnder interferometer) with two 50/50 directional couplers in the ends. Two arms of the MZI have equal length. One of the arms contains a ring coupled to a U-shape waveguide, which resembles the phase modulator presented in Chapter 4. Additionally, as shown in Fig. 5-2(a), a reference structure without ring is implemented as well. It is meant to test whether the MZI works correctly.

Since the two directional couplers are identical and placed anti-symmetrically, when a laser is launched into one of the inputs of the MZI, only the anti-symmetric output will see significant output. Therefore, when one of the arms is phase modulated, the output intensity changes. In the case of all passive structure, a wavelength sweep will effectively tune the resonance condition of the ring. When on resonance, an intensity drop should be observed from the output.

Experiments from the reference structure shows the MZI works as anticipated in the previous paragraph. The response from its bright output is shown in Fig. 5-3(a). However, no intensity drop is observed in structure with ring. Instead, similar intensity are observed from both outputs at all wavelength. The data from one of the outputs is shown in Fig. 5-3(b). This indicates one of the arms in MZI might be broken. Future investigation is needed to understand the problem. Additionally, full
Figure 5-2: (a) Layout for reference structure (b) Fabricated reference structure (c) Layout for modulator test structure (d) Fabricated modulator test structure
Figure 5-3: (a) Wavelength response of the bright branch of reference MZI (b) Wavelength response from one of the outputs of modulated MZI phase modulators may be included in future tape out.

5.2.4 Conclusion

This section summarizes possible future modification on hybrids, poly-silicon photodetectors and QPSK modulators. Some existing MMI designs need to be re-fabricated with correct vertical couplers. For photodetectors, intrinsic width and doping adjustment is expected to improve quantum efficiency and bandwidth. Additionally, the designed QPSK modulators need to be fabricated. With the implementation and modification of these devices, it is possible to develop a complete QPSK transceiver on a single CMOS chip. Unlike traditional transceivers, III-V chips are not necessary.
Appendix A

Coupling Mode Theory

The behavior of a single directional coupler can be described with coupling mode theory. As shown in Fig. A-1, two waveguides are parallel to each other and their fundamental modes are weakly coupled to each other. The total field can be expressed by superposition of two individual modal fields as shown in Eqn. [A.1], where $U$ and $\beta$ are unperturbed solution for individual modes.

$$ E(x,y,z) = E_1(z)U_1(x,y)e^{-j\beta_1 z} + E_2(z)U_2(x,y)e^{-j\beta_2 z} \quad (A.1) $$

Under week coupling assumption, index of one waveguide can be seen as perturbation to the other waveguide. Define $\Delta\epsilon_2 = \epsilon_2 - \epsilon_c$ as perturbation for $U_1$. Similarly, define $\Delta\epsilon_1 = \epsilon_1 - \epsilon_c$ as perturbation for $U_2$. Therefore, the total dielectric constant is
\( \epsilon(x, y) = \epsilon_c + \Delta \epsilon_1(x, y) + \Delta \epsilon_2(x, y). \)

By substitution of net field with dielectric perturbation into wave Eqn. A.2 and eliminate second order term and original unperturbed solutions, Eqn. A.3 can be obtained.

\[
\nabla^2 \mathbf{E} + \epsilon(x, y, z)k_0^2 \mathbf{E} = 0 \tag{A.2}
\]

\[
2j \beta_1 \mathbf{U}_1 \frac{dE_1}{dz} e^{-j \beta_1 z} + 2j \beta_2 \mathbf{U}_1 \frac{dE_2}{dz} e^{-j \beta_2 z}
= \Delta \epsilon_2(x, y)k_0^2 \mathbf{U}_1 e^{-j \beta_1 z} + \Delta \epsilon_1(x, y)k_0^2 \mathbf{U}_2 e^{-j \beta_2 z} \tag{A.3}
\]

Dot product \( \mathbf{U}_1^* \) and \( \mathbf{U}_2^* \) by Eqn. A.3 and integrate over the whole cross section, Eqn. A.4 and Eqn. A.5 can be obtained respectively.

\[
2j \beta_1 \frac{dE_1}{dz} |\mathbf{U}_1|^2 dA + 2j \beta_2 \frac{dE_2}{dz} |\mathbf{U}_2|^2 dA
= k_0^2 E_1 e^{-j \beta_1 z} \int \Delta \epsilon_2 |\mathbf{U}_1|^2 dA + k_0^2 E_2 e^{-j \beta_2 z} \int \Delta \epsilon_1 |\mathbf{U}_1|^2 dA \tag{A.4}
\]

\[
2j \beta_1 \frac{dE_1}{dz} |\mathbf{U}_1|^2 dA + 2j \beta_2 \frac{dE_2}{dz} |\mathbf{U}_2|^2 dA
= k_0^2 E_1 e^{-j \beta_1 z} \int \Delta \epsilon_2 |\mathbf{U}_2|^2 dA + k_0^2 E_2 e^{-j \beta_2 z} \int \Delta \epsilon_1 |\mathbf{U}_2|^2 dA \tag{A.5}
\]

These equations can be further simplified. Take Eqn. A.4 as example. On the left side, since the two modes are only weakly coupled to each other, \( |\mathbf{U}_1|^2 \) is much larger than \( \mathbf{U}_1^* \cdot \mathbf{U}_2 \). Therefore, the second term is much smaller than the first term and hence can be neglected. For the first term on the right side, the integration is localized around waveguide 2, since \( \Delta \epsilon_2 \) is only non-zero on waveguide 2. By the same weakly coupling argument, \( |\mathbf{U}_1|^2 \) is negligible around waveguide 2. Therefore, the first term on the right side can be neglected as well. The same argument applies on Eqn. A.5. The simplified equations are shown in Eqn. A.6 and Eqn. A.7. They
can be rearranged to Eqn. A.8 and Eqn. A.9 respectively.

\[
2j\beta_1 \frac{dE_1}{dz} e^{-j\beta_1 z} \int |U_1|^2 dA = k_0^2 E_2 e^{-j\beta_2 z} \int \Delta \epsilon_1 U_1^* \cdot U_2 dA \tag{A.6}
\]

\[
2j\beta_2 \frac{dE_2}{dz} e^{-j\beta_2 z} \int |U_2|^2 dA = k_0^2 E_1 e^{-j\beta_1 z} \int \Delta \epsilon_2 U_2^* \cdot U_1 dA \tag{A.7}
\]

\[
\frac{dE_1}{dz} = -j \frac{k_0^2}{2\beta_1} E_2 e^{-j(\beta_2 - \beta_1)z} \int \Delta \epsilon_1 U_1^* \cdot U_2 dA \int |U_1|^2 dA \tag{A.8}
\]

\[
\frac{dE_2}{dz} = -j \frac{k_0^2}{2\beta_2} E_1 e^{-j(\beta_1 - \beta_2)z} \int \Delta \epsilon_2 U_2^* \cdot U_1 dA \int |U_2|^2 dA \tag{A.9}
\]

Define normalized amplitude as shown in Eqn. A.10. \(\eta\) is referring to the impedance here. By substitution into Eqn. A.8 and A.9, Eqn. A.11 can be obtained,

\[
a_1(z) \sqrt{2\eta_1} = E_1(z) e^{-j\beta_1 z}
\]

\[
a_2(z) \sqrt{2\eta_2} = E_2(z) e^{-j\beta_2 z} \tag{A.10}
\]

\[
\frac{da_1}{dz} = -j \beta_1 a_1 - j \kappa_{12} a_2
\]

\[
\frac{da_2}{dz} = -j \beta_2 a_2 - j \kappa_{21} a_1 \tag{A.11}
\]

where

\[
\kappa_{12} = \frac{k_0^2}{2\beta_1} \frac{\int \Delta \epsilon_1 U_1^* \cdot U_2 dA}{\int |U_1|^2 dA}
\]

\[
\kappa_{21} = \frac{k_0^2}{2\beta_2} \frac{\int \Delta \epsilon_2 U_2^* \cdot U_1 dA}{\int |U_2|^2 dA} \tag{A.12}
\]

A general solution while the two waveguides are identical is shown in Eqn. A.13.

\[
a_1(z) = e^{-j\beta z}(A_1 e^{-j\kappa z} + A_2 e^{j\kappa z})
\]

\[
a_2(z) = e^{-j\beta z}(B_1 e^{-j\kappa z} + B_2 e^{j\kappa z}) \tag{A.13}
\]
A simple four-port directional coupler is shown in Fig. A-2, which is basically two parallel identical waveguides. Assuming a beam with normalized amplitude is injected into waveguide 1 and no input from waveguide 2, the two outputs can be calculated by applying boundary condition on Eqn. A.13. The results is shown in Eqn. A.14.

It can be observed that energy is oscillating between two waveguides. Additionally, since $|a_1(z)|^2 + |a_2(z)|^2 = |a_1(0)|^2$, total energy is conserved. At a specific length so that $\kappa z = \frac{\pi}{4}$, the optic power is equally split between two outputs. The propagation matrix for such a directional coupler is shown in Eqn. A.15.

\[
\begin{align*}
a_1(z) &= -a_1(0) \cos(\kappa z) e^{-j\beta z} \\
a_2(z) &= j a_1(0) \sin(\kappa z) e^{-j\beta z}
\end{align*}
\]  

(A.14)

\[
\begin{vmatrix}
a_1(L) \\
a_2(L)
\end{vmatrix} = \frac{\sqrt{2}}{2} \begin{vmatrix}
-1 & j \\
j & -1
\end{vmatrix} \begin{vmatrix}
a_1(0) \\
a_2(0)
\end{vmatrix}
\]  

(A.15)
Appendix B

Self-imaging

For a waveguide supporting a large number of modes, the input light will produce n-fold image of itself at designated length $L$, which is called self-imaging. In a high index contrast waveguide, for mode number $\nu$, the propagation constant can be approximately described by Eqn. B.1. Therefore, the difference of propagation constants between fundamental mode and each excited mode can be described. The beat length is defined in Eqn. B.2 and B.3.

$$\beta_\nu = k_0 n_r - \frac{(\nu + 1)^2 \pi \lambda_0}{4 n_r W_e^2}$$  \hspace{1cm} (B.1)

$$L_\pi = \frac{\pi}{\beta_0 - \beta_1}$$  \hspace{1cm} (B.2)

$$\beta_0 - \beta_\nu = \frac{\nu (\nu + 2) \pi}{3 L_\pi}$$  \hspace{1cm} (B.3)

The symmetric or anti-symmetric property of eigen modes of a multi-mode waveguide can be shown in Eqn. B.4 and B.5. Since any input field profile can be expressed by superposition of eigen modes, Eqn. B.6 can be obtained. By inspecting this equation, it can be seen that at $z = 3 L_\pi$, a mirrored image of the input field profile will be produced. At $z = \frac{3}{2} L_\pi$, the field can be represented in Eqn. B.7. It can be
observed that two images are produced, one replica and one mirror image.

\[ \nu(\nu + 2) = \begin{cases} 
\text{even for } \nu \text{ even} \\
\text{odd for } \nu \text{ odd} 
\end{cases} \quad (B.4) \]

\[ \phi_{\nu}(-y) = \begin{cases} 
\phi_{\nu}(y) & \text{for } \nu \text{ even} \\
-\phi_{\nu}(y) & \text{for } \nu \text{ odd} 
\end{cases} \quad (B.5) \]

\[ \Psi(y,0) = \sum_{\nu} c_{\nu} \psi_{\nu}(y) \]

\[ \Psi(y,z) = \sum_{\nu} c_{\nu} \psi_{\nu}(y) e^{j(\beta_{\nu} - \beta_{0})z} \]

\[ \Psi(y,z) = \sum_{\nu} c_{\nu} \psi_{\nu}(y) e^{j(\nu + 2)\pi / 3L_{\pi} z} \quad (B.6) \]

\[ \Psi(y,\frac{3}{2}L_{\pi}) = \sum_{\nu} c_{\nu} \psi_{\nu}(y) e^{j(\nu + 2)\pi / 2} \]

\[ \Psi(y,\frac{3}{2}L_{\pi}) = \sum_{\nu \text{ even}} c_{\nu} \psi_{\nu}(y) + \sum_{\nu \text{ odd}} -jc_{\nu} \psi_{\nu}(y) \]

\[ \Psi(y,\frac{3}{2}L_{\pi}) = \frac{1 - j}{2} \Psi(y,0) + \frac{1 + j}{2} \Psi(-y,0) \quad (B.7) \]

The general multi-fold image appears at intermediate length. The positions and phases of each image is analytically described in [12] using Fourier analysis. As shown in Fig. [B-1], the real MMI width is defined in region [0,W]. A virtue MMI width is added in the region [-W,0] and the anti-symmetric image of the input field is assumed in this region. Therefore, the total input field(real and virtual) can be represented by the superposition of anti-symmetric modes of the extended MMI region. At a minimal distance of \( \frac{3L_{\pi}}{N} \), \( N \) images with equal intensity will be formed. If they are numbered by \( q = 0, 1, \cdots, N - 1 \), the positions \( x_{q} \) and phases \( \phi_{q} \) can be described by Eqn. [B.8] As shown in Fig. [B-2] a total of \( 2N \) images will be produced in the
extended MMI region, when half of them are inverted. Therefore, in the real MMI region, there will be $N$ images.

$$x_q = (2q - N) \frac{W}{N}$$

$$\phi_q = q(N - q) \frac{\pi}{N}$$

(B.8)

When applying to a general $N \times N$ coupler, the positions of input and output in the case of $N$ is even are shown in Fig. B-3. The phase delay from input to output depends on both the position of input and output port. The relationship is show in
Eqn. B.9

\begin{align*}
\text{i + j even: } & \quad \phi_{ij} = \pi + \phi_{N-(j-i)/2} = \phi_0 + \pi + \frac{\pi}{4N} (j-i)(2N-j+i) \\
\text{i + j odd: } & \quad \phi_{ij} = \phi_{N-(j+i-1)/2} = \phi_0 + \frac{\pi}{4N} (j+i-1)(2N-j-i+1) \\
\end{align*}

A 4x4 coupler is shown in Fig. B-4. Two inputs are introduced into the MMI from port 4 and 2 respectively. The output delays are shown in Table B. It can be observed that the 4 phases is 90 degrees away from each other. Therefore, the 4x4 MMI is capable for serving as a 90 degree hybrid, which is essential in QPSK communication link.
\[
\begin{align*}
\phi_{41} &= \pi & \phi_{21} &= \frac{3}{4}\pi & \phi_{21} - \phi_{41} &= -\frac{1}{4}\pi \\
\phi_{42} &= -\frac{1}{4}\pi & \phi_{22} &= \pi & \phi_{22} - \phi_{42} &= \frac{5}{4}\pi \\
\phi_{43} &= \frac{3}{4}\pi & \phi_{23} &= \pi & \phi_{23} - \phi_{43} &= \frac{1}{4}\pi \\
\phi_{44} &= \pi & \phi_{24} &= \frac{7}{4}\pi & \phi_{24} - \phi_{44} &= \frac{3}{4}\pi
\end{align*}
\]

Since the performance of MMI depends on interference, it is intrinsically narrow band. Literature \cite{96} shows that the bandwidth is inversely proportional to the number of output ports and length of the device. Additionally, the fabrication tolerance is proportional to output channel separation.

A way to shorten the length of interferometric region length in MMI is to exploit restricted interference. Only a selection of modes are excited under this condition. The difference in propagation constants between \(\nu\)th mode and fundamental mode is proportional to \(\nu(\nu + 2)\) as shown in Eqn. \[B.3\] It can be calculated easily that for \(\nu\) excluding 2, 5, 8, etc, the value of \(\nu(\nu + 2)\) is multiple of 3. Therefore, if mode 2, 5, 8, etc is not excited or very weakly excited, the length where single image of the input field is reproduced is reduced 3 times. Subsequently, the N fold image will appear at length described by Eqn. \[B.10\] [11].

\[
L = \frac{1}{N} L_n
\]  \hspace{1cm} (B.10)

A possible approach it to place the excitation waveguide at \(\pm \frac{W_e}{6}\). As shown in Fig. \[B-5\], mode 2, 5, 8 hits zero with odd symmetry near these two positions. Therefore the excitation into these modes will be minimal. However, only two inputs can be used under this situation.

Another version of restricted interference is to excite only even modes by place the excitation at the center of the multimode waveguide. Since the value of \(\nu(\nu + 2)\) is all multiple of 4, the length needed for self-imaging can all be reduced by a factor of 4. However, this configuration will only be useful to realize \(1 \times N\) coupler.
Figure B-5: Normalized modes of multimode waveguide[11]
Appendix C

Derivation of Ring Resonator Response

An illustration of coupling between a bus waveguide and ring is presented in Fig. C-1. The relationship between two inputs and two outputs can be described by an unitary scattering matrix in Eqn. C.1. Additionally, the effect of travelling inside the ring can be described in Eqn. C.3. Therefore the transmission factor can be calculated and presented in Eqn. C.4. While resonance, where $\theta = m2\pi$, the transmission factor is reduced to Eqn. C.5. It can be observed that when $\alpha = |t|$, the transmitted energy is zero. This effect is called as critical coupling, where all input energy is going into the ring. All resonance photodetector should ideally work at critical coupling for
higher responsivity.

\[
\begin{aligned}
|b_1| &= \begin{vmatrix}
  t & \kappa \\
  \kappa^* & -t^*
\end{vmatrix} |a_1| \\
|b_2| &= \begin{vmatrix}
  \kappa^* & -t^*
\end{vmatrix} |a_2|
\end{aligned}
\]  \hfill (C.1)

\[|t|^2 + |\kappa|^2 = 1 \]  \hfill (C.2)

\[a_2 = b_2 \alpha e^{i\theta} \]  \hfill (C.3)

\[
\begin{aligned}
\frac{|b_1|^2}{a_1} &= \frac{\alpha^2 + |t|^2 - 2\alpha|t|\cos\theta}{1 + \alpha^2|t|^2 - 2\alpha|t|\cos\theta} \\
|b_1|^2 &= \frac{(\alpha - |t|)^2}{(1 - \alpha|t|)^2}
\end{aligned}
\]  \hfill (C.4)


[86] M. Frankel, “Ciena component specification (private communication),”


