Microwave Photonic Signal Processing
with Dynamic Reconfigurability

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ABSTRACT

Microwave photonics is an interdisciplinary subject that studies the interaction between microwave signals and photonic signals. It has wide applications in broad areas such as wireless networks, antennas, sensor networks and satellite communications. Microwave photonic signal processing is a powerful technique for processing high speed signals. It overcomes the electronic bottlenecks and possesses attractive advantages such as large instantaneous bandwidth, low loss, compact size and immunity to electromagnetic interference.

Microwave photonic signal processors with dynamic reconfigurability are of particular interest because they provide a route to robust, adaptable and lower power consumption systems. Most microwave photonic systems rely on highly stable lasers, which are usually thermally controlled and have external cavities. As the number of optical sources increases, the use of high performance lasers necessarily increases the cost, size and power consumption of the overall system.

An optical beamforming network that uses an uncooled Fabry-Perot laser is demonstrated. This is achieved by using a fast-scanning, high-resolution optical spectrum analyzer to track the frequency and power shift of the uncooled laser, and then reconfiguring a programmable Fourier-domain optical processor to provide compensation. In this way, the need for temperature control of the laser is eliminated, and the number of optical sources is reduced by using the output spectral lines of the laser. The system realizes six wideband microwave photonic phase shifters, and the resulting magnitude and phase responses vary within a $2\sigma$ deviation of 6.1dB and 14.8°, respectively, even when the laser current is changed during measurement.

A microwave photonic filter is presented based on a feedback structure, which uses a Fourier-domain optical processor as the control element and the fast-scanning optical spectrum analyzer as the feedback component. This system provides low-pass RF response. Experimental results demonstrate a 6-tap microwave photonic filter with a free spectral range of 2.5GHz. The power fluctuation of the first-order passband in RF response is within ±1dB over 20 minutes.

A novel tunable all-optical microwave photonic mixer is presented based on serial phase modulation and an on-chip notch filter. The notch filter breaks the out-of-phase symmetry between the upper and lower sidebands generated from phase modulation, resulting in bandpass response of frequency selection. This system is achieved through an all-optical approach, which does not require electrical components, thus increasing the operation bandwidth of the system. The tunability of frequency selection is achieved through adjusting the wavelength of the optical source. Experimental results verify the technique with a 3rd-order SFDR of 91.7dBm/Hz$^{2/3}$.
STATEMENT OF ORIGINALITY

This is to certify that, to the best of my knowledge, the content of this thesis is my own work. This thesis has not been submitted for any degree or other purposes. I certify that the intellectual content of this thesis is the product of my own work and that all the assistance received in preparing this thesis and sources have been acknowledged.

Jianqiao Ren
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Finally, I would like to dedicate this thesis to my parents, whose love has always been the greatest support through my whole life.
**PUBLICATIONS**


**Jianqiao Ren**, Xiaoke Yi, Suen Xin Chew and Liwei Li, “Tunable all-optical microwave photonic mixer based on serial phase modulation and on-chip notch filter”, in preparation for submission to IEEE Photonics Journal.
DECLARATION

This thesis contains the materials published in:


These are in Section 3.3. I designed the structure, conducted the simulation, carried out the experiment, analyzed the data and wrote the manuscript.

This thesis also contains the materials published in:


These are in Section 3.2. I participated in the structure design, conducted the simulation, carried out the experiment, analyzed the data and wrote the manuscript.

Finally, this thesis contains the materials in preparation for submission:

**Jianqiao Ren**, Xiaoke Yi, Suen Xin Chew and Liwei Li, “Tunable all-optical microwave photonic mixer based on serial phase modulation and on-chip notch filter”, in preparation for submission to IEEE Photonics Journal.

These are in Section 4.2. I designed the structure, conducted the simulation, carried out the experiment, analyzed the data and wrote the manuscript.

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<th>Description</th>
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<tr>
<td>CDR</td>
<td>compression dynamic range</td>
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<tr>
<td>DFB</td>
<td>distributed feedback</td>
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<td>DSB</td>
<td>double sideband</td>
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<td>EDFA</td>
<td>Erbium doped fiber amplifier</td>
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<td>EMI</td>
<td>electromagnetic interference</td>
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<td>EOM</td>
<td>electro-optic modulator</td>
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<td>FBG</td>
<td>fiber Bragg grating</td>
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<tr>
<td>FDOP</td>
<td>Fourier-domain optical processor</td>
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<tr>
<td>FIR</td>
<td>finite impulse response</td>
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<td>FP</td>
<td>Fabry-Perot</td>
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<td>FSR</td>
<td>free spectral range</td>
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<tr>
<td>HD</td>
<td>high-order distortion</td>
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<tr>
<td>HR-OSA</td>
<td>high resolution optical spectrum analyzer</td>
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<tr>
<td>IIR</td>
<td>infinite impulse response</td>
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<tr>
<td>IM</td>
<td>Intensity modulator</td>
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<tr>
<td>IMD</td>
<td>intermodulation distortion</td>
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<tr>
<td>LCoS</td>
<td>liquid crystal on silicon</td>
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<td>LED</td>
<td>light emitting diode</td>
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<td>MZM</td>
<td>Mach-Zehnder modulator</td>
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<td>NF</td>
<td>noise figure</td>
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<td>OIP</td>
<td>output intercept point</td>
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<td>PC</td>
<td>polarization controller</td>
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<td>PD</td>
<td>photodetector</td>
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<td>PM</td>
<td>phase modulator</td>
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<td>RF</td>
<td>radio frequency</td>
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<td>RIN</td>
<td>relative intensity noise</td>
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<td>SBS</td>
<td>stimulated Brillouin scattering</td>
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<td>SFDR</td>
<td>spurious-free dynamic range</td>
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<td>SMF</td>
<td>single-mode fiber</td>
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<tr>
<td>SNR</td>
<td>signal-to-noise ratio</td>
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<tr>
<td>SOA</td>
<td>semiconductor optical amplifier</td>
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<tr>
<td>SSB</td>
<td>single sideband</td>
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<tr>
<td>VNA</td>
<td>vector network analyzer</td>
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<td>WDM</td>
<td>wavelength division multiplexing</td>
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INTRODUCTION

1.1 Motivation

In 1966, Charles Kuen Kao proposed the use of glass optical fibers as the transmission medium for light [1]. In 1962, the first semiconductor laser and the first electro-optic modulator were developed for gigahertz transmission, laying the foundation for the development of microwave photonics [2, 3]. At the same time, the loss of silica fiber was reduced to less than 20dB/km by Corning [4]. These discoveries and development overcome the electronic bottlenecks and bring microwave photonics into a considerable range of applications.

Due to the unique advantages of microwave photonics, extensive investigations were carried out in the field of microwave photonic signal processing. These investigations have been implemented in broad applications such as optical filtering, RF phase shifting, frequency mixing and optical beamforming. Microwave photonic signal processing overcomes the electronic bottlenecks which limit the sampling speed and system bandwidth, and possesses the advantages such as large instantaneous bandwidth, low loss, compact size and immunity to electromagnetic interference [5]. The characteristics of low loss and large bandwidth make silica fibers suitable for transmitting high speed signals. Thus, all-optical microwave photonic signal processors are highly needed. In the past 40 years, many photonic processors with different functionalities have been designed and implemented [6, 7].

However, there are still many challenges existing in the development of microwave photonic signal processing. One critical issue with conventional microwave photonic signal processors is that they require highly stable lasers, which are usually thermally controlled and have external cavities [8, 9]. As the number of optical sources increase, the use of high performance lasers necessarily increases the cost, size and power consumption of the overall system. This is partly addressed by the use of spectrum sliced sources [10, 11], but amplified spontaneous emission has high intensity noise, while supercontinuum sources are complex to design. Therefore it is essential to replace the large numbers of lasers with a single uncooled laser in order to reduce the cost, size and complexity of the system.

Another main challenge existing in microwave photonic signal processors is the use of electrical components, which largely limit the bandwidth of the system and lead to electromagnetic interference [12-14]. For example, the use of RF couplers in microwave photonic mixers limits the range of frequency conversion and the signal processing speed. Therefore it is highly desirable that a microwave frequency converter in an all-optical scheme is designed and implemented in order to improve the system bandwidth and signal processing speed.

1.2 Objective

The main objectives of this thesis are summarized as follows:

(1) To design and implement a new optical beamforming system with dynamic reconfigurability in order to reduce the cost, size and complexity of conventional beamforming networks. This new system applies frequency tracking and dynamic controlling techniques to the system, obtaining stable system output, large operating bandwidth and a full 360° scanning range, even under rapid environmental changes.

(2) To investigate novel optical delay line filters that use uncooled FP lasers, which leads to frequency tracking microwave photonic signal processors that overcome the electronic bottlenecks and possess dynamic reconfigurability and compact size.
(3) To develop a new microwave photonic mixer based on all-optical scheme. This all-optical structure overcomes the electronic bottlenecks by eliminating the use of electrical components which limit the signal processing speed and system bandwidth, and shows a comparable spurious-free dynamic range.

1.3 Major contributions

The major contributions of this thesis are listed as follows:

- A photonic beamforming network that uses an uncooled FP laser as the optical source is proposed and experimentally demonstrated. This is achieved by using rapid, high-resolution optical spectral measurements to track the frequency drift of the uncooled laser, and then reconfiguring a programmable Fourier-domain optical processor to provide compensation. By using an uncooled laser, the need for temperature control of the laser is eliminated, and the number of optical sources is reduced by using the output spectral lines of the laser. The system realizes six wideband microwave photonic phase shifters, and the resulting magnitude and phase responses vary within a $2\sigma$ deviation of 6.1dB and 14.8°, respectively, even when the laser current is changed during the measurement.

- A microwave photonic filter that uses an uncooled FP laser as the optical source is demonstrated. This filter realizes a low-pass magnitude response and wide passband characteristic, by optically shaping the laser signal. The configuration can overcome the frequency and power drift of the uncooled FP laser by using a feedback structure. Experimental results demonstrate a 6-tap microwave photonic filter with a free spectral range of 2.5GHz. The power fluctuation of the first-order passband in RF response within ±1dB over 20 minutes.

- A novel tunable all-optical microwave photonic mixer is presented based on serial phase modulation and an on-chip notch filter. The notch filter breaks the out-of-phase balance between the upper and lower sidebands generated from phase modulation, resulting in bandpass response of frequency selection. This system is achieved through an all-optical approach, which does not require electrical components, thus increasing the operation bandwidth of the system. The tunability of frequency selection is achieved through adjusting the wavelength of the optical source. Experimental results verify the technique with a 3rd-order SFDR of 91.7dBm/Hz$^{2/3}$. 
2.1 Fiber Optic Communication Links

The need of high-speed communication over long distance has been expanded over decades. Previously, the transmission of microwave signals depended on electrical cables and free space. Glass fibers were believed not suitable as information transmission medium due to its high signal loss (1000dB/km). In 1966, Charles Kuen Kao discovered that optical fibers could be used as the transmission medium for light [1]. Kao pointed out that the contaminants in glass fibers were the main limitation of high signal loss. Thus by removing the contaminants, purified fibers could be used as light transmission medium over long distance. In the 1970s, the single mode fibers with attenuation loss less than 20dB/km were realized by Corning [2]. At the same year, GaAs semiconductor lasers were demonstrated to emit light waves, making it possible for transmitting light through fiber optic cables over long distance [3].

In 1977, the very first fiber optic communication link was installed at Bell Laboratory, with a GaAs semiconductor laser as the transmission source operating at 0.8um, and repeaters spaced up to 10km [6, 7]. Since then, fiber optic communication links have been growing exponentially. The spacing between adjacent repeaters was then extended to 50km in fiber optic links, which used InGaAsP semiconductor lasers as the transmitting source operating at 1.5um, and fibers with an attenuation loss of 0.5dB/km. After that, dispersion shifted fibers were demonstrated to reduce the dispersion at 1.55um, which has the lowest loss (0.2dB/km) across optical spectrum [15, 16]. At the same time, the fiber optic communication structure known as wavelength division multiplexing (WDM) was introduced, in which an optical fiber was connected with multiple optical sources at the transmission end [17]. This scheme overcomes the narrow optical bandwidth of the typical point-to-point link, and increases the information capacity of the fiber-optic link dramatically. Recently, the multi-core technique has further improved information transmission capacity [69-72].

Fig.2.1 shows the basic diagram of a fiber-optic communication link. The optical transmitter converts the signal from electrical domain to optical domain, and launches the optical output to the transmission link [18]. In the transmission channel, optical amplifiers are often applied to overcome the transmission loss [19]. At the receiver side, the received signals are converted from optical domain back to electrical domain by photodetectors [20].

Nowadays, fiber optic communication links are widely used because of its advantages of large instantaneous bandwidth, light weight, high transmission speed and immunity to electromagnetic interference. They are applied in the areas such as broadband access networks, submarine systems, defence systems and major telecommunication infrastructure, and perform the complex functions that are impossible to be realized in microwave and RF communication systems [21-23]. The analog link of fiber optic communication systems has become an important field known as microwave photonics.
2.2 Microwave Photonic Signal Processing

Microwave photonics signal processing focuses on processing the microwave signals using optical techniques. It overcomes the drawbacks of electrical signal processing, and has the features such as high sampling frequency, fast transmission speed and dynamic reconfigurability.

The schematic of a general microwave photonic system is shown in Fig.2.2. An optical signal is generated from an optical source, operating at extremely high frequency (commonly 193THz). The optical source provides a continuous optical wave, which can be modulated by an RF signal by internal or external modulation. Electro-optic modulators and phase modulators are the most commonly used devices for external modulation [18,38]. In direct modulation, the optical signal from the light source is modulated by the input current, bringing the disadvantages such as linewidth instability, low extinction ratio and chirping. The typical directly modulated light source is light emitting diodes. In comparison, the externally modulated optical signals have narrower linewidth. One typical externally modulated laser is distributed feedback lasers. Therefore, externally modulated lasers are desirable. The optical-domain signal is obtained by the electrical-domain RF signal producing amplitude modulation via electro-optic modulators or phase modulation via phase modulators [24-26, 74-78]. The RF/microwave signal is generated from an RF source such as antenna, commonly operating from 300MHz to 300GHz.

The output of electro-optic modulator then enters the photonic signal processor, which is the fundamental element of microwave photonic systems [48]. The photonic signal processor modifies the spectral features of optical carrier and sidebands according to the specific requirements set by a particular application [27, 28]. The modified features can be amplitude, frequency, phase, polarization, time delay and dispersion. The photonic signal processor is described by $H(f)$, which is the Fourier transform of the impulse response and which is known as the transfer function. Commonly used optical devices are fiber Bragg Gratings [29-31], Erbium Doped Fiber Amplifiers [32], optical couplers, circulators, polarization controllers and etc. The optically processed signals are then detected by a photodetector, which performs optical-to-electrical conversion. The recovered RF signals are detected at the output of photodetector. Microwave photonic systems support various functions such as optical filtering, phase shifting, sensing and frequency mixing. These functions will be introduced and discussed in the following chapters.

The schematic diagram of a general microwave photonic filter is shown in Fig.2.3. The optical signal from laser source is modulated by an RF input signal in an electro-optic modulator, where the RF domain information is converted into the optical domain. Time delay is added to the modulated optical signal in optical delay line processors, such as chirped fiber Bragg gratings, optical couplers, and Liquid-crystal-on-sicilon-based dispersive devices. The received signals are combined at the photodetector, which performs the optical-to-electrical conversion.
Fig. 2.4 A general optical delay line signal processor

The schematic of a general optical delay line based signal processor is shown in Fig. 2.4, where the input RF signal is delayed, weighted and summed [150, 151].

The output signal at the optical signal processor \( y(t) \) is given by [8]

\[
y(t) = \sum_{n=0}^{N-1} w_n x(t - n\Delta \tau)
\]

(2.1)

where \( n \) is the filter coefficient, \( N \) is the number of taps, \( w_n \) is the amplitude weighting of the \( n \)th tap, \( \Delta \tau \) is the basic time delay. The optical delay line filter is known as Infinite Impulse Response (IIR) with an infinite number of \( N \), and known as Finite Impulse Response (FIR) with finite number of taps \( N \). Due to the discrete and linear time-invariant characteristics, the photonic signal processor can be given in the impulse response form [8]

\[
w(t) = \sum_{n=0}^{N-1} w_n \delta(t - n\Delta \tau)
\]

(2.2)

The transfer function of the photonic signal processor thus can be expressed as [22]

\[
W(f_m) = \sum_{n=0}^{N-1} w_n e^{j2\pi f_m n \Delta \tau}
\]

(2.3)

where \( f_m \) is the microwave frequency of the photonic signal processor. Two critical functions are used to describe the flexibility of microwave photonic filters: tunability and reconfigurability. The tunability of the center frequency is a critical function of realizing flexible filters. Optical delay line filters can be
tuned by changing the basic time delay. A series of taps which are equally spaced in the time domain show a periodic spectral response in the frequency domain due to the discrete characteristic of optical delay line filters. A critical parameter known as Free Spectral Range is used to describe the filter tunability. The FSR is defined as the frequency spacing between the adjacent maximum values of the RF passband. The FSR is represented as the reciprocal of the basic time delay:

\[
FSR = \frac{1}{\Delta \tau}
\]  

(2.4)

Fig.2.5 shows the simulation of the FSR of a 2-tap positive coefficient filter with unity tap weights under different basic time delays. The basic time delay is set at 100ps, 120ps, 140ps and 160ps respectively. It can be seen that the FSR and the basic time delay are inversely proportional, which makes the FSR 10GHz, 8.33GHz, 7.14GHz and 6.25GHz respectively.

Fig.2.5 FSR of a 2-tap positive coefficient filter with unity tap weights under different basic time delays

The filter reconfigurability is the capacity to reconfigure the filter shape through applying different window functions. For the transfer function in Eq.2.3, the reconfigurability can be achieved by changing the tap weights. In practical implementations, this is accomplished by changing the power of individual laser sources, the optical amplifier coefficients or the attenuation coefficients of the optical processor.

Based on the number of filter taps, microwave photonic filters can be classified as Finite Impulse Response filters and Infinite Impulse Response filters. FIR filters, which have finite number of filter taps, can be realized through single source configuration or multi-source configuration [22]. The schematic of the FIR filter based on single source is shown in Fig.2.6. The optical signal from a continuous wave optical source is modulated by an RF signal, where the RF signal is converted to the optical domain. The modulated optical signals are fed into a \(1 \times N\) splitter, which is connected to N optical delay lines. Each optical delay line has its amplification factor \(a_n\) and delay factor \(n\tau\). The weighted, delayed optical signals are summed by an \(N \times 1\) coupler. The combined optical signals are sent to the photodetector for optical-to-RF conversion. The schematic of the FIR filter based on multiple optical sources is shown in Fig.2.7. Multiple optical signals generated from independent optical sources experience the same time delay. The power of the optical signals are controlled to form different windowing functions. The sideband suppression ratio of the multi-source FIR filters can be controlled by applying different windowing functions. Compared with single source based FIR filters, multi-source based FIR filters are more easily to be programmable. The power of each optical signal can be controlled to achieve desired frequency response.
The schematic of a general IIR filter is shown in Fig.2.8. The infinite number of taps in IIR filters can be generated from a recursive structure, which provides higher frequency selectivity [22]. An optical coupler is applied to connect the modulated optical signals with the loop which provides optical gain and time delay. The gain medium and the delay element decide the FSR and the transfer function of the filter. The number of recirculating taps can be adjusted through controlling the gain medium. The multiple optical taps at the output of the optical loop are summed and combined at the photodetector, causing the phase induced intensity noise (PIIN) to become the dominant noise. The PIIN noise can be reduced by using differential photodetection, which breaks the coherence time of the optical source.

2.3 Analog Performance of Fiber Optic Links

In this section, the optical aspects of microwave photonic systems are demonstrated, with the most important analog performance parameters defined and discussed. This section begins with an introduction of the scattering matrix, which is used to quantify a noisy RF system. Next, the definition and mathematical expression of noise sources are introduced. In reality, a fiber optic link is nonlinear, thus the nonlinear distortion under tone test is analyzed, yielding a term Spurious-free dynamic range, which is often used to describe the performance of microwave photonic mixers [139]. Finally, this
section is concluded with cascade analysis, which describes how individual performance metrics of each stage in a cascaded system influence the overall performance.

2.3.1 Scattering Matrix

The scattering matrix is often used in the analysis of RF network to define the relationship between the RF input and RF output [33]. For a linear N-port RF network, the scattering matrix is expressed as

\[
\begin{bmatrix}
V_{1\text{out}}^	ext{in} \\
V_{2\text{out}}^	ext{in} \\
\vdots \\
V_{N\text{out}}^	ext{in}
\end{bmatrix}
=
\begin{bmatrix}
S_{11} & S_{12} & \cdots & S_{1N} \\
S_{21} & S_{22} & \cdots & \vdots \\
\vdots & \vdots & \ddots & \vdots \\
S_{N1} & S_{N2} & \cdots & S_{NN}
\end{bmatrix}
\begin{bmatrix}
V_{1\text{in}} \\
V_{2\text{in}} \\
\vdots \\
V_{N\text{in}}
\end{bmatrix}
\]

(2.5)

Fig.2.9 Schematic of scattering matrix

where \(V_{N\text{in}}^{\text{in}}\) and \(V_{N\text{out}}^{\text{out}}\) are the input and output voltage of the \(n\)th port, and the elements are known as scattering coefficients. Most of the RF networks are two-port as shown in Fig.2.9, thus reducing Eq.2.1 to

\[
\begin{bmatrix}
V_{1\text{out}}^	ext{in} \\
V_{2\text{out}}^	ext{in}
\end{bmatrix}
=
\begin{bmatrix}
S_{11} & S_{12} \\
S_{21} & S_{22}
\end{bmatrix}
\begin{bmatrix}
V_{1\text{in}} \\
V_{2\text{in}}
\end{bmatrix}
\]

(2.6)

For a general externally modulated optic fiber link, Port 1 is a modulation device, connected with an RF input signal, and Port 2 is a photodetector, connected with load impedance (which is often the vector network analyzer). Thus \(V_{1\text{out}}^{\text{out}} = S_{11}V_{1\text{in}}^{\text{in}} + S_{12}V_{2\text{in}}^{\text{in}}\), and \(V_{2\text{out}}^{\text{out}} = S_{21}V_{1\text{in}}^{\text{in}} + S_{22}V_{2\text{in}}^{\text{in}}\). The elements in \(S\) matrix are often described as a function of frequency. Specifically, \(S_{11}\) and \(S_{22}\) describe the reflection of the modulation device and photodetector, exhibiting periodic peaks and nulls. \(S_{12}\) describes the noise floor of the signal analyzer, because the RF signal cannot be transmitted in the counter direction, which means from the photodetector to the modulator [140]. The \(S_{21}\) element is the Port1-to-Port2 transmission coefficient [33], which is often known as the RF power gain, or the system conversion efficiency.

The conversion efficiency is the power ratio of the RF output to the RF input[141]. Because the conversion between RF domain and optical domain and the optical signal processing involves inevitably signal loss, \(S_{21}^2\) is often expressed as the insertion loss of the optical devices [33]. The mathematical expression for the conversion efficiency is [34]
\[
G_{\text{conv}} = \frac{P_{\text{out,RF}}}{P_{\text{in,RF}}}
\]  

(2.7)

In a microwave photonic mixer, the interested mixing component is the IF signal, thus \( P_{\text{out,RF}} \) refers to the output IF signal power in mixers [141]. In practice, the linear \( G_{\text{conv}} \) is often expressed in logarithmic scales with a unit of \( dB \), which gives \( G[dB] = 10 \log(G_{\text{conv}}) \). The final expression of the conversion efficiency in a microwave photonic mixer under different modulation schemes will be discussed in Chapter 4.

### 2.3.2 Noise Sources

The noise performance of a microwave system is often described by Noise Figure, which describes the degradation of the signal-to-noise ratio [142]. Noise Figure is defined as the ratio of 1) the total output noise power spectral density to 2) the portion of 1) generated at the input termination [33]. The output noise power spectral density is denoted as \( N_{\text{out}} \). The input thermal noise power is given by

\[
P_{\text{th}} = K_B T B
\]

(2.8)

where \( K_B \) is Boltzmann constant, \( T \) is temperature, and \( B \) is the signal bandwidth [33]. Multiplying Eq.2.8 with \( G_{\text{conv}} \) and normalizing to a unit bandwidth, gives the Noise Figure as

\[
F = \frac{N_{\text{out}}}{gK_B T_s}
\]

(2.9)

As shown in Eq.2.9, to minimize the noise figure, the conversion efficiency has to be increased. The definition of noise factor can be expressed by using SNR, which gives [33]

\[
NF = \frac{\text{SNR}_{\text{in}}}{\text{SNR}_{\text{out}}} = \frac{s_{\text{in}}/n_{\text{in}}}{s_{\text{out}}/n_{\text{out}}}
\]

(2.10)

where \( s_{\text{in}} \) and \( s_{\text{out}} \) are the power of the input signal and the output signal, and \( n_{\text{in}} \) and \( n_{\text{out}} \) are the total noise power of the input signal and the output signal [153]. Note that for a microwave photonic mixer, \( s_{\text{out}} \) is the output power of the desired mixing product, while \( s_{\text{in}} \) is the input RF signal power. The noise figure in dB scale under the standard temperature is expressed as

\[
NF[\text{dB}] = 10 \log(F) = 174 + N_{\text{out}}[\text{dBm/Hz}] - G[\text{dB}]
\]

(2.11)

where \( 10 \log(K_B T_s) = -174 dB / Hz \). Since the thermal noise cannot be eliminated, the lowest noise is the thermal noise level, which makes the range of noise figure as \( F \geq -1 \) and \( NF \geq 0 dB \).

There are three major sources of noise in fiber optic links, which are thermal noise, shot noise and laser intensity noise [33, 35]. Thermal noise is generated from the thermal motion of the electrons in resistors. It exists with and without applying voltage to resistors. Thermal noise is constant over the frequency spectrum. Thus, thermal noise is also called white noise, and exists both in load resistors and modulation devices. Note that the thermal noise generated from modulation device is amplified by a factor of conversion efficiency \( G_{\text{conv}} \).
Shot noise appears at the photodetector because of the random statistical fluctuations of the arrived photons \([34]\). It is linearly dependant on the detected average photocurrent. Shot noise power is given as

\[
N_{\text{shot}} = 2qI_{\text{avg}} BR_L
\]  

(2.12)

where \(q\) is the electron charge, \(I_{\text{avg}}\) is the average detected photocurrent, and \(R_L\) is the load resistor.

Laser intensity noise is generated from the fluctuations of phase and frequency of the optical signals when they are not modulated. These fluctuations are caused by the spontaneous emission of phonons \([33, 36]\). The laser intensity noise relied more on the photocurrent than shot noise, and is expressed as

\[
N_{\text{RIN}} = \text{RIN} I_{\text{avg}}^2 BR_L
\]  

(2.13)

where \(\text{RIN}\) is the laser relative intensity noise \([33]\). With the increase of laser input power, \(N_{\text{RIN}}\) becomes the dominant noise in RF networks. Assuming the three dominant noise sources are independent from each other, the total link noise in fiber optic links is the sum of all noise sources, which is given as \([33]\)

\[
N_{\text{tot}} = N_{\text{th}} + N_{\text{th,amp}} + N_{\text{shot}} + N_{\text{RIN}}
\]

\[
= (1 + G_{\text{tot}})K_BT B + 2qI_{\text{avg}} BR_L + \text{RIN} I_{\text{avg}}^2 BR_L
\]

(2.14)

where \(N_{\text{th,amp}}\) is the amplified thermal noise, which depends on the system conversion efficiency. Therefore, at low laser input power, the noise power is mainly from thermal noise, which cannot be decreased as it is independent of optical injection. At high laser input power, RIN becomes the dominant noise source in the fiber optic link. The only noise that cannot be decreased is the shot noise. Therefore, the aim for many microwave photonic mixers is to reach the limit of shot noise \([152]\).

### 2.3.3 Nonlinear Distortions

The derived expressions of conversion efficiency and noise figure are mainly applied to linear fiber optic links. A system is said to be linear if the principle of superposition applies. However, no fiber optic link is perfectly linear. A nonlinear distortion analysis of the fiber optic link is very important especially for microwave photonic mixers, since the major element performing the mixing function, i.e. the modulation device, has intrinsic nonlinearity. Other optical components such as the photodetector and the photonic signal processor also have nonlinearity. However, their nonlinear effects can be ignored as they are much smaller than that of the modulation device.

The nonlinearity of a fiber optic link can be represented by a Taylor series, where the output voltage is described as a function of the input voltage:

\[
V_{\text{out}}(V_{\text{in}}) = a_0 + a_1(V_{\text{in}} - V_b) + a_2(V_{\text{in}} - V_b)^2 + a_3(V_{\text{in}} - V_b)^3 + \ldots
\]

(2.15)

where \(V_b\) is the bias voltage, and \(a_m\) is given by

\[
a_m = \frac{1}{m!} \frac{d^m V_{\text{out}}}{dV_{\text{in}}^m} \bigg|_{V_{\text{in}} = V_b}
\]

(2.16)
The Eq.2.15 uses a small-signal approximation, where an expansion at a bias point is applied under the condition that the input is very small. Therefore, when the deviation from the operating point is minimal, linear theory can be used to determine the system output.

A simple measure of the linearity of the fiber optic link is by performing a tone test [3]. With a single-tone modulation, an input signal with only one frequency \( V_{in}(t) = V_b + V \sin(\omega t) \) is launched into system, yielding an output

\[
V_{out} = \left( a_0 + a_1 V \right) + \left( a_V + \frac{3a_1 V^3}{4} \right) \sin(\omega t)
- \frac{a_2 V^2}{2} \cos(2\omega t) - \frac{a_1 V^3}{4} \cos(3\omega t) + \cdots
\]  

(2.17)

The output terms in Eq.2.17 can be classified into three types: a DC offset with no oscillation, a term at fundamental frequency \( \omega \), and harmonic distortions with multiple integer of fundamental frequency. Note that the term with \( n \) times of the fundamental frequency is called the \( n \)th order harmonic distortion. A term Compression Dynamic Range was created to describe the influence of harmonic distortions made on the fundamental distortions. CDR is the power range that the input signal is above the noise floor, and the output signal is suppressed relative to linear a response by a certain amount [33]. Mathematically,

\[
CDR_{dB} = \frac{P_{out}^{10dB}}{N_{out}^B}
\]

(2.18)

where \( P_{out} \) is the output power at \( xdB \) compression. However, it is not practical to determine the system nonlinearity from CDR due to the multioctave nature of harmonic distortion [33]. For example, a nonlinear system with an input signal at 30GHz would yield second- and third-order harmonic distortions at 60GHz and 90GHz respectively, which are beyond the measurement range of network analyzers. Rather, a two-tone test is often applied to determine the intermodulation distortion.

Under a two-tone test, an input signal expressed as \( V_{in}(t) = V_b + V \sin(\omega_1 t) + V \sin(\omega_2 t) \) is launched into the nonlinear system, where \( \omega_1 = 2\pi f_1 \) and \( \omega_2 = 2\pi f_2 \) are the two-tone angular frequencies. By substituting the input equation into Eq.2.15 and applying trigonometric relations, the resultant output is as follows [33]

\[
V_{out} = \left( a_0 + a_1 V \right) + \left( a_V + \frac{9a_1 V^3}{4} \right) \sin(\omega_1 t)
+ \left( a_V + \frac{9a_1 V^3}{4} \right) \sin(\omega_2 t)
- \frac{a_2 V^2}{2} \cos(2\omega_1 t) - \frac{a_2 V^2}{2} \cos(2\omega_2 t)
+ a_1 V^3 \cos[(\omega_1 - \omega_2)t] - a_1 V^3 \cos[(\omega_1 + \omega_2)t]
- \frac{a_3 V^3}{4} \sin(3\omega_1 t) - \frac{a_3 V^3}{4} \sin(3\omega_2 t)
+ \frac{3a_2 V^3}{4} \sin[(2\omega_1 - \omega_2)t] + \frac{3a_2 V^3}{4} \sin[(2\omega_2 - \omega_1)t]
- \frac{3a_3 V^3}{4} \sin[(2\omega_1 + \omega_2)t] - \frac{3a_3 V^3}{4} \sin[(2\omega_2 + \omega_1)t] + \cdots
\]

(2.19)

From Eq.2.19, it is clear that besides the harmonic distortions, additional mixing products are generated at the output. These components are known as intermodulation distortions. The terms at \( \omega_1 - \omega_2 \) and \( \omega_1 + \omega_2 \) are known as second-order intermodulation distortions (IMD2), which possess the sum and the difference of the input fundamental frequencies \( \omega_1 \) and \( \omega_2 \). Similarly, the terms at \( 2\omega_1 - \omega_2, 2\omega_2 - \omega_1, \)
$2\omega_1 + \omega_2$ and $2\omega_2 + \omega_1$ are known as the third-order intermodulation distortions (IMD3), which are the combination of sum and difference of the two modulating frequencies.

Fig. 2.10 shows an illustrative power spectrum for a single-tone test and a two-tone test. In practice, when the two fundamental frequencies are close to each other, the IMD3 terms at $2f_1 - f_2$ and $2f_2 - f_1$ are problematic because they are too close to the fundamental frequencies to be filtered out. For example, a two-tone input signal at 5.9GHz and 6GHz will yield two IMD3 terms at 5.8GHz and 6.1GHz. Given that many existing optical filters have a 3-dB bandwidth wider than 1GHz, the two spurious mixing products cause severe interference.

### 2.3.4 Spurious-free Dynamic Range

The influence of IMD on the fiber optic link is described with Spurious-free dynamic range. The SFDR is defined as the range that the output signal power is above the output noise floor and all spurious signals are less than or equal to the output noise floor [3, 4, 33]. Fig. 2.11 illustrates the definition of SFDR in dB scale. In the figure, the x-axis is the RF input power, and the y-axis is the output power of the fundamental component and the limiting distortion component. Note that the slopes of the fundamental component and the nth-order IMD are 1 and n respectively. The output noise is also included in the plot. The SFDR decreases with the rise of noise floor. It is also clear from the figure that the SFDR can either be measured along the noise floor or to the noise floor.

![Fig.2.11 Spurious-free dynamic range](image)
The SFDR calculation involves a variable known as the nth-order output intercept point \((OIP_n)\) [143]. \(OIP_n\) is the intercept point of the fundamental component and the nth-order \(IMD\) under the same laser input power. The nth-order SFDR \((SFDR_n)\) can be expressed as a function of the \(OIP_n\) [143]

\[
SFDR_n = \left(\frac{OIP_n}{N_{out}^B}\right)^{(n-1)/n}
\] (2.20)

The dB-scale form of Eq.2.16 is given as [143]

\[
SFDR_n[\text{dB}] = -\frac{n-1}{n} \left\{ OIP_n[\text{dBm}] - N_{out} \left[ \frac{\text{dBm}}{\text{Hz}} \right] - 10\log(B[\text{Hz}]) \right\}
\] (2.21)

The \(SFDR_n\) given in Eq.2.20 and Eq.2.21 has a unit of \(Hz^{n-1/n}\) and \(dB \cdot Hz^{n-1/n}\) respectively. The use of \(OIP_n\) in the calculation of SFDR is based on the assumption that the output noise remains constant across the RF input power.

To determine the value of \(OIP_n\), a widely used method is by measuring the power of fundamental component, \(IMD_n\) and output noise power under different input RF power, and plotting the measured data to extrapolate the intersection point [144]. Another method is simply by measuring a single-point RF response, which is only acceptable when the measured data matches the device characterisation. For the second measurement method, the \(OIP_n\) is given as [143]

\[
OIP_n = \left(\frac{P_n}{P_{\Omega}}\right)^{1/(n-1)}
\] (2.22)

where \(P_{\Omega}\) and \(P_n\) are the output power of the fundamental or desired component and the nth-order distortion \((IMD_n)\) respectively. The \(OIP_n\) in dB form is given as [143]

\[
OIP_n[\text{dBm}] = \frac{1}{n-1} \left( n \cdot P_{\Omega}[\text{dBm}] - P_n[\text{dBm}] \right)
\] (2.23)

As shown in Eq.2.23, the amplitude of the \(IMD_2\) is twice larger than \(HD_2\), and the \(IMD_3\) is three times larger than \(HD_2\). Since power is the quadratic function of amplitude, the power of \(IMD_4\) is four times larger than \(HD_2\), and the \(IMD_3\) is nine times larger than \(HD_3\). The SFDR can also be calculated by involving the input intercept point \((IIP_n)\) by simply dividing the \(OIP_n\) with conversion efficiency \(G_{conv}\), which gives

\[
IIP_n = \frac{OIP_n}{G}
\] (2.24)

Thus, another mathematical expression for \(SFDR_n\) is derived as Eq.2.25 with the use of the definition of noise figure:

\[
SFDR_n = \left(\frac{IIP_n}{FK_B T B}\right)^{(n-1)/n}
\] (2.25)

\(SFDR_n\) can also be given in the dB form as [143]
A high SFDR is important in many RF applications. There are a few ways to improve the SFDR. Reducing the noise figure will cause the increase of SFDR. Also, removing the limiting IMD will also improve the SFDR.

2.3.5 Cascade Analysis

The presentation of RF performance in previous sections is applied in a singular system. In practice, a cascaded system which is composed of a series of independent systems is often used. The schematic of an N-stage cascaded system is shown in Fig.2.12. Each individual stage has a known conversion efficiency, noise figure and output intercept point [33].

The conversion efficiency of a cascaded RF system is simply the multiplication of each stage, which is given as

\[ g = \prod_{p=1}^{N} g_p \]  

(2.27)

Its dB form is given as

\[ G[dB] = \sum_{i=1}^{N} G_i [dB] \]  

(2.28)

The cascaded noise figure is as follows

\[ F = F_1 + \sum_{i=2}^{N} \left( F_i - 1 \right) \prod_{p=1}^{i-1} g_p \]  

(2.29)

When each individual noise figure is relatively low, the first-stage noise figure often becomes the dominant noise. The \( OIP_n \) in a cascaded architecture is given as [33]

\[ OIP_n = \left\{ \sum_{i=2}^{N} \left[ \prod_{p=i}^{N} g_p \right]^{(1-n)/2} \right\} + OIP_{aN}^{(1-n)/2} \]  

(2.30)
2.4 Optical Beamforming

In this section, the fundamental principles of optical beamforming is given. First, the concept of phased array antenna is introduced, with its basic characteristics given. Next, the optical beamforming techniques based on different optical phase shifters are analyzed. These techniques involve the use of fiber Bragg grating, stimulated Brillouin scattering, dual-electrooptic modulators and semiconductor optical amplifiers.

2.4.1 Phased Array Antenna Beamforming

Phased array antenna is an array of antennas in which the relative phases of the respective signals feeding the antennas are varied, so that the radiation pattern is strengthened in one direction and weakened in other directions (Fig.2.13) [37]. Beamforming techniques for phased array antenna can be realized either through electronic or photonic approaches. Photonic beamforming approaches have been widely used because of its advantages such as compact size, large instantaneous bandwidth, low loss and immunity to EMI [154].

![Fig.2.13 Schematic of phased array antenna [37]](image)

Two common approaches are used for photonic beamforming. One is by creating phase shift to the signals in radiating elements. The phase shift creates interference between the radiating signals, so that some signals are strengthened by constructive interference and some signals are weakened by destructive interference. In this way the desired beamforming direction can be controlled [38-41]. Another approach is by creating time delay among the signals in different radiating elements. Different time delay result in different radiating angles [42-47]. Before discussing these two approaches, the basic characteristics for phased array antenna are introduced in this section.

**Radiation pattern**

Radiation pattern is the energy radiated from the phase array antennas. To measure the radiation pattern, the measurement point is set a fixed location from the antenna, and the energy is measured. The excitement of antenna array includes amplitude control and phase control, which can be achieved through Fourier calculation [37]. Each array is controlled individually and their combination forms the whole array pattern. A general linear array pattern is shown in Fig.2.14. The pattern is given by

\[
AF(\theta) = \sum A_n \exp\left[jkd(n-1)(\sin \theta - \sin \theta_n)\right]
\]

(2.31)
where \( \lambda \) is wavelength, \( d \) is the element spacing, \( k = \frac{2\pi}{\lambda} \) is the angular frequency in spacing, \( u = \sin \theta - \sin \theta_0 \) is the angular variable, \( \theta_0 \) is the scan angle. For uniform excitation, Eq.2.31 can be expressed as [37]

\[
AF(\theta) = \exp \left[ j\pi(N-1)(\sin \theta - \sin \theta_0) \frac{\sin \left( N\pi \frac{d}{\lambda} (\sin \theta - \sin \theta_0) \right)}{N \sin \left( \pi \frac{d}{\lambda} (\sin \theta - \sin \theta_0) \right)} \right] \quad (2.32)
\]

where \( N \) is the number of array elements, \( kd \sin \theta_0 \) is the phase shift between elements. The beam direction can be tuned by changing the phase shift.

### Beam squint

Beam squint is a phenomenon that the beam changes its direction as a function of operating frequency [37]. This causes a variation of the gain at a certain direction, limiting the bandwidth of system. Beam squint can be eliminated by photonic time-delay approaches because it is independent of frequency, thus widening the operation bandwidth.

### Beamwidth

The 3dB beamwidth is the width of the angle at the half-power point. The increase of the beamwidth causes the decrease of the side lobe level. The 3-dB point of an array pattern is given by [37]

\[
\frac{\sin \left( N\pi \frac{d}{\lambda} (\sin \theta - \sin \theta_0) \right)}{N \sin \left( \pi \frac{d}{\lambda} (\sin \theta - \sin \theta_0) \right)} = 0.5 \quad (2.33)
\]

### Bandwidth

The bandwidth of phase array antenna beamforming is influenced by elements spacing and beamwidth. For a uniform array, the bandwidth can be expressed as [37]
\[ BW = \frac{f_2 - f_1}{f_0} = \frac{(\sin \theta_2 - \sin \theta_1) \sin \theta_0}{\sin \theta_1 \sin \theta_2} \]  

(2.34)

where \( f \) is the scanning frequency and \( \theta \) is the pointing angle.

**Directivity**

Directivity describes the radiation ratio of the pointing direction that over all directions. For a linear phased array, the directivity is given by [37]

\[
D_{array} = \frac{4\pi |F|^2}{\int_0^{2\pi} \int_{-\pi/2}^{\pi/2} |F(\theta, \phi)|^2 \cos \theta d\theta d\phi}
\]

(2.35)

2.4.2 Optical Beamforming based on Microwave Photonic Phase Shifters

As introduced in Section 2.4.1, optical beamforming can be achieved through phase shifting or time delay. This section will focus on the optical beamforming techniques based on different phase shifting methods.

**FBG based microwave photonic phase shifter**

Fiber Bragg Grating has wide application in microwave photonic phase shifters. FBG is an optical reflector constructed in a length of fiber. The optical wavelength reflected along the FBG is decided by the refractive index. By having a periodic change in the refractive index, a specific wavelength is reflected while all other wavelengths are transmitted. By tilting FBG or polarization control, the FBG can achieve fast and slow light effect, and this effect can be used in phase shifters.

The schematic diagram of the microwave photonic phase shifter based on tilted FBG is shown in Fig.2.15 [48]. In normal FBGs, the refractive index has the same direction as the core of fiber. However, in tilted FBGs, there is an angle between the refractive index and the core of fiber. Thus, two coupling modes are generated from this angle. One is between the forward direction and backward direction, and the other is between the cladding and the counter-propagation direction. As a result of the two coupling modes, two wavelength resonances are created, which are given by [48]

\[
\lambda_{Bragg} = \frac{2n_{eff,core} \Lambda_g}{\cos \theta}
\]

(2.36)

\[
\lambda_{coupling} = \left(n_{eff,cladding} + n_{eff,core}\right) \frac{\Lambda_g}{\cos \theta}
\]

(2.37)

where \( \theta \) is the tilt angle, \( \Lambda_g \) is the grating period, \( n_{eff,core} \) and \( n_{eff,cladding} \) are the refractive indices of the core mode and the cladding mode. Based on Karmers-Kronig relations, the change of amplitude causes the change of phase. Therefore, by changing the refractive index along the TFBG, the wavelength resonance is shifted, thus the phase shift can be adjusted. The frequency response of this structure can be tuned by adjusting the power of pump wave to the tilted FBG. A 280° phase shifting tuning range is achieved in this structure.
Another method of FBG based microwave photonic phase shifters is combining a polarization-maintaining fiber Bragg grating (PM-FBG) with a variable retardation plate (VRP) [38]. The structure is shown in Fig.2.16. An optical signal enters the PM-FBG after single-sideband modulation. The PM-FBG has a slow axis and a fast axis, both have a reflection point. The reflection point of the fast axis is set at the optical carrier frequency, which means that all the wavelengths can transmit through the fast axis except the optical carrier. The reflection point of the slow axis is set at another frequency which is different from the optical carrier frequency and the RF modulating frequency. This means that both the carrier and the sideband can pass the slow axis. The fast axis and slow axis are orthogonally polarized. The output of the PM-FBG enters the VRP, where a phase shift is applied to the polarized signals. This method achieves a full 360° phase shifting range from 10 to 40GHz.

Stimulated Brillouin scattering based microwave photonic phase shifter

Brillouin scattering is a phenomenon caused by the nonlinearity of transmission medium. This nonlinearity converts the incident photon into a photon and a phonon [49, 50]. The scattered photon transmits in the counter-propagating direction as the incident photon. Also, due to the scattering, the scattered photon has lower energy than the incident photon. At low optical powers, this effect can occur spontaneously. At high optical power, the scattering has a stimulated effect. The stimulated Brillouin scattering (SBS) involves an optical pump. This pump provides optical gain at a certain frequency, thus amplifying the counter-propagating wave [51]. SBS is frequently encountered in fiber. For higher pump power, SBS gain increases greatly, but it leads to chaotic fluctuations of the powers instead of stable situation.

The advantages of narrow-linewidth and low threshold makes SBS widely used in microwave photonic phase shifters [50-57]. The principle is to use a pump wave which transmits in the opposite direction to counter-propagate with a modulated signal. This pump wave also gives a phase shift to the modulated optical signal. Thus, the beating between the phase shifted optical signal and its sideband generates an RF phase shifter.

One SBS based microwave photonic phase shifter was proposed as shown in Fig.2.17 [54]. In this structure, the optical sideband generated from single-sideband modulation works as the pump wave, which applies phase shift to the optical carrier. The pump wave provides optical gain, which is called Brillouin gain, and the Stokes wave provides Brillouin loss. The location of the gain and loss can be tuned by changing the frequency differences among the pump wave, Stokes wave and the optical signal. To generate only phase shift without amplification or loss, the frequency can be changed so that the gain and loss can be compensated [54].

![Fig.2.15 Schematic of the TFBG based microwave photonic phase shifter [48]](image)

![Fig.2.16 Schematic of the PM-FBG and VRP based microwave photonic phase shifter [38]](image)
Another structure is shown in Fig.2.18 [155], which uses the SBS effect and vector-sum technique [155]. In this scheme, the phase shift is applied on the sideband. Two optical sidebands are generated through phase modulation (PM) and introduced to the SBS processing module. Generally, the symmetrical sidebands generated from phase modulation have $\pi$-phase difference, making no RF signal recovered at the PD. To obtain phase shift, a technique is needed to break the $\pi$ phase difference between the symmetrical sidebands, which is SBS in this structure. The SBS processing is employed in a dispersion shifted fiber (DSF), in which the dispersion effects on the power variation can be ignored.

The phase shift to a modulated optical signal can be either added on the optical carrier or the generated sideband. In this scheme, the phase shift is added on the sideband by a pump wave. The pump wave interacts with the selective optical sideband, resulting in the phase shift as well as amplification on the sideband, which is known as the phase-to-intensity modulation conversion. To achieve a broadband phase shifter, the pump signal and the sideband are set at a fixed frequency distance. The desired RF signal is obtained after photodetection through vector-sum technique.

Microwave photonic phase shifters based on SBS have the advantages of creating a different phase shift while maintaining the signal magnitude [145, 146]. However, this structure requires a high pump power. Also, the environmental change would affect the stability of the system.
Dual-electrooptic modulators based microwave photonic phase shifter

Another phase shifting method is based on dual-electrooptic modulators [40]. In this scheme, two independent optical sources with the same wavelengths connected with two intensity modulators respectively (Fig.2.19). An RF coupler is used to create an in-phase and quadrature component of the modulating signal. The intensities of the input signal of the modulators are $I_2$ and $I_3$, and the bias voltages are $V_2$ and $V_3$. An optical coupler is used to combine the output signals at the two modulators. In traditional structures, when the two optical sources operate at the same wavelength, optical interference will occur. However, this structure uses two individual sources, so that the interference at the optical coupler is minimized [147]. If we set $I_2 = I_3 = I$, and $V_3 = \frac{V_2}{2} - V_2$, the RF signal recovered from photodetection can be expressed as [40]

$$E_{\text{RF}}(t) \propto J_1 \left( \pi V_m / V_\pi \right) I \left[ \sin \left( \omega t + \phi + \theta \right) \right]$$  \quad (2.38)

where $J_1$ is the first-order Bessel function of the first kind, $V_m$ is the amplitude of the RF signal, $\omega$ and $\phi$ are the angular frequency and the initial phase of the RF signal [147], and $\theta = \pi V_3 / V_\pi$ is the RF phase shift. Therefore, the RF phase shift can be tuned by adjusting the bias voltage. Note that the change of bias voltage would cause the change of output power of the modulators, we can compensate this by tuning the output power of the optical sources. The fiber lengths are calibrated to eliminate the frequency dependence of the phase shift.

![Schematic of the dual-EOM](image)

In the proof-of-concept experiment, two optical signals from two individual laser sources operating at the same wavelength are launched into the two intensity modulators. To reduce the loss caused by polarization, two polarization controllers are applied in the system. A hybrid coupler is used for the modulating signal. The phase shift is measured on the VNA through RF frequency sweeping.

The tuning range of this structure can be obtained by changing the modulator bias voltage. However, one common problem for intensity modulators is the use of electrical components, which limit the bandwidth of the system. Experimental results demonstrate a phase shift range of RF signal from 7GHz to 10GHz. Also, to apply this scheme in other applications such as filters, more EOMs are needed, which increase the size, cost and complexity of the system.
Semiconductor optical amplifier based microwave photonic phase shifter

A $2\pi$ microwave photonic phase shifter based on SOA was reported in [58]. This scheme also applies the slow and fast light effects in a SOA, achieving 360° RF phase tuning range. This structure applies the phase shift on the sideband, thus creating a RF phase shift. The structure of this phase shifter is shown in Fig.2.20 [58]. The tuning of phase shift is obtained by changing the feed current, which is controlled by the EDFA. An optical notch filter which is FBG is used for phase shifting.

![Fig.2.20 Schematic of the SOA based microwave photonic phase shifter [58]](image)

The output at PD is given by [58]

$$i(\Omega) \propto |E_0^{\text{out}}| |E_1^{\text{out}}| e^{i(-\phi_0 + \phi_1)} + |E_0^{\text{out}}| |E_{-1}^{\text{out}}| H_{\phi_0 - \phi_1} e^{i(\phi_0 - \phi_1 - \phi_H)}$$  (2.39)

where $\phi_0$, $\phi_1$ and $\phi_H$ are the phase shift applied on the optical carrier, upper sideband and lower sideband. The phase shifts imposed on the upper sideband and the lower sideband both have positive value. Since the lower sideband has larger power than the upper sideband, by imposing a negative phase change on the lower sideband, the phase transition can be achieved. In the experiment, the phase transition is realized through applying an FBG as the notch filter. The FBG used provides an attenuation over 40 dB at the notch center and a 3dB bandwidth of 10GHz, which can provide slow and fast light effects to the total phase shift.

In a single SOA, only tens of degrees of phase shift are obtainable. This scheme increases the phase shifting range by using cascaded SOAs. However, the amplitude response of this structure varies when the phase response changes, thus limiting the operating bandwidth of this phase shifter.
3.1 Introduction

Optical beamforming techniques possess the advantages of large instantaneous bandwidth, antenna remoting capability, compact size and immunity to electromagnetic interference [148]. These systems are usually either based on true-time delay of the optical taps or phase shifting the taps relative to each other[58-68]. Various methods for microwave photonic phase shifters have been demonstrated previously, including the use of Bragg gratings, stimulated Brillouin scattering, dual electrooptic modulators (EOM), and semiconductor optical amplifiers. These approaches are not reconfigurable, and have limitation in their operation bandwidth or amplitude response which is varied with phase response. A promising approach for optically beamforming system is the use of spatial light modulators (SLM) due to the advanced optical signal processing functions they provide, and the use of a Fourier-domain optical processor (FDOP) has been demonstrated to dynamically tailor the amplitude and phase response of a modulated laser to provide a programmable RF phase shift [41, 79].

One potential disadvantage of optical beamforming systems is the reliance on highly stable lasers, which are usually thermally controlled and have external cavities. As the number of beamforming elements increases, the use of high performance lasers necessarily increases the cost, size and power consumption of the overall system. This is partly addressed by the use of spectrum sliced sources, but amplified spontaneous emission has high intensity noise, while supercontinuum sources are complex to design [10, 11].

In Section 3.2, a multiple wavelength optical controlled beamforming network is demonstrated using an uncooled Fabry-Perot laser and an LCoS-based FDOP as the programmable amplitude and phase element; a fast-scanning, high resolution optical spectrum analyser is used to monitor the multiwavelength lasing spectrum of the FP laser, and the detected information is fed back into the control system, which modifies the FDOP in real time. In this way, the power and frequency shift due to ambient temperature is compensated without thermally controlling the laser, which reduces the power consumption of the laser. Furthermore, since the FDOP is based on industrial hardware that exists in modern wavelength selective switches, and high resolution optical channel monitors are now commercially available, this system is a route to a robust, low power consumption beamforming system. Additionally, since the microwave phase shifters in the array are independent and have quasi-continuous phase control from 0 to $2\pi$, arbitrary scanning beam angles can be realized.

In the field of microwave photonic signal processing, finite impulse response (FIR) microwave photonic filters are of particular interest as the amplitude and phase response are easy to manipulate [84-86]. A common FIR structure includes a laser array, which is modulated by RF signals an then sent to a chirped fiber Bragg grating, which performs as a discrete time microwave photonic signal processor [87]. Nevertheless, high resolution signal processing requires a large number of wavelengths, and consequently a large number of laser arrays are needed, which presents intractable problems. To date, most reported FIR filters require wavelength stable lasers. While these lasers are commonly available, removing the need for temperature control can reduce the size, cost and complexity of the optical source.

In Section 3.3, a new reconfigurable FIR microwave photonic filter that uses an uncooled FP laser is demonstrated to provide a low-pass magnitude response [88]. The technique is based on the use of an LCoS-based FDOP to tailor the amplitude of the laser [41]. This is an efficient solution that can compensate for temperature fluctuations in the uncooled FP by monitoring the optical spectrum during operation, and avoids the requirement of using the expensive wavelength stable lasers.
3.2 Frequency Tracking Photonic Beamforming System Using Uncooled Fabry-Perot Lasers

In this section, a novel dynamic reconfigurable optical beamforming network that uses an uncooled Fabry-Perot laser as the optical source is presented. This is achieved by tracking the frequency and power drift of an FP laser and compensating these changes with programmable amplitude and phase in a liquid crystal-on-silicon (LCoS) Fourier domain optical processor. The system realizes six microwave phase shifters that operate simultaneously over 12GHz to 39.5GHz with amplitude and phase response variation of 6.1dB and 14.8° respectively, under the condition of changing laser current.

3.2.1 Principle of Operation

The topology of the frequency tracked photonic beamformer is illustrated in Fig.3.1. An array of uncooled lasers, either single- or multi-wavelength, could be used at the transmitter side. In this experiment, we elected to use FP lasers as they are multiwavelength, which increases the number of tones available from a single laser.

In order to avoid dispersion fading effects in the beamforming system, single sideband with carrier (SSB+C) modulation was generated based on a dual electrode Mach-Zehnder modulator, using an RF hybrid coupler to drive the two electrodes. The FP laser spectrum, with each tone modulated, is then passed through the FDOP, which is able to control amplitude and phase with a center frequency resolution of 1GHz. The FDOP is used to demultiplex each lasing line to a dedicated output port; additionally, the power of each tone is equalized, and a phase shift is imparted between the sideband and carrier. Thus, when each output port is directed to a photodetector, the resulting beat signal will be an RF signal with a frequency equal to our input RF frequency, but with a phase change that is set by the FDOP.

![Fig.3.1 Concept of the frequency tracked photonic beamformer, where n uncooled lasers are used to produce m spectral lines, each phase shifted, detected and emitted to produce a radiation pattern](image)

To enable the use of uncooled FP lasers, the optical spectrum is measured before the FDOP, and the measured frequency of the lasing lines was used to dynamically shift the filter profile accordingly. As environmental changes cause the source lasers to experience frequency drift, this feedback system is able to track every tone in real-time and ensure that the power of each beamforming taps is equalized, and the phase set by design. As the number of required taps increases, the system can be scaled by simply reconfiguring the software to handle the addition of more lasers.
The detected photocurrent at the nth output port is given by

\[ I_\Omega(t) \propto B_n P_n \cos(\omega_m t + \Psi_n) \quad (3.1) \]

where \( \Psi_n \) is the optical phase difference between the nth optical carrier with power \( P_n \) and its sideband at angular RF frequency \( \omega_m \). The amplitude coefficient, \( B_n \), is set by the relative attenuation in the FDOP. From this, we can estimate the far field radiation pattern, also called a space factor, given by

\[ F(\theta) = \sum_{n=0}^{N-1} B_n P_n \exp\left[j nk \Lambda (\sin \theta - \sin \theta_0)\right] \quad (3.2) \]

where \( N \) is the total number of radiating elements, \( k \) is the wavenumber, and \( \Lambda \) is the spacing between elements. The mainbeam, \( \theta_0 \), is related to the set of phase differences by

\[ \psi_n = -nk \Lambda \sin(\theta_0) \quad (3.3) \]

3.2.2 Driving Circuit of Fabry-Perot Laser

A Fabry-Perot cavity is a cavity with two reflecting mirrors as shown in Fig.3.2. The light is limited between the mirrors, forming a standing wave. Constructive interference occurs if the transmitted and reflected beams are in phase, resulting in a large transmission gain and high-transmission peak. In comparison, destructive interference is formed if the two beams are out of phase, resulting in a transmission minimum [80]. The type of interference depends on the wavelength of the optical signal and the refractive index of the material.

![Fig.3.2 Fabry-Perot cavity](image)

Fabry-Perot lasers use the Fabry-Perot cavity as the fundamental element, with its wavelength selected from gain region [81]. Multiple wavelengths can be formed in the cavity because the gain is wavelength dependent. In optical spectrum analysis, FP laser has a large free-spectral range, which is the wavelength or frequency spacing between adjacent lasing modes. The typical FSR for FP laser is 100-125GHz. The bandwidth of resonance is then given by the FSR divided by finesse. To improve the wavelength accuracy of the FP cavity, the distance between the two mirrors can be increased because more loops will be contained in the cavity. However, this would reduce the FSR. Also, the resonance frequency of the FP cavity can be tuned by adjusting the cavity length [82].

The statistical analysis for FP laser is analysed as follows. The mode field propagating in the forward direction in a Fabry-Perot cavity is given as
\[ E(x, y, z) = \{ \hat{x}E_x(x, y) + \hat{y}E_y(x, y) + \hat{z}E_z(x, y) \} e^{i\beta z} \] (3.4)

where \( E_x, E_y \) and \( E_z \) are the field in x, y and z direction respectively, \( z \) is the position along cavity, and \( \beta \) is the propagation constant. The mode generated corresponds to the propagation vector. For an optical cavity, suppose the amplitude reflection coefficients at the two reflecting surface is \( r_1 \) and \( r_2 \), and \( R_1 = |r_1|^2 \) and \( R_2 = |r_2|^2 \). The reflection coefficient in terms of amplitude and phase can be expressed as

\[ r_1 = \sqrt{R_1} e^{i\phi_1} \quad r_2 = \sqrt{R_2} e^{i\phi_2} \] (3.5)

For a certain cavity mode, the phase change of light in a roundtrip is an integral of \( 2\pi \), which is given by

\[ 2\beta L + \phi_1 + \phi_2 = 2p\pi \] (3.6)

where \( p \) is integer. The expression of adjacent modes is then given by

\[ 2\Delta\beta L + \Delta\phi_1 + \Delta\phi_2 = 2\pi \] (3.7)

The frequency spacing between adjacent modes can be obtained through dividing the propagation vector by delta omega, which is given by

\[ \left( 2 \frac{\partial\beta}{\partial\omega} L + \frac{\partial\phi_1}{\partial\omega} + \frac{\partial\phi_2}{\partial\omega} \right) \Delta\omega = 2\pi \]

\[ \Rightarrow \Delta\omega = \frac{\pi}{L} - \frac{1}{v_g} \left( \frac{\partial\phi_1}{\partial\omega} + \frac{\partial\phi_2}{\partial\omega} \right) \] (3.8)

The phase is usually weakly dependant on the frequency, thus for a long cavity, the Free Spectral Range is given by

\[ \Delta\omega = \frac{\pi c}{L n_g} \] (3.9)

The FSR is commonly expressed in the form of

\[ \Delta\lambda = \frac{\lambda^2}{2n_g L} \] (3.10)

The gain region of semiconductors is typically between 10-50nm, which generates many modes with this bandwidth as shown in Fig.3.3.

Due to the diode configuration of FP lasers, a driving circuit is needed to launch optical signals with the FP laser. A driving circuit designed for the uncooled FP laser used in the experiment is shown in Fig.3.4. A few parameters of the electronic components in the driving circuit is listed in Table.3.1. A pair of MOSFET current mirrors is used as the driving current supplier. Two voltage regulators are applied to control the driving current. The voltage regulators are composed of operational amplifiers and BJTs. The FP laser has an output power of 6dBm under a DC voltage supply of 7.62V.
Table 3.1 Design parameters of the FP laser driver

<table>
<thead>
<tr>
<th>Electronic component</th>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC supply voltage</td>
<td>$V_{cc}$</td>
<td>7.62V</td>
</tr>
<tr>
<td>Capacitor 1</td>
<td>$C_1$</td>
<td>0.1uF</td>
</tr>
<tr>
<td>Capacitor 2</td>
<td>$C_2$</td>
<td>0.1uF</td>
</tr>
<tr>
<td>Resistor 1</td>
<td>$R_1$</td>
<td>100 ohm</td>
</tr>
<tr>
<td>Adjustable resistor 1</td>
<td>$R_{adj1}$</td>
<td>0-1k ohm</td>
</tr>
<tr>
<td>Resistor 2</td>
<td>$R_2$</td>
<td>100ohm</td>
</tr>
<tr>
<td>Adjustable resistor 2</td>
<td>$R_{adj2}$</td>
<td>0-10k</td>
</tr>
<tr>
<td>Resistor 3</td>
<td>$R_3$</td>
<td>8.2 ohm</td>
</tr>
<tr>
<td>Capacitor 3</td>
<td>$C_3$</td>
<td>4.7uF</td>
</tr>
</tbody>
</table>

A typical laser output power against driving current under different temperature is shown in Fig. 3.5. The threshold current is significantly influenced by temperature, and it increases by over 20mA when temperature increases from 20° to 80°. Typically, laser threshold increases exponentially with temperature as $I_{th} = u_i \exp\left(\frac{T}{T_0}\right)$, where $u_i$ is the current coefficient, $T$ is the laser Kelvin temperature and $T_0$ is the characteristic temperature of the laser.

Due to the requirement of no temperature controller or external cavities, FP lasers have the advantages such as low cost and light weight over Distributed feedback (DFB) lasers. Furthermore, an FP laser launches multiple optical wavelengths simultaneously while a DFB laser only generates one optical wavelength. For a photonic beamforming system, each element requires one lasing mode. As the number of beamforming elements increases, the use of high performance tunable lasers necessarily increases the cost, size, weight and power consumption of the overall system. Therefore, a beamforming system using an uncooled FP laser is needed. The fluctuations of the FP laser caused by the environment temperature change can be solved by using a dynamic reconfigurable system with a Fourier-domain optical processor as its fundamental element.
3.2.3 Fourier-domain Optical Processor Simulation

The core component of the FDOP is a Liquid-crystal-on-silicon. LCoS is derived from Liquid Crystal and semiconductor technologies [41, 47]. It can be used to produce beam-steering by controlling the amplitude and phase of light at its each pixel with a high resolution. The schematic of LCoS-based FDOP is shown in Fig.3.6 [83]. In the FDOP, a large number of pixels are used to create optical switch. The light from a fiber array passes through a polarization diversity optics, where the polarization state of the light is aligned to be in the high-efficient state. The light is then reflected from the imaging mirror to the dispersion grating. The diffraction grating disperses the light and reflects the light to different
portions of the imaging mirror which is composed of LCoS pixel arrays [47]. The reflected light is distributed on the LCoS horizontally while overlapping vertical pixels. For the phase control, a phase scheme is applied on the horizontal axis, which can be programmable. The phase tuning range of LCoS is from 0 to $2\pi$. For the amplitude control, a second phase scheme is applied on the vertical axis. The modulated signal is then retraced to the fiber array from the LCoS, with each wavelength routed to the desired output port. Due to the independence of the channels, each wavelength can be directed to different ports without influencing other wavelengths. Therefore, the output from each port can be mapped to the desired photodetector, in which the optically phase shifted signals are converted to microwave signals.

Fig.3.6 LCoS-based Fourier-domain optical processor. The inset shows the incremental phase retardation on each pixel [83]

The simulation of different filters was carried out to investigate the characteristics and functions of the FDOP. The FDOP has a frequency setting resolution of ±1GHz (approximately 8pm), a filter bandwidth of 10GHz to 5THz, an attenuation conrol range from 0-35dB, and a bandwidth setting accuracy of ±5GHz. A 40Gbps DPSK filter centered at 194THz and a sinc function filter with a bandwidth of 200GHz were loaded on the FDOP as shown in Fig.3.7. The phase response of filters can also be controlled by changing the time delay of dispersion, since time delay is the frequency derivative of phase, and dispersion is the first-order frequency derivative of time delay.

Fig.3.7 Magnitude response of (a) 40Gbps DPSK filter and (b) Sinc function filter
3.2.4 Beam Pattern Analysis

Radiation pattern is the energy field radiated from an antenna. Each array element is controlled and adjusted individually and the combined elements form the whole radiation array. Based on Eq.3.2, MatLab simulation was carried out to study the beam characteristics such as the direction of mainlobe, the amplitude ratio of mainlobe and sidelobe, beamwidth and directivity. The element spacing $d$ is set as half wavelength, which means $d = \frac{\lambda}{2}$. The main beam angle $\theta_0$ is set at different values of $20^\circ$, $40^\circ$, $60^\circ$, $80^\circ$, $90^\circ$, $-20^\circ$ and $-40^\circ$. A full $360^\circ$ scanning range is covered in the simulation. As discussed in Section 2.4, optical beamforming can be obtained through time delay approach or phase shifting approach. For an optical beamforming network based on microwave photonic phase shifters, an important factor in simulation is to find the relationship between the mainbeam direction and the optical phase shift. In microwave photonic phase shifters, the frequency difference between the optical carrier and the sideband equals to the RF modulating frequency. The output signals at the EOM, which include the optical carrier and sideband, are launched into the Fourier-domain optical processor. The FDOP implements amplitude control and phase shift to the signals, so that a phase difference is created between the optical carrier and its sideband. In our proposed method, the phase difference is achieved through applying a phase shift to the optical carrier and keeping the phase of the sideband unchanged. The optically processed signals then enter the photodetector, where the optical carrier beats with its sideband, generating an RF signal. This generated RF signal has a phase which is equal to the phase difference between the optical carrier and its sideband. According to the phased array antenna theory, the mainbeam angle of the antenna element is related to the phase of the RF signal. The relationship between the mainbeam angle $\theta_0$ and the phase shift $\Psi_n$ is given by

$$\Psi_n = - nk \Lambda \sin(\theta_0)$$  \hspace{1cm} (3.11)

The mathematical expression of beamwidth is given in Eq.2.29. For simplicity, the 3dB beamwidth for a uniformly illuminated array can be approximated as [37]

$$\theta_{3\text{dB}} \approx \frac{5.57 c}{\omega Nd \cos(\theta_0)}$$  \hspace{1cm} (3.12)

The pointing angle error $\Delta \theta_0$ for small frequency deviations from the center frequency is expressed as

$$\Delta \theta_0 = \frac{\Delta \omega}{\omega} \tan(\theta_0)$$  \hspace{1cm} (3.13)

where $\Delta \omega$ is the small frequency deviation. The condition that the pointing error $\Delta \theta_0$ is less than half the beamwidth $\theta_{3\text{dB}}$ is fulfilled when

$$\Delta \omega < 0.886 \pi \frac{c}{Nd \sin(\theta_0)}$$  \hspace{1cm} (3.14)

The simulation of beam pattern of the phased array antenna with six radiating elements is conducted. The amplitude of each element is set as unity, and the phase shift applied on each element is calculated and summarized in Table.3.2. When the mainbeam angle is tuned to the opposite direction, the phase shift of each element is also adjusted to the opposite value. The simulation of beam pattern is shown in Fig.3.8.
<table>
<thead>
<tr>
<th>Main beam direction</th>
<th>Phase shift of the nth element</th>
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<tbody>
<tr>
<td></td>
<td>1st</td>
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<tr>
<td>20°</td>
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Table 3.2 Phase shift applied on each element under different beam directions

![Beam pattern with directions](image)

Fig. 3.8 Beam pattern with directions of (a) 20° (b) -20° (c) 40° (d) -40° (e) 60° (f) 80°
3.2.5 Experimental Results

A proof-of-concept experiment of the proposed beamforming network was carried out. The experimental setup is shown in Fig.3.9. A single coaxial FP laser was used, with a driving current of 30mA, resulting in an optical output power of +6dBm. No cooling was used to stabilize the temperature of the laser in the experiment. The power and frequency of laser modes were measured using a HR-OSA (WaveAnalyzer 1500S), where constant measurements of six lasing modes were taken over 30 minutes as shown in Fig.3.10.

![Fig.3.9 Experimental setup to demonstrate the frequency tracking photonic beamformer](image)

In our proposed system, each lasing mode is intensity modulated via SSB+C modulation. The FDOP (WaveShaper 4000S) allows us to dynamically set the phase on each individual lasing mode, by simply creating a phase shift between the carrier and sideband. The phase for each lasing mode is calculated based on the desired beam direction from Eq.2.31. Additionally, the power of each mode can be equalized by tailoring the attenuation at the carrier frequency. The measured spectrum of the FP laser after SSB+C modulation is shown in Fig.3.11, where the RF frequency was set to 20GHz. The lasing spectrum of the FP laser contains more than 13 modes that are measurable, but we select the six central tones with the most optical power.

![Fig.3.10 Measured (a) power and (b) frequency stability over 30 minutes, for each of the six FP lasing tones used for the beamformer.](image)
Using the uncooled laser reduces the power consumption of the transmitter, but makes the setup extremely vulnerable to environmental changes. If left free running, the laser would quickly drift outside of the FDOP filter passband, extinguishing all RF power at the photodetector. By using the high resolution measurement from the HR-OSA, we are able to track these changes and adjust the FDOP passband locations to suit. To measure the S21 magnitude and phase response of the system, a 43.5GHz Vector Network Analyzer (Agilent N5234A) was used to inject RF signals to the modulator and recover the response from a photodetector (u²t).

Fig.3.11 Lasing spectrum of the FP laser after SSB+C modulation with a 20GHz sinusoid; only the central six modes used in the beamformer are shown here.

To introduce significant ambient temperature variations on a short time scale, we used PCB cooling spray to reduce the laser package temperature from 24°C to 7°C, causing an optical frequency shift of 1400GHz. The lasing mode corresponding to the 180° phase shift was directed to the measurement setup, as the near pi phase shift is the most sensitive to perturbations. 177 consecutive measurements of amplitude and phase responses were acquired as the FP laser was changing frequency and power, as displayed in Fig.3.12. The mean response for amplitude and phase is shown in this figure, as well as the 2σ bound, corresponding to a 95% confidence interval.

The phase response of the system shows that the frequency tracking implementation is successful at maintaining the phase shift while the source is drifting. A region of large instability is observed from roughly 4-12GHz, which corresponds to the transition point between the carrier and the sideband regions in the FDOP. However, if we consider the stable region of 14 to 40GHz, we measure a phase shift of 182.2° ±14.8° over the 177 measurements. From Fig.3.12(a), we observed large fluctuations in the amplitude response, with a 2σ boundary of ±6.1dB, even in the stable phase response region of 14 to 40 GHz.

The amplitude fluctuations are attributed to several factors. First, FP lasers are known to have mode partition noise, which causes the power in each lasing line to fluctuate. These fluctuations occur much more rapidly than our compensation scheme can track, as the FDOP can only be reconfigured in roughly 300ms. Second, measuring narrow-linewidth signals in the HR-OSA at a refresh rate of 4 Hz shows large power uncertainty. Thus, the ability to track the power of the FP laser lines is inherently limited to the HR-OSA power repeatability. To accommodate this uncertainty, a single-coefficient Kalman filter was implemented per laser mode, to smooth the estimation of the power in each mode.
Fig. 3.12 Averaged measurements of (a) amplitude response and (b) phase response of a single phase shifting element, aggregating 177 consecutive measurements (dashed line indicates $2\sigma$ boundary).

Over the band of 12 to 39.5GHz, the phase response of six phase shifts at room temperature was measured, as shown in Fig. 3.13, where the traces indicate the average of 12 consecutive measurements. While the stability of five of the angles is reasonably good, with an average $2\sigma$ deviation of 2.9° (dotted lines), we note that the variance for the phase shift of 113.8° has a larger variance, which we found was a single measurement that had shifted over the passband in the FDOP. We attribute this to improper synchronization between the spectral measurements and the setting of the FDOP, and highlight a complication in the proposed system. Nevertheless, except for this outlier, we believe that this indicates that the frequency tracking beamformer operates consistently for the range of phase shifts required, and has good phase response stability.

Fig. 3.13 Aggregated phase responses of 12 consecutive S21 measurements, for each of the six angles created in the beamformer, calculated to form a mainbeam angle of 20° (dotted line indicate $2\sigma$ boundary).

To assess the suitability of the proposed photonic beamforming system, a photonic phased array that would hypothetically drive 6*Ku-band antennas was emulated. First, the mainbeam direction to 20° was
set, which from Eq.3.11, gives six RF phase shifts of $0^\circ$, $-61.6^\circ$, $-123.1^\circ$, $175.3^\circ$, $113.8^\circ$, and $52.2^\circ$. For simplicity, the power of each mode was set to be constant, which means that the FDOP is actively performing power equalization of the six modes. The measured variance in the S21 response was converted into variance in amplitude and phase setting for the antenna coefficients, and then the antenna radiation pattern was numerically synthesized. The emulated beam pattern for an array operating at 39GHz is shown in Fig.3.14(a), where the blue trace is the ideal far field pattern, and the dashed red trace is the emulated pattern.

To demonstrate the reconfigurability of the proposed system, the beam direction was changed by changing the phase shifts in the FDOP, and synthesizing new antenna patterns. The scanning angles of $40^\circ$, $-20^\circ$, and $-40^\circ$ was set, and the patterns are shown in Fig.3.14(b)-(d), respectively. We note that the fluctuations of the S21 response cause a widening of the main radiation lobe, and distort the sidelobes. These distortion effects could be mitigated by increasing the number of array elements. Our proposed beamforming system could be easily scaled to include more FP or DFB lasers, though the FDOP would need to have more output ports to accommodate the increased number of elements. High port count wavelength selective processors are commercially available in 1*20 configurations, and researchers have demonstrated devices that have 100 output ports.

![Synthesized antenna radiation patterns based on emulating a six-element antenna array directing the beam to (a) $20^\circ$, (b) $40^\circ$, (c) $-20^\circ$ and (d) $-40^\circ$. The ideal pattern is shown in blue, and the pattern calculated from experimental results is shown in red.](image)

### 3.2.6 Summary

In this section, a dynamic reconfigurable photonic beamforming system that uses an uncooled FP laser as the optical source is demonstrated. This is achieved by using rapid, high-resolution optical spectral measurements to track the frequency drift of the uncooled laser, and then reconfigure a programmable...
FDOP to compensate the changes caused by environmental changes. By using the uncooled FP laser, we eliminate the need for temperature control of the laser, and reduce the number of optical sources by using the spectral lines in the output optical spectrum of the laser. The system realizes six wideband microwave photonic phase shifters, and the resulting magnitude and phase responses vary within a 2σ deviation of 6.1 dB and 14.8°, respectively, even when the laser current is changed during the measurement.

3.3 Reconfigurable Microwave Photonic Filter using Uncooled Fabry-Perot Laser

In this section, a new reconfigurable microwave photonic filter that uses an uncooled Fabry-Perot laser as the optical source is presented. This filter realizes low-pass magnitude response and wide passband characteristic, by optically shaping the laser signal. This configuration can overcome the frequency and power drift of the uncooled FP laser by using a feedback structure with a fast update rate. Experimental results demonstrate a 6-tap microwave photonic filter with a free spectral range of 2.5GHz.

3.3.1 Principle of Operation

The schematic of the proposed microwave photonic filter is shown in Fig.3.15. An uncooled FP which launches multiple optical wavelengths is spectrally filtered by an LCoS-based FDOP, and is intensity modulated by an electro-optic modulator through double-sideband modulation. The modulated optical signals are then launched into a chirped fiber Bragg grating (CFBG) [90]. The CFBG offers linear time delay characteristic against frequency within its passband. The delayed optical signals are detected and summed at the photodetector, which converts the signals from optical domain back into electrical domain. The output signal of microwave photonic filter based on optical delay line is given by

\[ y(t) = \sum_{k=1}^{N} h_k x(t - k\tau) \] (3.15)

where \( x(t) \) is the input signal, \( N \) is the number of taps, \( k \) is the filter coefficient, \( h_k \) is the weight of the \( kth \) tap, and \( \tau \) is the basic system delay.

The optical delay line filter can be represented in the impulse response form due to its discrete and linear-time-invariant characteristics:
\[ h(t) = \sum_{k=1}^{N} h_k \delta(t - k\tau) \]  (3.16)

Thus the transfer function of the filter can be obtained through Fourier transform [89]:

\[ H(f) = \sum_{k=1}^{N} h_k e^{j2\pi fk} \cos(\pi df^2) \]  (3.17)

where \( f \) is the modulation frequency, \( df = D\lambda^2 / c \), \( \lambda \) is the optical wavelength, and \( D \) is the chromatic dispersion. The first item in Eq.3.17 is the transfer function of a standard FIR filter. The shape of the filter passband can be controlled through applying windowing functions. The second term is the RF decay, which is known as the carrier suppression effect. For the microwave photonic systems with delay elements and those work under double-sideband modulation, the carrier suppression effect is intrinsic [149].

Although this structure works in principle, it requires a highly stable laser which is under temperature control, increasing the complexity of the optical source. Any frequency or power drift can cause the optical signal to be offset from the FDOP, which can cause fluctuations in RF response. Therefore, in order to improve the stability of the system, the FDOP is dynamically reconfigured to compensate the power and frequency drift. A high-resolution optical spectrum analyzer is used as the feedback signal, with an update rate fast enough to track the signal fluctuations. The HR-OSA detects and tracks the optical signals, and pass the signal information to the FDOP. The FDOP reconfigures its filter settings when the feedback signal exceeds the offset boundary of the previous loop.

3.3.2 Microwave Photonic Filters Analysis

According to the signs of the taps, delay line filters can be classified as positive coefficient filters and complex coefficient filters [91-101]. Positive coefficient filters have the taps which add in intensity, and complex coefficient filters have the taps which are out-of-phase. The impulse response of a 10-tap positive coefficient filter and a 10-tap complex coefficient filter with unity tap weights is shown in Fig.3.16. The bipolar taps in complex coefficient filters can be generated in a dual-input EOM, in which its two branches create 180° phase difference.

**Fig.3.16 Tap coefficients of a (a) positive coefficient filter and (b) complex coefficient filter**

Complex coefficient filters have differences with positive coefficient filters in several ways. First, complex coefficient filters have a bandpass response in the frequency domain, while positive coefficient filters have a low-pass response in the frequency domain. Second, the center frequency of complex
coefficient filters can be tuned by adjusting the filter tap weights without influencing the filter shape. In comparison, the center frequency of positive coefficient filters can be tuned by changing the basic time delay.

The simulation of frequency response of positive and complex coefficient filters under different number of taps is shown in Fig.3.17. The basic time delay is selected as 199.2ps, and the tap weights are set at unity. For 2-tap, 3-tap, 4-tap and 6-tap filters, the impulse responses for positive coefficient filters are \( h_{p1} = [1 \ 1] \), \( h_{p2} = [1 \ 1 \ 1] \), \( h_{p3} = [1 \ 1 \ 1 \ 1] \) and \( h_{p4} = [1 \ 1 \ 1 \ 1 \ 1 \ 1] \), and those for complex coefficient filters are \( h_{c1} = [1 \ -1] \), \( h_{c2} = [1 \ -1 \ 1] \), \( h_{c3} = [1 \ -1 \ 1 -1] \) and \( h_{c4} = [1 \ -1 \ 1 -1 \ 1 -1] \). It can be seen from the figure that positive coefficients filters show low-pass response and complex coefficient filters show bandpass response. For unity tap weights, the FSR is only related to the basic time delay under different number of taps. The FSR, which is inversely proportional to the basic time delay, is 5.02GHz. The center frequencies of the mainlobes of positive coefficient filters are at 0, FSR, 2*FSR, 3*FSR and etc. The center frequencies of complex coefficient filters are at 0.5*FSR, 1.5*FSR, 2.5*FSR, 3.5*FSR and etc.

For optical delay line filters with unity taps, the time delay obtained is multiple integers of the basic time delay. Thus, the minimum time delay can be achieved is the basic time delay. However, a fraction of the basic time delay is also achievable in complex coefficient filters [84]. This can be obtained through adjusting the tap weights of the filter. The weight for each filter tap can be derived as follows. Assume
the sampling time of the signal is \( t = nT \), where \( n \) is integer and \( T \) is the sampling interval. The discrete time signal for processing is expressed as

\[
y(n) = L \{ x(n) \} = x(n-D) \tag{3.18}
\]

where \( D \) is the time delay. For a fractional delay line filter, \( D \) is composed of a positive integer part and a fractional part as \( D = \text{Int}(D) + d \). The time delay is often presented in the form of phase in frequency response and is defined as the negative frequency derivative of the phase [100]. The ideal impulse response of the fractional delay line filter is obtained through the inverse discrete-time Fourier transform and is given as

\[
h_{id}(n) = \frac{\sin[\pi(n-D)]}{\pi(n-D)} = \sin c(n-D) \quad \text{for all } n \tag{3.19}
\]

It can be seen that fractional delay line filters require a sampled and shifted sinc function. Fig.3.18 shows the simulation of a 10-tap fractional delay line filter with \( D = 4.1, 4.2, 4.3 \) and 4.4. The tap weights of fractional delay line show both positive and negative values. In experiment, the positive taps which operating at different optical frequencies are sent to the first input port of the EOM, and the other negative taps are sent to the second port of the EOM. The output of the EOM is launched into a dispersive medium such as a chirped FBG, which provides frequency-dependent linear time delay, and detected by a photodiode.

Fig.3.18 The impulse response of a 10-tap fractional delay line filter with \( D = (a) 4.1, (b) 4.2, (c) 4.3 \) and (d) 4.4
The simulation of the group delay of the fractional delay line filter is shown in Fig. 3.19. The number of filter taps is 8 and the basic system delay $\tau$ is selected as 20ps, which results in a FSR of 50GHz. To minimize the error, the integer part of the fractional delay is placed at the center of the impulse response. This means that for an 8-tap fractional delay line filter, the overall delay is selected at 4. Fig. 3.19 shows the group delay with $D=\{4, 4.1, 4.2, \ldots, 4.9\}$. The group delay under $D=4$ is set as offset. It is clear that the group delay with a given $D$ is a fractional part of the basic system delay. For example, the group delay under $D=4.2$ accounts for 0.2 of the basic system delay, resulting in a 4ps delay.

![Fig.3.19 Group delay of a fractional delay line filter with $D=\{4, 4.1, 4.2, \ldots, 4.9\}$](image)

### 3.3.3 Optical Delay Medium Theory

The optical delay medium applied in optical delay line filters can be classified mainly into two types: optical fibers and fiber gratings. For optical fibers, multimode fibers were implemented in previous works, which limited the system performance. Later, single-mode fibers (SMF) became the dominant optical delay element due to its wideband signal processing functions and low loss [9]. As optical delay medium, single-mode fibers possess wavelength dependent group delay characteristic. The induced group delay is given by

$$\Delta t = \Delta \lambda \ DL$$

where $D$ is the dispersion (ps/nm), $L$ is the length of fiber, and $\Delta \lambda$ is the wavelength spacing between two optical signals. For an n-tap FIR filter, the time delay can be adjusted by changing the wavelength spacing between optical taps.

Light spreads out during the transmission in fiber, which is known as chromatic dispersion. The total dispersion of an SMP includes the material dispersion as well as the waveguide dispersion [53]. The total dispersion increases along optical wavelengths, with the zero-dispersion around 1.31um. Note that the minimum fiber loss region is around 1.55um. Thus, the wavelength of no dispersion was designed at 1.55um for dispersion shifted fibers. The dispersion curve of standard, dispersion flattened and dispersion shifted fibers are shown in Fig. 3.20. Therefore, through tailored dispersion shifting, the loss of optical fibers at 1.55um is lower than other wavelengths. By implementing optical amplifiers, the signal transmission distance can exceed several thousand kilometres.
A limitation of optical fiber delay lines is the RF degradation caused by the dispersion of the fiber. For example, when an optical carrier is modulated by an RF signal, the generated sidebands experience different time delay in the optical fiber. Destructive interference occurs during the beating between sidebands at the photodetector, resulting in RF degradation. Single-sideband modulation overcomes the dispersion-induced RF degradation, enhancing the RF bandwidth.

Fiber gratings are applied in broad areas such as optical sensing, beamforming and frequency mixing. They possess the advantages of all-fiber geometry and low loss. The fiber Bragg grating is one of the most widely used fiber gratings. The FBG is built in a short optical fiber, with the core refractive index varying periodically. The FBG reflects certain optical wavelengths and transmits all others. The behavior of fiber Bragg gratings can be explained with the coupled-mode theory [33]. The resultant perturbation of the refractive index $n_{\text{eff}}$ can be expressed as [102]

$$
\delta n_{\text{eff}}(z) = \overline{\delta n_{\text{eff}}}(z) \left[ 1 + v \cos \left( \frac{2\pi}{\Lambda} z + \phi(z) \right) \right]
$$

(3.21)

where $\overline{\delta n_{\text{eff}}}(z)$ is the index change offset, $v$ is the fringe visibility, $\Lambda$ is the grating period, $z$ is the location along the fiber core, and $\phi(z)$ is the chirp rate. According to the index change, fiber Bragg gratings can be categorized into two types: uniform gratings and non-uniform gratings [150].
The uniform Bragg grating has a constant index change and zero grating chirp. Fig. 3.21 shows the boundary conditions of a uniform Bragg grating, where \( A(z) \) is the amplitude of propagating mode and \( B(z) \) is the amplitude of counter-propagating mode [102]. The light propagation in FBG needs to meet the boundary conditions, which means that the light only transmits in one direction without back-propagating wave for \( z \geq \frac{L}{2} \), making \( A^+\left(-\frac{L}{2}\right) = 1 \) and \( B^+\left(\frac{L}{2}\right) = 0 \). With the chirp \( \frac{d\phi}{dz} = 0 \), the amplitude reflection coefficient \( \rho \) can be expressed as [150]

\[
\rho = \frac{-\kappa \sinh(\sqrt{\kappa^2 - \hat{\sigma}^2} L)}{\hat{\sigma} \sinh(\sqrt{\kappa^2 - \hat{\sigma}^2} L) + i\sqrt{\kappa^2 - \hat{\sigma}^2} \cosh(\sqrt{\kappa^2 - \hat{\sigma}^2} L)}
\] (3.22)

where \( \kappa \) is the coupling coefficient offset and \( \hat{\sigma} \) is the self-coupling coefficient [103]. The amplitude reflectivity reaches maximum when \( \hat{\sigma} = 0 \). This also results in the maximum wavelength, which is given by

\[
\lambda_{\text{max}} = \left(1 + \frac{\hat{\sigma}_{\text{eff}}}{n_{\text{eff}}} \right) \lambda_0
\] (3.23)

The time delay and dispersion of the uniform Bragg grating can be obtained from the amplitude reflection coefficient \( \rho \). The time delay is the frequency derivative of the phase of \( \rho \), which is given by [102, 103]

\[
\tau_p = \frac{d\psi_p}{d\omega} = -\frac{\lambda^2}{2\pi c} \frac{d\psi_p}{d\lambda}
\] (3.24)

The dispersion is the frequency derivative of the time delay:

\[
D_p = \frac{d\tau_p}{d\lambda} = \frac{\lambda^2}{\lambda} \frac{d^2\psi_p}{d\lambda^2}
\]

\[
= -\frac{2\pi c}{\lambda^2} \frac{d^2\psi_p}{d\omega^2}
\] (3.25)

Chirped fiber Bragg grating is one of the non-uniform Bragg gratings, which as a varying grating period. Due to the periodic change of the refractive index, only one wavelength is reflected with its reflectivity reaching maximum. This reflected Bragg wavelength is given by [25, 29]

\[
\lambda_\beta = 2n_{\text{eff}} \Lambda
\] (3.26)

where \( n_{\text{eff}} \) is the effective refractive index, and \( \Lambda \) is the grating period. It is clear from Eq. 3.28 that the desired Bragg grating can be obtained from adjusting the refractive index and the grating period [30]. In fabrication, chirped FBGs can be converted from uniform gratings through changing the refractive index and the grating period [90]. Fig. 3.22 shows the structure of a chirped fibre Bragg grating.
Chirped FBGs are applied in broad areas such as optical filtering and sensing. Therefore, the operating bandwidth of CFBG is an important parameter, which is given by

$$\Delta \lambda_{chirp} = 2n_{eff} (\Lambda_{long} - \Lambda_{short}) = 2n_{eff} \Delta \Lambda_{eff}$$  \hspace{1cm} (3.27)

3.3.4 Experimental Results

A proof-of-concept system is established using an uncooled FP laser, driven by a current of 30mA, producing an optical output power of +6dBm. The measured laser spectrum is shown in Fig.3.23. The power and frequency of the laser are altered by tuning the driving current. Using the HR-OSA, the peak frequency and power of the FP laser is monitored over 40 minutes. While the frequency of the lasing lines changed equally among the modes, the mode competition in the FP laser caused lines to suddenly disappear. For example, the change over time for the peak at 1551.19nm is shown in Fig.3.24, with a frequency drift of $\pm 7GHz$ and power fluctuations of $\pm 3.4dB$.

Using this laser, a microwave photonic filter based on Fig.3.15 is set up using a double-sideband EOM (from EOSpace), and a high speed photodetector. The FDOP was configured to choose six optical carriers and their sidebands, and filtered out the lower sideband through amplitude attenuation. The optical spectrum was monitored on the HR-OSA, and as the detected signal was offset from FDOP, the FDOP was reconfigured to stay aligned to the laser, as well as equalize the power of each peak, with an update rate of approximately 5 seconds.
The modulated optical signals were then launched into a linear chirped fiber Bragg grating with a dispersion of 300ps/nm. The optical signals were all in the wavelength reflection range of the CFBG. The wavelength spacing was 1.33nm, with a corresponding FSR of 2.5GHz. A Keysight Vector Network Analyzer was used to measure the transfer function of the microwave photonic filter, and we acquired consecutive measurements over 20 minutes. The experimental magnitude responses measured from 10MHz to 11GHz are shown in Fig.3.25. The power fluctuation of the first order passband is $\pm 1dB$ over 20 minutes.

![Graph showing frequency drift and power fluctuation](image)

**Fig.3.24** Frequency drift and power fluctuation of the lasing peak at 1551.19nm

![Graph showing magnitude response](image)

**Fig.3.25** Measured magnitude response of the microwave photonic filter at 5min, 10min, 15min, 20min.
3.3.4 Summary

In this section, a new reconfigurable microwave photonic filter that uses an uncooled FP laser as the optical source and an FDOP as the programmable amplitude and phase controller is presented. An HR-OSA is used to provide feedback signal to the system by monitoring the frequency drift and power fluctuation of the optical signals with an update rate fast enough to track the changes. Experimental results demonstrate a 6-tap microwave photonic filter that shows a low-pass magnitude response, with an FSR of 2.5GHz. The power fluctuation of the first-order passband is ±1dB over 20 minutes.

3.4 Conclusion

In the field of microwave photonics, there has been growing attention to adding dynamic reconfigurability to optical beamforming networks and microwave photonic filters. This chapter has reported a new technique based on feedback loop and dynamic control, which can be used to improve the environmental adaptability of optical beamforming networks and microwave photonic filters. The dynamic reconfigurability is achieved through the use of a Fourier-domain optical processor, which implements amplitude tailoring and phase shift to optical signals. A high-resolution optical spectrum analyser is applied to monitor the frequency drift and power fluctuation of optical signals, thus providing feedback information to the system. Under this dynamic reconfigurability, an uncooled Fabry-Perot laser is used as the optical source, which reduces the cost and size of the microwave photonic system. The proposed system has an update rate fast enough to track the system variation under environmental changes.

In Section 3.2, a photonic beamforming system that uses an uncooled FP laser as the optical source is demonstrated. The system realizes six wideband microwave photonic phase shifters, and the resulting magnitude and phase responses vary within a $2\sigma$ deviation of 6.1 dB and 14.8°, respectively, even when the laser current is changed during the measurement. In Section 3.3, a microwave photonic filter that uses an uncooled FP laser as the optical source and an FDOP as the programmable amplitude and phase controller is presented. Experimental results demonstrate a 6-tap microwave photonic filter that shows low-pass magnitude response, with an FSR of 2.5GHz. The power fluctuation of the first-order passband is ±1dB over 20 minutes.
4.1 Introduction

Microwave signal mixing is an important characteristic for microwave photonic systems [104]. The signal mixing in fiber optic links can be achieved through electrical domain or optical domain. Electrical domain approaches mix the received RF frequency with the local oscillator after photodetection, while optical domain approaches mixing the RF signal with LO signals before photodetection [104-132]. Electrical domain approaches are limited by the signal processing speed and bandwidth. Moreover, the requirement of a high-speed photodetector adds nonlinearity to the system, limiting the conversion efficiency and the dynamic range of the system. In comparison, optical domain approaches avoid these limitations.

Previously, different structures have been proposed for microwave photonic mixers, including the structures based on cascaded intensity modulators [103], cross-gain modulation in a semiconductor optical amplifier [127], optical Four-wave mixing [112], and phase coherent orthogonal optical carriers [133]. However, these approaches either have low conversion efficiency and large mixing spurs generated by useless sidebands, suffer the disability of frequency conversion for phase modulated signals, or limit the lowest possible frequency of the local oscillator. Electro-optic modulators were widely used for microwave photonic frequency conversion. Approaches such as cascaded or parallel electro-optic modulators have been proposed to achieve frequency mixing [156, 157]. However, the requirement of bias voltage control limits the operation bandwidth of the system. Also, the suppression of intermodulation distortions is achieved through additional digital signal processing after photodetection [156], which requires the use of analog-to-digital convertors, thus limiting the signal processing speed of the system.

Recently, there is growing interest in using phase modulators for microwave photonic frequency conversion. Compared with electro-optic modulators, phase modulators avoid the need of bias voltage control, and have lower insertion loss and the capability of handling higher input power. Single-mode fiber was used to provide bandpass filtering function to the phase modulated signals [158]. However, this system lacks the ability of tuning the center frequency due to the fixed length of the SMF. A serial phase modulation approach with optical notch filtering was proposed in [159]. This system has limited frequency selectivity and only performs frequency downconversion instead of upconversion due to the limited 3dB-bandwidth of the FBG.

In this chapter, a novel tunable all-optical microwave photonic mixer is presented based on serial phase modulation and on-chip notch filter. The notch filter breaks the π-phase difference between the symmetrical sidebands generated from phase modulation, resulting in bandpass response of frequency selection. This system is achieved through an all-optical approach, which does not require electrical components, thus increasing the operation bandwidth of the system. The tunability of frequency selection is achieved through adjusting the wavelength of the optical source. Experimental verify the technique with a 3rd-order SFDR of 91.7dBm/Hz^2/3.

The content of this chapter is as follows. First, the principle of the proposed microwave photonic mixer is demonstrated in Section 4.2.1. Next, the performance of microwave photonic mixers based on different modulation schemes is investigated. The optical spectrum analysis after modulation is given in Section 4.2.2, and the electrical spectrum analysis after photodetection is given in Section 4.2.3. The conversion efficiency and noise performance is presented in Section 4.2.4. The VPI simulation of SFDR is given in Section 4.2.5. Finally, the experimental results are demonstrated in Section 4.2.6.
4.2 Tunable All-optical Microwave Photonic Mixer Based on Serial Phase Modulation and On-chip Notch Filter

4.2.1 Principle of Operation

The schematic diagram of the proposed microwave photonic mixer is shown in Fig.4.1. The system consists of a tunable laser source, polarization controllers PC1 and PC2, serial connected phase shifters PM1 and PM2, a tunable on-chip notch filter and a photodetector. The optical signal from the laser source is launched to the first phase modulator after polarization control, modulated by the first RF signal. The modulated optical signals of the PM1 then enter the PM2, modulated by the second RF signal. The output signals at the serial phase modulators are then fed into the on-chip notch filter. Finally, the output signals at the notch filter are detected by the photodiode, where optical signals are converted to corresponding RF signals.

The optical field at the output of the serial phase modulators under small signal modulation can be expressed as

\[
E_{\text{out}} = E_0 \left( \sum_{n=-\infty}^{\infty} \sum_{k=-\infty}^{\infty} J_n(\alpha)J_k(\beta) \cdot \cos[(\omega_0 + n\omega_1 + k\omega_2)t + \frac{1}{2}n\pi + \frac{1}{2}k\pi] \right) \quad (4.1)
\]

where \( E_0 \) is the optical field of the laser source, \( \omega_0 \), \( \omega_1 \) and \( \omega_2 \) are the angular frequencies of the optical carrier, RF1 and RF2 respectively. \( n \) and \( k \) represent the order of harmonics, and \( J_n(\alpha) \) and \( J_k(\beta) \) are the \( nth \) and \( kth \) order Bessel functions of the first kind respectively, with \( \alpha \) and \( \beta \) indicating the phase modulation indices in which \( \alpha = \pi \frac{V_1}{V_{x_1}} \) and \( \beta = \pi \frac{V_2}{V_{x_2}} \), and \( V_{x_1} \) and \( V_{x_2} \) are the half-wave voltages of the PM1 and PM2 respectively.

Phase modulation produces an output with 180° difference between the upper and lower sidebands. Generally, if the phase modulated signals are sent directly to the photodetector, there will be no RF signal detected at the output of the photodetector, because the beating of the upper sideband and the optical carrier cancels that of the optical carrier and the lower sideband. However, the existence of the on-chip notch filter breaks the balance between the out-of-phase symmetry. The principle of operation is shown in Fig.4.1. When the frequency of optical carrier \( \omega_0 \) does not overlap with the center frequency of the optical notch filter \( \omega_c \), the sideband which falls within the bandstop region will not be symmetrically passed through the filter as its out-of-phase counterpart. Therefore, the signal located at the notch region
will remain after photodetection, resulting in bandpass filtering of frequency selection. By simply changing the wavelength of the laser source or the center frequency of the on-chip notch filter, different frequency components of the mixing products can be selected. The transfer function of the microwave photonic mixer can be expressed as

\[ H(\omega_m) = |H(\omega_0 - \omega_m)|e^{i[\phi(\omega_0 - \omega_m)]} - |H(\omega_0 + \omega_m)|e^{i[\phi(\omega_0 + \omega_m)]} \]  

(4.2)

where \( |H(\omega)| \) and \( e^{i\omega} \) represent the amplitude and phase response of the on-chip notch filter. \( \omega_m \) is the modulation frequency in which \( \omega_m = n\omega_1 + k\omega_2 \).

### 4.2.2 Modulation Spectrum Analysis

The mixing of RF signals requires multiple modulators or a single modulator with hybrid couplers. The idea behind the proposed microwave photonic mixer is to use a notch filter to break the out-of-phase symmetry between the symmetrical sidebands generated from phase modulation. Thus, it is desired to know how the combination of different modulation schemes with phase modulation influences the frequency mixing. The spectrum analysis of microwave photonic mixers under different modulation schemes is given in this section. These modulation schemes include:

- Serial phase modulation
- Single phase modulation
- Single-sideband modulation + phase modulation (SSB+PM)
- Double-sideband modulation + phase modulation (DSB+PM)
- Single-sideband suppressed carrier modulation + phase modulation (SSB-SC+PM)
- Double sideband suppressed carrier modulation + phase modulation (DSB-SC+PM)

The spectrum analysis under each modulation scheme is divided into two parts: the optical spectrum at the output of modulators, and the RF spectrum at the output of photodetector. This section focuses on the optical spectrum at the output of modulators. The details of each modulation scheme are expanded as follows.

#### 4.2.2.1 Serial phase modulation

The schematic of the serial phase modulation is shown in Fig.4.2. The input optical signal is modulated by an IF signal in the first modulator PM1. The output signals from PM1 enter the second modulator PM2, modulated by an LO signal. Note that the output signals from PM1 all work as optical carriers in the PM2. The PM1 output optical field is described as follows.
where $E_{in}$ is the optical field of the input optical signal to PM1, $l_i$ is the insertion loss of PM1, $\omega_0$ is the optical signal frequency, $\omega_{IF}$ is the modulating IF signal frequency, $\alpha$ is the modulation index of the IF signal. Bessel function expansion is used on Eq.4.3 to observe the frequency components. By expanding Eq.4.3 in terms of Bessel function of the first kind, $E_i(t)$ can be expressed as

$$E_i(t) = E_{in}I_i \exp(j\omega_0 t) \exp(j \alpha \cos \omega_{IF} t)$$  

(4.3)

In Eq.4.4, only first-order sidebands are considered due to the small signal modulation. However, theoretically, there are infinite sidebands after modulation. The PM2 output optical field is given

$$E_2(t) = E_i(t)l_2 \exp(j \beta \cos \omega_{LO} t)$$  

(4.5)

Similarly, only first-order sidebands are considered in Eq.4.6. It is clear from Eq.4.6 that besides the IF frequency and LO frequency, the intermodulation terms at $\omega_0 + \omega_{IF} + \omega_{LO}$, $\omega_0 + \omega_{IF} - \omega_{LO}$, $\omega_0 - \omega_{IF} + \omega_{LO}$ and $\omega_0 - \omega_{IF} - \omega_{LO}$ are also generated from serial phase modulation. These terms are known as second-order intermodulation terms (IMD2), which result from the nonlinearity of phase modulators. Higher order intermodulation terms exist in theory, which can be derived from expanding Eq.4.3 and Eq.4.5 with Bessel function to higher orders.

Note that the sequence of the modulating signals entering the phase modulators does not influence the modulation result. In other words, the modulation output is still the same as Eq.4.6 when the LO signal enter PM1 and the IF signal enter the PM2.

The VPI simulation of the optical spectrum after serial phase modulation is shown in Fig.4.3. The input optical signal is centered at 193.5THz, with an output power of 10dBm. The IF signal and LO signals are
at 11GHz and 9GHz respectively. These settings are set as the same in the other simulations in this section. The sidebands generated from serial phase modulation are at ±2GHz, ±9GHz, ±11GHz, ±18GHz, ±20GHz, and ±22GHz relative to the optical carrier frequency.

![Fig.4.3 Optical spectrum at the output of modulators under serial phase modulation](image)

### 4.2.2.2 Single phase modulation

The schematic of the single PM modulation is shown in Fig.4.4. The modulating IF signal and LO signal are combined with an RF coupler. The electrical field of the coupled modulating signals can be expressed as 

\[ E(t) = l_{\text{coupler}} \left( \exp(j\omega_{\text{IF}}t) + j\beta_{\text{LO}}t \right), \]

where \( l_{\text{coupler}} \) is the loss of RF coupler.

![Fig.4.4 Single phase modulation scheme](image)

The field function of the optical signals after being phase modulated by the coupled microwave signals is given by
\[ E_i(t) = E_i l_{pm} l_{coupler} \exp(j \omega_i t) \exp(j \alpha \cos \omega_{RF} t + j \beta \cos \omega_{LO} t) \]

\[ = E_i l_{pm} l_{coupler} \exp(j \omega_i t) \sum_{m=-\infty}^{\infty} \sum_{n=-\infty}^{\infty} J_m(\alpha) J_n(\beta) \exp[j(m \omega_{RF} + n \omega_{LO}) t] \]

\[ = E_i l_{pm} l_{coupler} \{ J_0(\alpha) J_0(\beta) \exp(j \omega_i t) + J_0(\beta) J_1(\alpha) \exp[j(\omega_i + \omega_{RF}) t + \frac{\pi}{2}] \]

\[ - J_0(\alpha) J_1(\beta) \exp[j(\omega_i - \omega_{LO}) t - \frac{\pi}{2}] + J_0(\alpha) J_1(\beta) \exp[j(\omega_i + \omega_{RF} + \omega_{LO}) t + \pi] \]

\[ - J_1(\alpha) J_1(\beta) \exp[j(\omega_i + \omega_{RF} - \omega_{LO}) t] - J_1(\alpha) J_1(\beta) \exp[j(\omega_i - \omega_{RF} + \omega_{LO}) t] \]

\[ + J_1(\alpha) J_1(\beta) \exp[j(\omega_i - \omega_{RF} - \omega_{LO}) t - \pi] \]

Comparing Eq.4.6 and Eq.4.7, we can see that the frequency components generated from serial phase modulation and single phase modulation with RF coupler are the same, except that the sidebands of the later one have lower power due to the loss of the RF coupler. The VPI simulation of the single PM case is shown in Fig.4.5.

![Fig.4.5 Optical spectrum at the output of modulators under single phase modulation scheme with RF coupler](image)

4.2.2.3 SSB+PM modulation

Another option for microwave photonic mixers involves the use of intensity modulators. Fig.4.6 shows the schematic of EOM+PM modulation. A bias voltage is needed for the EOM to control the sideband suppression ratio [119, 135]. Intensity modulation can be classified into the schemes of SSB, SSB-SC, DSB and DSB-SC.

![Fig.4.6 EOM+PM modulation scheme](image)
The optical field at the output of the EOM is given by

\[ E_i(t) = E_0^l \text{eom} \left\{ J_0(\alpha) \exp(j\omega_0 t) + J_1(\alpha) \exp[j(\omega_0 + \omega_{IF})t] \right\} \] (4.8)

Where \( E_0 \) is the amplitude of the input optical signal, \( l_{\text{eom}} \) is the insertion loss of the EOM, \( \omega_0 \) is the angular frequency of the input signal, \( J_n(\alpha) \) is the \( nth \) order Bessel function of the first kind and \( \alpha \) is the RF modulation index. The output optical signals at EOM, which include the optical carrier and the single sideband, are then modulated by an LO signal in the phase modulator. The field function of the optical signals at the output of the PM is expressed as

\[
E_z(t) = E_i(t) l_{\text{pm}} \exp(j \beta \cos \omega_{LO} t)
= E_0^l \text{eom} l_{\text{pm}} \left\{ \left[ J_0(\alpha) \exp(j\omega_0 t) + J_1(\alpha) \exp[j(\omega_0 + \omega_{IF})t] \right] 
- J_0(\alpha) J_1(\beta) \exp(j\omega_0 t + \frac{\pi}{2}) - J_1(\alpha) J_0(\beta) \exp[j(\omega_0 + \omega_{IF})t - \frac{\pi}{2}] 
+ J_0(\alpha) J_1(\beta) \exp[j(\omega_0 + \omega_{IF})t + \frac{\pi}{2}] - J_1(\alpha) J_0(\beta) \exp[j(\omega_0 - \omega_{LO})t - \frac{\pi}{2}] \right\} 
\] (4.9)

The VPI simulation of the SSB+PM modulation is shown in Fig.4.7. It can be seen that the upper sidebands have more frequency components than the lower sidebands due to the single-sideband modulation of EOM.

![Fig.4.7 Optical spectrum at the output of modulators under SSB+PM modulation](image)

4.2.2.4 SSB-SC+PM modulation

The optical carrier is suppressed under the SSB-SC modulation in the EOM. The optical field at EOM output is given by

\[ E_i(t) = E_0^l \text{eom} J_1(\alpha) \exp[j(\omega_0 + \omega_{IF})t] \] (4.10)
The single sideband at $\omega_0 + \omega_{IF}$ works as the optical carrier in the phase modulator. The optical field at the output of the PM is given by

$$E_2(t) = E_1(t) I_{pm} \exp(j \beta \cos \omega_{LO} t)$$

$$= E_1 I_{pm} I_1(\beta) \{ J_1(\beta) \exp[j(\omega_0 + \omega_{IF}) t] + J_1(\beta) \exp[j(\omega_0 + \omega_{IF} + \omega_{LO}) t + \frac{\pi}{2}] - J_1(\beta) \exp[j(\omega_0 + \omega_{IF} - \omega_{LO}) t - \frac{\pi}{2}] \}$$

(4.11)

It can be observed from Eq.4.11 that SSB-SC+PM essentially works as a phase modulator. It does not provide frequency mixing function. Therefore, SSB-SC+PM scheme cannot be used in microwave photonic mixers for frequency conversion. The simulation result is shown in Fig.4.8.

4.2.2.5 DSB+PM modulation

The field function of the EOM output under double-sideband modulation is given by [135]

$$E_1(t) = E_1 I_{con} \{ J_0(\alpha) \exp(j \omega_0 t) + J_1(\alpha) \exp[j(\omega_0 + \omega_{IF}) t] + J_1(\alpha) \exp[j(\omega_0 - \omega_{IF}) t] \}$$

(4.12)

Consequently, the optical field at the output of the PM is given by

$$E_2(t) = E_1(t) I_{pm} \exp(j \beta \cos \omega_{LO} t)$$

$$= E_1 I_{pm} I_1(\beta) \{ J_0(\beta) \exp[j(\omega_0 t) + J_1(\beta) \exp[j(\omega_0 + \omega_{IF}) t] + J_1(\beta) \exp[j(\omega_0 - \omega_{IF}) t] \}$$

$$\times \{ J_0(\beta) + J_1(\beta) \exp(j \omega_{LO} t + \frac{\pi}{2}) - J_1(\beta) \exp(-j \omega_{LO} t - \frac{\pi}{2}) \}$$

$$= E_1 I_{pm} I_1(\beta) J_0(\beta) \exp(j \omega_0 t) + J_0(\beta) J_1(\alpha) \exp[j(\omega_0 + \omega_{IF}) t]$$
The VPI simulation of the optical spectrum at EOM output under DSB+PM modulation is shown in Fig.4.9. It can be seen that the sidebands generated from DSB+PM are the same as serial phase modulation, except that the sideband suppression ratio is different, which can be adjusted through changing the RF input power or the modulation index of the EOM.

\begin{equation}
\begin{aligned}
+J_i(\beta)J_i(\alpha)\exp[j(\omega_b - \omega_{IF})t] + J_o(\alpha)J_o(\beta)\exp[j(\omega_b + \omega_{LO})t + \frac{\pi}{2}] \\
-J_i(\alpha)J_i(\beta)\exp[j(\omega_b - \omega_{LO})t - \frac{\pi}{2}] + J_i(\alpha)J_i(\beta)\exp[j(\omega_b + \omega_{IF} + \omega_{LO})t + \frac{\pi}{2}] \\
-J_i(\alpha)J_i(\beta)\exp[j(\omega_b + \omega_{IF} - \omega_{LO})t - \frac{\pi}{2}] + J_i(\alpha)J_i(\beta)\exp[j(\omega_b - \omega_{IF} + \omega_{LO})t + \frac{\pi}{2}] \\
+J_i(\alpha)J_i(\beta)\exp[j(\omega_b - \omega_{IF} - \omega_{LO})t - \frac{\pi}{2}]
\end{aligned}
\end{equation}

The field function of the optical signals at the output of the EOM is given by

\begin{equation}
E_i(t) = E_{0i}\exp\{J_i(\alpha)\exp[j(\omega_b + \omega_{IF})t] + J_i(\alpha)\exp[j(\omega_b - \omega_{IF})t]\}
\end{equation}

The field function of the optical signals at the output of the EOM is expressed as

\begin{equation}
E_2(t) = E_i(t)\exp(j\beta \cos \omega_{LO}t)
= E_{0i}\exp\{J_i(\alpha)\exp[j(\omega_b + \omega_{IF})t] + J_i(\alpha)\exp[j(\omega_b - \omega_{IF})t]\}
\times[J_0(\beta) + J_i(\beta)\exp(j\omega_{LO}t + \frac{\pi}{2}) - J_i(\beta)\exp(-j\omega_{LO}t - \frac{\pi}{2})]
\end{equation}
The VPI simulation of the DSB-SC+PM is shown in Fig.4.10. The optical carrier at $\omega_0$ is suppressed, and the first RF modulating signal generated from EOM intensity modulation work as the optical carriers in the PM, in which sidebands are generated with frequency spacing equal to the second RF frequency.

\[
E_{\text{out}}(\nu) = J_0(\beta)J_1(\alpha)\exp[j(\omega_0 + \omega_{RF})t] + J_0(\beta)J_1(\alpha)\exp[j(\omega_0 - \omega_{RF})t] + J_1(\alpha)J_1(\beta)\exp[j(\omega_0 + \omega_{RF} + \omega_{LO})t + \frac{\pi}{2}] - J_1(\alpha)J_1(\beta)\exp[j(\omega_0 + \omega_{RF} - \omega_{LO})t - \frac{\pi}{2}] + J_1(\alpha)J_1(\beta)\exp[j(\omega_0 - \omega_{RF} + \omega_{LO})t + \frac{\pi}{2}] + J_1(\alpha)J_1(\beta)\exp[j(\omega_0 - \omega_{RF} - \omega_{LO})t - \frac{\pi}{2}] 
\]

\(4.15\)

To summarize, for EOM+PM structures, SSB+PM, DSB+PM and DSB-SC+PM all provide frequency mixing functions, while SSB-SC+PM does not providing mixing function. For all-PM structures, the single PM structure requires the use of RF couplers, which results in limited operation bandwidth in electrical domain. In comparison, the PM+PM structure does not require RF coupler or bias voltage, thus providing the potentiality of achieving a microwave photonic mixer based on all-optical approach.

### 4.2.3 RF Spectrum Analysis

This section focuses on the electrical spectrum analysis after photodetection as shown in Fig.4.11. The aim of this section is to investigate how the notch filtering influences the final RF output. Mathematical expressions under different modulation schemes are derived. The center frequency of the notch filter is denoted as $\omega_c$ and the bandwidth of the notch filter as B.
4.2.3.1 Serial phase modulation

Under PM+PM scheme, no RF signal will be detected after photodetection when no filtering function is applied. By adding an optical notch filter, the out-of-phase balance between the sidebands is broken, resulting in bandpass response in electrical domain. The center frequency of the bandpass response is equal to the frequency difference between the optical carrier and the notch center.

Assume an optical notch filter centered at $\omega_i + \omega_r + \omega_{LO}$ is applied to the phase modulated signals, the mathematical expression of the output optical spectrum at filter can be obtained simply by subtracting Eq.4.6 with the notch term:

$$E_r(t) = E_i(t) - E_m J_1 J_1(\alpha) J_1(\beta) \exp\{j(\omega_i + \omega_r + \omega_{LO})t + \pi\}$$

The photocurrent is proportional to the detected optical power at photodetector. In mathematical form, the photocurrent is proportional to the product of the optical spectrum and its complex conjugate:

$$i(t) \propto E_r(t) E_r(t)^*$$

$$\approx J_1^2(\alpha) J_1^2(\beta) \exp\{j(\omega_r + \omega_{LO})t\} - J_1(\alpha) J_1(\beta) \exp\{j\omega_{LO}t + \frac{\pi}{2}\}$$

$$+ J_1(\alpha) J_1^2(\beta) \exp\{j(2\omega_r + \omega_{LO})t + \frac{\pi}{2}\} - J_1(\alpha) J_1^2(\beta) \exp\{j(\omega_r t + \frac{\pi}{2}\}$$

Since $J_1^2(x)$ is much smaller than $J_0^2(x)$ under small signal modulation, the terms with coefficient $J_1^4(x)$ are dropped in Eq.4.17. It can be seen that the notch filtering function generates a few components at $\omega_r + \omega_{LO}$, $\omega_r + 2\omega_{LO}$, $2\omega_r + \omega_{LO}$, $2\omega_r + 2\omega_{LO}$. By observing the amplitude coefficient of each frequency term in Eq.4.17, the dominant RF frequency component is at $\omega_r + \omega_{LO}$, which is generated from the beating between the optical carrier at $\omega_i$ an the sideband at $\omega_i - \omega_r - \omega_{LO}$. The simulation of the electrical spectrum at the output of photodetector is shown in Fig.4.12. The power of the dominant frequency component is approximately 17dB higher than that of other new generated frequency components. The power ratio can be adjusted by changing the modulation index. The influence of these new generated sidebands to the RF dominant component will be described with SFDR, which will be discussed in Section 4.2.4.
4.2.3.2 Single phase modulation

Since the modulation output of the PM+PM and single PM scheme are the same, the RF output after notch filtering of the two schemes are expected to be identical. The VPI simulation is shown in Fig.4.13.

Fig. 4.13 Output electrical spectrum at photodetector after notch filtering under single PM scheme

4.2.3.3 SSB+PM modulation

The photocurrent generated from single-sideband modulation without filtering consists of the modulating frequency and its higher order harmonics, which can be expressed as

\[ i(t) \propto \sum_{n=1}^{\infty} I_n \cos(n\omega_{IF}t) \]  \hspace{1cm} (4.18)

where \( n \) is the order of harmonics and \( I_n \) is the photocurrent of the \( nth \) order harmonics. When no filtering function is applied to the SSB+PM modulated signals, the RF output only consists of the SSB modulating frequency and its harmonics as shown in Fig.4.14 (a). In the simulation, the IF signal and LO signal are 11GHz and 9GHz respectively.

A notch at the \( \omega_R + \omega_{IF} + \omega_{LO} \) is applied to the SSB+PM signals, and the RF output is shown in Fig.4.14(b). The new generated RF terms include \( \omega_{IF} + \omega_{LO} \), \( \omega_{LO} \), \( 2\omega_{LO} \), \( \omega_{IF} + 2\omega_{LO} \) and \( \omega_{IF} - \omega_{LO} \). Among these terms, \( \omega_{IF} + \omega_{LO} \) is the fundamental component, and \( \omega_{IF} - \omega_{LO} \) is the IMD2. Note that \( 2\omega_{LO} \) and \( 2\omega_{IF} \) also limit the dynamic range of the mixer. Also, the IF power is higher than the fundamental component, thus limiting the performance of the mixer.

Fig.4.14 RF spectrum at the output of PD under SSB+PM (a) without optical filtering (b) after notch filtering
4.2.3.4 DSB+PM modulation

The VPI simulation of the DSB+PM scheme is shown in Fig. 4.1. Comparing Fig.4.14 with Fig.4.15, the recovered RF frequency components after notch filtering are the same under DSB+PM as SSB+PM. However, the second-order harmonics at $2\omega_{rf}$ under DSB+PM is lower than that under SSB+PM. This gives DSB+PM larger dynamic range than SSB+PM.

Fig.4.15 RF spectrum at the output of PD under DSB+PM after notch filtering

4.2.4 Conversion Efficiency and Noise

This section focuses on the conversion efficiency and noise of the proposed microwave photonic mixer based on serial phase modulation and on-chip notch filter. As given in Chapter 2, the conversion efficiency of optical fiber links is the power ratio of the RF output to the RF input. The power of the RF output can be derived from the photocurrent. The detected DC photocurrent at the photodetector is given by

$$I_{dc} = \Re P_{0} I_{pm, total} g_{0}$$

(4.19)

where $P_{0}$ is the optical input power, $I_{pm, total}$ is the total insertion loss of the modulation module. For the serial phase modulation structure, $I_{pm, total} = I_{pm1} I_{pm2} \cdot g_{0}$ is the net optical gain or loss between the modulation module and photodetector. In mathematical form, $g_{0}$ is the product of the insertion loss of the on-chip notch filter and the gain of amplifiers. $\Re$ is the responsivity of the photodetector. Among all the mixing products, the photocurrent of the fundamental/wanted component is given by

$$I_{\Omega} = I_{dc} \cdot a$$

(4.20)

where $a$ is the coefficient of the fundamental component. The output power of the fundamental component can be obtained from Ohm’s Law:

$$P_{\Omega} = I_{\Omega}^2 R_{L} |H_{pd}|^2$$

(4.21)

where $R_{L}$ is the load resistance, and $|H_{pd}|$ is the filter function of the photodetector. Thus, the system conversion efficiency can be obtained from Eq.2.3:
\[ g = \frac{P}{P_{in}} = \frac{I_{dc} e^{2} \alpha^{2} R_{l} |H_{pd}|^{2}}{P_{in}} \]  \hspace{1cm} (4.22)

where \( P_{in} \) is the input RF power. Note that for microwave photonic mixers, \( P_{in} \) is the input IF power. The input voltage of IF signal can be given as \( V_{IF} = \sqrt{2P_{in} R_{m}} \), where \( R_{m} \) is the input resistance of modulator. The filter function of the PD \( |H_{pd}| \) is often taken as 1. For microwave photonic mixers, the wanted RF component is the mixing product of the two modulating signals. Thus, when the desired frequency component is at \( \omega_{IF} + \omega_{LO} \), the conversion efficiency as be obtained as follows:

\[ G_{conv} = I_{p1}^{2} I_{p2}^{2} J_{1}(m_{LO})^{2} P_{in}^{2} \left( \frac{\pi}{V_{g}} \right)^{2} \Re \beta_{1} R_{l} R_{m} \]  \hspace{1cm} (4.23)

In the equation above, \( g_{0} \) is set as 1 based on the assumption that the gain of EDFA compensates the loss of the notch filter. Fig.4.16 shows the MatLab simulation of the conversion efficiency. The parameters used in the simulation are shown in Table.4.1. It can be seen that the conversion efficiency is a linear function of the input optical power. The conversion efficiency is -20dB at an input optical power of 10dBm.

![Conversion efficiency of the microwave photonic mixer](image)

**Fig.4.16 Conversion efficiency of the microwave photonic mixer based on serial phase modulation**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Name</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>( q )</td>
<td>Electron Charge</td>
<td>( 1.6 \times 10^{-19} ) C</td>
</tr>
<tr>
<td>( K )</td>
<td>Boltzmann Constant</td>
<td>( 1.38 \times 10^{-23} ) J / K</td>
</tr>
<tr>
<td>( R_{m} )</td>
<td>Input Resistance</td>
<td>50Ω</td>
</tr>
<tr>
<td>( R_{l} )</td>
<td>Load Resistance</td>
<td>50Ω</td>
</tr>
<tr>
<td>( m_{LO} )</td>
<td>LO Modulation Index</td>
<td>0.1</td>
</tr>
<tr>
<td>( I_{p1} )</td>
<td>PM1 Insertion Loss</td>
<td>4.02dB</td>
</tr>
<tr>
<td>( I_{p2} )</td>
<td>PM2 Insertion Loss</td>
<td>7.3dB</td>
</tr>
<tr>
<td>( \Re \beta )</td>
<td>Photodetector Responsivity</td>
<td>0.6A / W</td>
</tr>
<tr>
<td>( T )</td>
<td>Absolute Temperature</td>
<td>298K</td>
</tr>
<tr>
<td>( RIN )</td>
<td>Relative Intensity Noise</td>
<td>( 1 \times 10^{-16} ) Hz(^{-1})</td>
</tr>
</tbody>
</table>

**Table.4.1 Parameters for the microwave photonic mixer simulation**
The simulation of noise sources is shown in Fig.4.1. The thermal noise is constant over the laser input power under room temperature. The shot noise is a linear function to the laser input power, while the RIN noise is a parabolic function to the laser input power. It can be seen that the shot noise and RIN noise are -170dBm and -180dBm under the input power of 10dBm. The total noise floor relies more on the shot noise and RIN noise with the increase of optical input power.

Fig.4.18 shows the simulation of the Noise Figure. The NF is 25dBm under an input of 10dBm. The practical noise figure is expected to be higher than the simulation due to the use of notch filter and optical amplifiers.
4.2.5 Spurious-free Dynamic Range Simulation

The VPI simulation of SFDR is demonstrated in this section [117, 136]. The front panels of the VPI simulation under serial phase modulators and single phase modulator is shown in Fig.4.19. The optical signal from the laser source is at 193.5THz with an output power of 10dBm. The two modulating RF signals are at 11GHz and 9GHz respectively. Each phase modulator has an insertion loss of 3dB. The RF coupler in the single PM structure has a loss of 5dB. The phase modulated signals are amplified by 10dB, then enter the optical notch filter. The notch filter has 20dB insertion loss and a notch depth of 14dB. The center frequency of the notch filter is adjustable. The filtered signals are detected by a photodetector, which is connected with a 2-tone analyzer.

Fig.4.19 VPI front panel of the microwave photonic mixer based on (a) serial phase modulation and (b) single phase modulation

Fig.4.20 shows the VPI simulation of the modulation output. The suppression ratio between the optical carrier and the first-order sideband under serial phase modulation scheme is 5dB higher than the single PM scheme, which is caused by the RF coupler loss.
The center frequency of the optical notch filter is set at 193.511THz. The RF spectrum after photodetection is shown in Fig.4.21. The fundamental component is at 11GHz, which is introduced by the notch filtering. The limiting distortion is the IMD3 term, which is the closest to the fundamental component with a frequency spacing of 2GHz. In practice, when the bandwidth of the filter is not narrow enough, the IMD3 cannot be filtered out.

To obtain the 3rd-order SFDR of the system, the power of the fundamental component and IMD3 is measured at different RF input power. The numerical data of the VPI simulation is shown in Table.4.2. The noise floor is set at -150dBm/Hz. The analytical SFDR is obtained by extracting the numerical data to the noise floor. Fig.4.22 shows that the microwave photonic mixer based on serial phase modulation has a 3rd-order SFDR of 95.5dBm/Hz, and the one based on single phase modulation has a 3rd-order SFDR of 80dBm/Hz. Thus, it can be concluded that the single PM scheme not only limits the operating bandwidth of the system, but also limits the SFDR.
The VPI front panel for the microwave photonic mixer under EOM+PM scheme is shown in Fig.4.23. The laser output signal is at 193.5THz and 10dBm. The first RF modulating signal driving the EOM is at 11GHz and the second RF modulating signal driving the PM is at 9GHz. Double-sideband modulation is applied to the EOM with a modulation index of 0.3. To investigate the 2\textsuperscript{nd}-order SFDR of the system, the notch frequency is set at 193.520GHz with a notch depth of 14dBm. The simulation of the 2\textsuperscript{nd}-order SFDR is shown in Fig.4.24. The grey dots and brown dots are the numerical data for the fundamental component and the IMD2, and the grey and brown lines are the analytical lines for the SFDR extrapolation. The simulation shows a 2\textsuperscript{nd}-order SFDR of 43dBm/Hz\textsuperscript{1/2} for the microwave photonic mixer based on EOM+PM.
<table>
<thead>
<tr>
<th>RF input [dBm]</th>
<th>Fundamental component [dBm]</th>
<th>IMD3 [dBm]</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Single PM</td>
<td>Serial PM</td>
</tr>
<tr>
<td>-17.276</td>
<td>-53.664</td>
<td>-55.504</td>
</tr>
<tr>
<td>-16.948</td>
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<tr>
<td>-16.327</td>
<td>-52.808</td>
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</tr>
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<td>-16.033</td>
<td>-52.547</td>
<td>-54.668</td>
</tr>
<tr>
<td>-15.748</td>
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<td>-15.472</td>
<td>-52.058</td>
<td>-54.334</td>
</tr>
<tr>
<td>-15.205</td>
<td>-51.827</td>
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</tr>
<tr>
<td>-14.945</td>
<td>-51.605</td>
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</tr>
<tr>
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</tr>
<tr>
<td>-14.211</td>
<td>-50.992</td>
<td>-53.708</td>
</tr>
<tr>
<td>-13.979</td>
<td>-50.803</td>
<td>-53.615</td>
</tr>
</tbody>
</table>

Table 4.2 VPI numerical data of the RF output power for the microwave photonic mixer based on serial phase modulation and single phase modulation.

![VPI front panel for the microwave photonic mixer based on EOM+PM and notch filter](image1)

Fig. 4.23 VPI front panel for the microwave photonic mixer based on EOM+PM and notch filter.

![2nd-order SFDR of the microwave photonic mixer based on EOM+PM and notch filter](image2)

Fig. 4.24 2nd-order SFDR of the microwave photonic mixer based on EOM+PM and notch filter.
4.2.6 Experimental Results

A proof-of-concept experiment of the proposed microwave photonic mixer based on serial phase modulation and notch filter was carried out. A contrast experiment for the EOM+PM experiment was also carried out to compare with the serial PM structure.

The experimental setup of the microwave photonic mixer based on EOM+PM is shown in Fig.4.25. The tunable laser source launches an optical signal at 1559.4227nm with an output power of 10dBm. The optical signal enters the EOM through polarization control, modulated by the first RF signal through double-sideband modulation. RF₁ has a frequency of 11GHz and a power of -1dBm. The power ratio between the optical carrier and the sidebands is adjusted through changing the DC voltage of the EOM. The output signals of the EOM enter the PM, modulated by the second RF signal. RF₂ has a frequency of 9GHz and a power of 5dBm. The output signals after EOM+PM modulation then enter the notch filter. The notch filter employed in the system is based on a microring resonator fabricated on a SOI wafer via ePIXfab, with its center frequency adjustable through temperature control. The filtered optical signals then enter the photodetector, which is connected with an electrical spectrum analyzer.

The optical spectrum at the output of EOM+PM was measured with a Wave Analyzer as shown in Fig.4.26. The sidebands generated from modulation include the modulating signals (RF₁, RF₂), mixing products (RF₁+RF₂, RF₁- RF₂), and second-order harmonics (2RF₁ and 2RF₂). Compared with the simulation result shown in Fig.4.8, it can be seen that harmonics appear in the experiment. This is because the RF power applied in the experiment is much larger than that in simulation. In the simulation, both modulating signals are small signals, thus the second-order harmonics very small and buried under noise floor.
Next, the optical spectrum after notch filtering was measured. The notch center was set at 192.359THz through temperature control. Fig.4.27 shows the notch filter response and the optical spectrum after notch filtering. The on-chip microring resonator shows a notch depth of 12dB. Compared with Fig.4.26, the sideband at the notch center is suppressed down, while other frequency components remain the same.

The output RF spectrum after photodetection was measured with ESA as shown in Fig.4.28. The frequency component at 20GHz, which is the mixing product of RF₁ and RF₂, is selected from notch filtering as the fundamental component. The IMD2 term at 29GHz also appears due to the nonlinearity of modulators. The EOM modulating signal RF₁ is also recovered from photodetection, which cannot be eliminated. Other frequency components detected at ESA include RF₁-RF₂, RF₂, 2RF₁ and 2RF₂.
As shown in Fig.4.28, the dynamic range is mainly limited by the 2\textsuperscript{nd}-order harmonics, as they are the closest to the fundamental component. The dynamic range is obtained through extrapolating the power measurement of the fundamental component and the 2\textsuperscript{nd}-order harmonics. As shown in Fig.4.29, the dynamic range is reduced to 40dBm/Hz and 45dBm/Hz caused by 2RF\textsubscript{1} and 2RF\textsubscript{2}.

![Graph showing dynamic range](image)

**Fig.4.29** The limitation of the dynamic range caused by the 2\textsuperscript{nd}-order harmonics (a) 2RF\textsubscript{1} and (b) 2RF\textsubscript{2} for the microwave photonic mixer based on EOM+PM and notch filter

The experimental setup of the microwave photonic mixer based on serial phase modulation is shown Fig.4.30. The tunable laser source launches an optical signal at 1559.4427nm with an output power of 11.5dBm. The optical signal enters the first phase modulator through polarization control, modulated by the first RF signal from the RF signal generator. The frequency of RF\textsubscript{1} is 11GHz, and the insertion loss of the PM1 is 2.3dB. The output signals of the PM1 enter the PM2, modulated by the second RF signal which is generated from another RF source. RF\textsubscript{2} has a frequency of 9GHz and the PM2 has an insertion loss of 3.4dB. The power of the two RF modulating signals is kept the same in the experiment. The output signals after serial phase modulation are polarization controlled and amplified by an EDFA, then enter the notch filter. The notch filter employed in the system is based on a microring resonator fabricated on a
SOI wafer via ePIXfab, with its center frequency adjustable through temperature control. The filtered optical signals then enter the photodetector, which is connected with an electrical spectrum analyzer.

Fig.4.30 Experimental setup of the microwave photonic mixer based on serial phase modulation

The optical spectrum at the output of the serial phase modulators under different RF power was measured with a Wave Analyzer as shown in Fig.4.31. Only the first order sidebands at 9GHz and 11GHz are detectable at low RF modulating power. With the increase of RF power, higher order harmonics and intermodulation sidebands also appear at the output of serial phase modulators. Under the RF power of -10dBm, the power ratio between carrier and the first order sideband is 32dBm, while this reduces to 12dBm under the RF power of 10dBm.

Next, the optical spectrum after notch filtering was measured. The center frequency of the notch filter was set at 192.365THz, with the resistance of its temperature controller adjusted to 10.36kΩ. Fig.4.32
shows the notch filter response and the optical spectrum after notch filtering. The on-chip microring resonator shows a notch depth of 12dB. Compared with Fig.4.31(d), the sideband at 11GHz is suppressed by 12dB. Note that the sideband at 9GHz is also suppressed by approximately 5dB because it falls into the notch region.

![Fig.4.32 Optical spectrum after notch filtering (blue) and notch filter response (green), serial phase modulation scheme](image)

The output RF spectrum after photodetection was measured with ESA as shown in Fig.4.33. The frequency component at 11GHz was selected as the fundamental component. It can be seen that the fundamental component under serial phase modulation has the largest power, which is different from the EOM+PM scheme. The IMD3 term at 13GHz also appears due to the nonlinearity of modulators. Other frequency components include RF1-RF2, 2RF2-RF1, RF2, 2RF2, RF1+RF2 and 2RF1.

![Fig.4.33 RF output spectrum measured at ESA after photodetection, serial PM scheme](image)
To measure the 3rd-order SFDR of the system, the RF input power was adjusted from 3.5dBm to 10dBm. The power of the fundamental component and the IMD3 component was measured with the ESA as shown in Fig.4.34. The noise floor was measured as -142.35dBm/Hz. The analytical power of the RF components was extracted from the measurement data. It can be seen that the microwave photonic mixer based on serial phase modulation and notch filtering has a 3rd-order SFDR of 91.7dBm/Hz$^{2/3}$. The RF component at 9GHz would also limit the dynamic range of the system. However, this component can be eliminated by using notch filters with smaller bandwidth.

![Fig.4.34 Measured 3rd-order SFDR of the microwave photonic mixer based on serial phase modulation and notch filtering](image)

4.3 Conclusion

In the field of microwave photonics, there has been growing attention to realize microwave photonic frequency mixers based on all-optical scheme, since it provides higher signal processing speed and larger operating bandwidth. This chapter has reported a microwave photonic mixer based on serial phase modulation and optical notch filtering. Each RF modulating signal is fed into a single phase modulator without using RF couplers. The optical notch filter breaks the out-of-phase symmetry between the sidebands generated from phase modulation, resulting in bandpass response in RF domain. The fundamental component of the mixing products can be selected by either changing the optical wavelength or the notch frequency of the optical filter.

Microwave photonic mixers based on other modulation schemes are also investigated. Compared with serial phase modulation structure, other modulation schemes all need to use electrical components, either to control the bias voltage of the modulator, or to combine the RF modulating signals. Furthermore, EOM+PM structure results in additional RF component that cannot be eliminated, thus influencing the dynamic range of the system. The single PM structure has lower SFDR than the serial PM structure.

The proposed microwave photonic mixer based on serial phase modulation and notch filtering is a route to realizing frequency mixing of multiple signals. Additional signals can be mixed by adding phase modulators to the serial structure. Experimental results verify the proposed technique with a measured 3rd-order SFDR of 91.7dBm/Hz$^{2/3}$. 


CONCLUSION

5.1 Summary

Microwave photonic signal processing refers to the use of photonic techniques to manipulate microwave signals. Photonic signal processing overcomes the electronic bottlenecks, and offers the advantages such as large instantaneous bandwidth, low loss, high speed, compact size and immunity to electromagnetic interference. They provide the possibility of meeting the increasing demand for high-speed signals and large bandwidth systems.

This thesis has investigated microwave photonic signal processing techniques with dynamic reconfigurability, which either solves the problems or improves the system performance. The achievements of the work completed in this thesis are summarized below.

- A photonic beamforming system that uses an uncooled FP laser as the optical source is demonstrated. This is achieved by using rapid, high-resolution optical spectral measurements to track the frequency drift of the uncooled laser, and then reconfigure a programmable FDOP as compensation. By using an uncooled laser, we eliminate the need for temperature control of the laser, and reduce the number of optical sources by using the spectral lines in the output optical spectrum of the laser. The system realizes six wideband microwave photonic phase shifters, and the resulting magnitude and phase responses vary within a $2\sigma$ deviation of 6.1 dB and 14.8°, respectively, even when the laser current is changed during the measurement.

- A microwave photonic filter that uses an uncooled FP laser as the optical source and an FDOP as the programmable amplitude and phase controller is presented. An HR-OSA is used to provide feedback signal to the system by monitoring the frequency drift and power fluctuation of the optical signals with an update rate fast enough to track the changes. Experimental results demonstrate a 6-tap microwave photonic filter that shows low-pass magnitude response, with an FSR of 2.5GHz. The power fluctuation of the first-order passband is $\pm 1\text{dB}$ over 20 minutes.

- A novel tunable all-optical microwave photonic mixer is presented based on serial phase modulation and on-chip notch filter. The notch filter breaks the out-of-phase balance between the upper and lower sidebands generated from phase modulation, resulting in bandpass response of frequency selection. This system is achieved through an all-optical approach, which does not require electrical components, thus increasing the operation bandwidth of the system. The tunability of frequency selection is achieved through adjusting the wavelength of the optical source. Experimental verify the technique with a 3rd-order SFDR of 91.7dBm/Hz$^{2/3}$.

5.2 Future Work

Although microwave photonic signal processing techniques with dynamic reconfigurability have been presented, there are a number of ways in which the present work can be extended. The optical beamforming network reported in Chapter 3 achieves high processing speed and avoids the use of a large number of stable lasers. However, the setting resolution of the system is limited by the number of pixels. Increasing the number of pixels would enhance the setting resolution, however, this would also result in larger size and complexity of the system. Thus, the recommended work is to explore different techniques to develop optical beamforming networks. Also, the system update rate is expected to increase, which would adapt better with environmental changes.
The microwave photonic filter based on FDOP and feedback loop reported in Chapter 3 achieves low-pass magnitude response. It is expected to realize a bandpass filter, which requires complex coefficient taps. This requires a higher phase setting resolution of the FDOP, since an out-of-phase needs to be added to half of the filter taps. It is desired that the FDOP could process more taps to increase the resolution of the filter.

Microwave photonic mixers based on different modulation schemes are demonstrated in Chapter 4. For the EOM+PM structure, it is expected to eliminate the unwanted frequency component which has larger magnitude than the wanted frequency component. This can be solved by adding an electrical filter, however, it would limit the signal processing speed and the operating bandwidth. For the serial phase modulation structure, the recommended work is to improve the dynamic range of the system. This can be achieved by suppressing the intermodulation components, which can be realized by adjusting the polarization between sidebands.

Microwave photonic signal processing is a unique technology that brings new opportunities to revolutionize the microwave field and photonics field. It is hoped that this thesis can provide some insight into the design and analysis of microwave photonic signal processing, and encourage further research in this field.
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