Microwave signal processing using photonic techniques

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**NANYANG
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**MICROWAVE SIGNAL PROCESSING USING
PHOTONIC TECHNIQUES**

ZHOU JUNQIANG

**SCHOOL OF ELECTRICAL AND ELECTRONIC
ENGINEERING**

2011

Microwave Signal Processing Using Photonic Techniques

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ABSTRACT

Microwave signal processing using photonic technologies is a technique to process microwave or radio frequency (RF) signals with the help of photonic devices or subsystems. Processing of RF signals conveyed by an optical carrier directly in the optical domain offers great flexibility in selecting the radio frequency of operation, RF bandwidth, and the filter response. This technique can overcome the limitations of conventional electrical signal processors such as limited bandwidth and electromagnetic interference. It also has the advantages of high rate-distance product, low loss, and tunable and adaptive functions. The major parts of the thesis are devoted to the development of the microwave signal processing techniques such as microwave photonic filter (MPF) and microwave signal instantaneous frequency measurement (IFM) using photonic techniques.

In order to realize microwave photonic filter (MPF), various schemes of division, delay and summing of a modulated optical signal have been proposed in the literature. The coherent summing of optical signals is sensitive to polarization fluctuation caused by environmental perturbation. Therefore the incoherent approach is more attractive in practice for stable operation. To satisfy the requirement of incoherent summing, either the laser coherence time is needed to be shorter than the optical delay time, or the state of polarization (SOP) of the optical signals after division needs to be orthogonal.

Tunability is an important feature of microwave photonic filters. Based on the working principles of microwave photonic filters, tunability can be realized through changing the number of taps, tap weights, and time delay between taps. Consequently,

the first major aspect of this thesis targets the development of different types of incoherent MPFs with tunability including the MPFs based on chirped fibre Bragg grating (FBG), the MPFs based on multi-wavelength laser and RDL type infinite impulse response (IIR) MPF.

The second major aspect of this thesis focuses on the instantaneous frequency measurement of microwave signals using photonic means. For many radar and other electronic warfare (EW) systems, instantaneous frequency measurement (IFM) of a microwave signal is required to enable scanning, identification and analysis of the microwave signal over a large frequency range with a high probability of interception. In these systems, a number of specialized receivers are jointly employed to reduce the processing load of a single receiver. Therefore, the carrier frequency of a microwave signal is needed to be measured instantaneously using an IFM receiver before passing it to a specialized receiver for further processing. The conventional electronic solutions for realizing such an IFM receiver generates two out-of-phase signals and calculates the frequency through the comparison of these two signals. However, because of the limited bandwidth of the microwave components, with this approach it is difficult to meet the wide bandwidth requirement of the modern EW environment. Therefore, in the literature, photonic approaches have been proposed including optical scanning receiver, optical channelizer receiver, and frequency-to-time mapping and frequency-to-power mapping based IFM receivers. For the measurement of frequency of a single-frequency signal, IFM techniques based on frequency-to-power mapping are attractive because of the large frequency measurement range and high measurement resolution. Combined with a channelizer receiver, the single frequency limitation for an IFM receiver is possible to be overcome. Hence this thesis mainly focuses on the frequency-to-power mapping based IFM receivers. Keeping in mind

the requirements of practical applications, we have tried to realize IFM receiver designs that have reduced number of components, wide measurement range, and high resolution.

In all cases, the performance has been predicted theoretically and measured results that match well the theoretical ones have been obtained. The thesis concludes by summarizing the major achievements and suggestions for future work.

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ACRONYMS

A

ACF	Amplitude Comparison Function
ASE	Amplified Spontaneous Emission

C

CD	Chromatic Dispersion
CW	Continuous Wave

D

DC	Direct-Current
DCF	Dispersion Compensation Fibre
DD-MZM	Dual-Drive Mach-Zehnder Modulator
DFB	Distributed Feedback
DGD	Differential Group Delay
DSB	Double Sideband

E

EAM	Electro-Absorption Modulator
EDF	Erbium-Doped Fibre
EDFA	Erbium-Doped Fibre Amplifier
EOM	Electro-optic Modulator
EW	Electronic Warfare

F

FBG	Fibre Bragg Grating
FIR	Finite Impulse Response
FRDL	Forward Recirculating Delay Line
FSR	Free Spectral Range
FWHM	Full Width Half Maximum
FWM	Four-Wave-Mixing

H

Hi-Bi	High-Birefringence
HNDSF	Highly-Nonlinear Dispersion Shifted Fibre
HNLF	Highly-Nonlinear Fibre

I

IIR	Infinite Impulse Response
IFM	Instantaneous Frequency Measurement

L

LCFBG	Linearly Chirped Fibre Bragg Grating
LD	Laser Diode

M

MSSR	Main to Secondary Sidelobe Ratio
MWP	Microwave photonics
MZM	Mach-Zehnder Modulator

N

NLCFBG Nonlinearly Chirped Fibre Bragg Grating

P

PBS Polarization Beam Splitter

PC Polarization Controller

PD Photodetector

PM Polarization Maintaining

PMF Polarization Maintaining Fibre

R

RDL Recirculating Delay Line

RF Radio Frequency

RRDL Reflective Recirculating Delay Line

S

SMF Single-mode Fibre

SOA Semiconductor Optical Amplifiers

SOP State of Polarization

T

TLS Tunable Laser Source

V

VNA Vector Network Analyzer

Chapter 1 INTRODUCTION

1.1 Background and motivation

Microwave photonics (MWP) consists of techniques to process microwave or radio frequency (RF) signals with the help of photonic devices or subsystems [1]. The original intention in MWP is to introduce photonic technology into microwave systems for eliminating the so-called electronic bottleneck and, at the same time, obtaining the advantages from photonics such as high rate-distance product, immunity to electromagnetic interference, tunability, low loss, flat frequency responses etc. These advantages result in many important functions which are complex or even impossible to be realized directly in electrical or microwave domain, and have been used in diverse applications, such as signal processing, instantaneous frequency measurement (IFM), radio-over-fibre systems, optical beam-forming, photonic analog-to-digital conversion, and arbitrary waveform generation [2, 3], etc.

Considering first the topic of signal processing, the traditional RF and microwave filters represent a class of electronic filters working over a certain frequency range, usually in the megahertz to gigahertz range, for broadcast radio, wireless communication, and radar systems, etc. The electronic filters usually have a bandwidth constraint; these also suffer from electromagnetic interference and high loss especially when the signal frequency is high. On the other hand, processing of RF signals conveyed by an optical carrier directly in the optical domain can overcome these limitations and offers great flexibility in selecting the radio frequency of

operation, RF bandwidth, filter response etc. and provides tunable and adaptive functions [4].

The concept of using an optical fibre as a delay medium for RF signal processing was first proposed by K. Wilner and A. P. Van Den Heuvel in 1976 [5]. C. T. Chang, J. A. Cassaboom and H. F. Taylor realized a bandpass filter using an optical fibre in the following year [6]. These works opened a new era for realizing high-resolution broadband processing of microwave signals in the optical domain.

With the development of key elements such as laser sources, external modulators, different types of low loss fibres, and photodetectors (PDs), high performance microwave photonic filters (MPFs) can be built [7]. MPFs have potential applications in advanced communication and tactical systems. Some specific applications are: channel selection and channel rejection tunable filters in radio over fibre systems, suppression of manmade interfering signals in radio astronomy applications, lightweight filters for interference suppression in satellite communication, noise and clutter suppression in radar systems and relaxation of the required resolution of Analog/Digital converters, and band-select filters in microwave photonic access networks, etc. [4, 7, 8]. Because of their advantages and potentially wide applications, it is important to develop various types of MPFs.

With regard to the topic of instantaneous frequency measurement, the rapid and ongoing developments in telecommunication and electronic warfare technology demand faster and more flexible systems. Wideband signal processing is thus needed to implement such systems. As explained later on, in electronic warfare (EW) systems, instantaneous frequency measurement (IFM) receivers play an important role. EW is the science or the systems approaches for exploiting and controlling the

electromagnetic spectrum to the maximum possible extent. It exploits the enemy's electromagnetic emissions for obtaining the intelligence on the enemy's order of battle, intentions, and capabilities; it provides countermeasures to deny effective use of communications and weapon systems in all battle phases while preserving the same spectrum for friendly forces [9-11].

Because of the complexity of the electromagnetic environment in a battlefield, there is always a strong demand for signal scanning, identification or analysis over a large frequency range with high possibility of interception. In some EW systems, a number of specialized receivers are jointly employed to reduce the processing load of a single receiver. Therefore, the carrier frequency of a microwave signal is needed to be measured instantaneously using an instantaneous frequency measurement (IFM) receiver before passing it to a specialized receiver for further processing [12]. The modern EW systems are usually required to cover the frequency range 0.8 – 20.0 GHz (L to Ku band); this range is projected to go up to 40 GHz (Ka band) [9]. Many electronic IFM receivers have been developed previously [13-15]. However, because of the limited bandwidth of the microwave components, it is difficult to reach 40 GHz measurement range or even the 20 GHz range. Besides, the electronic solutions also suffer from the previously mentioned common limitations of electronics such as high loss, high power-consumption, bulky implementation, etc. Therefore photonic solutions have been proposed to overcome these limitations.

1.1.1 A typical scheme for MPF

Microwave filtering can be realized by an optical signal processor which is an optical subsystem including delay lines, couplers, fibre Bragg gratings (FBGs) etc. Since the signal processing is in the optical domain, electrical/optical and optical/electrical

conversions are needed and these are carried out using an electro-optic modulator (EOM) and a PD.

(a)

(b)

Fig. 1.1 Microwave photonic filter (a) general layout, and (b) a possible scheme for a transversal microwave photonic filter. RF: radio frequency; CW: continuous wave; FBG: fibre Bragg grating; PD: photodetector.

The general layout of a microwave photonic filter is shown in Fig. 1.1 (a). In an optical signal processor, the modulated optical signal has to be divided, delayed, weighted and summed at the PD as illustrated in Fig. 1.1 (b). Since photonic signal processing structures are linear time-invariant systems, the usual z and discrete-time Fourier transform techniques can be used to analyze MPFs. The general expression for a MPF response is [1, 8]

$$H(\Omega) = \sum_{n=0}^N a_n e^{-jn\Omega T} \quad (1.1)$$

where Ω represents the RF signal angular frequency, T denotes the time delay, and a_n is the weight function for each tap. The number of taps of the MPF is decided by N .

Fig. 1.2 Typical periodic response of a 3-tap MPF. MSSR: main to secondary sidelobe ratio, FSR: free spectral range, FWHM: full width half maximum.

An example of a 3-tap filter response is shown in Fig. 1.2 which illustrates the definitions of some basic filter parameters [1]. The filter response is usually a periodic function and the period is known as the filter free spectral range (FSR). For bandpass filters, the spectral selectivity for any passband is given by the full width half maximum (3 dB bandwidth), $\Delta\Omega_{FWHM}$. The filter selectivity of a given resonance is determined by its quality or Q factor

$$Q = \frac{FSR}{\Delta\Omega_{FWHM}} \quad (1.2)$$

When the number of taps is high (>10), the Q factor can be approximated by N for uniform filters. Another important parameter for multi-tap filters is the main to secondary sidelobe ratio (MSSR) which is a measure of rejection of the nonadjacent channels.

1.1.2 A typical scheme for IFM

Photonics-based IFM techniques can be divided into two types: frequency-to-time mapping and frequency-to-power mapping. Because of the advantages of relatively simpler configuration, higher resolution and accuracy, and insensitivity to the type of modulation of the pulsed signal under measurement, frequency-to-power mapping-based IFM receivers have attracted considerable interest.

(a)

(b)

Fig. 1.3 A frequency-to-time mapping IFM receiver (a) concept and (b) power fading functions and ACF [16]. EOM: electro-optic modulator; WDM: wavelength division multiplexer; LD: laser diode; DWDM: dense wavelength division multiplexer; PD: photodetector; DLVA: detector logarithmic video amplifiers; ACF: amplitude comparison function.

The basic concept of a frequency-to-time mapping IFM receiver is to construct an amplitude comparison function (ACF) which is the ratio of two different optical or microwave power functions. An example is shown in Fig. 1.3 (a) [16]. Two optical carriers with different wavelengths are modulated by the same electro-optic modulator and pass through a dispersive medium. Because of the wavelength difference, the experienced chromatic dispersion (CD) is also different. Therefore, two power fading functions are generated with different notch frequencies as shown in Fig. 1.3 (b). By comparing the two functions, ACF can be obtained which is monotonic from Direct-Current (DC) to the notch with lower frequency. ACF establishes the relation between the input microwave signal frequency and the output power in the monotonic region. From this relation, a lookup table can be established for estimating the microwave frequency from the measured power. Actually, the two power fading functions are also monotonic from DC to their respective notch frequency. However, these cannot be used to estimate the input signal frequency because the detected power values depend upon the power of the unknown microwave signal. By using ACF, this power dependence can be eliminated.

1.2 Scope

The scope of our research involves two aspects: novel tunable MPF configurations and new low cost high resolution IFM methods. Keeping in view the principles of the IFM receivers, the IFM can actually be considered as an application of MPF.

Similar to digital filters, MPFs also have two categories, namely finite impulse response (FIR) and infinite impulse response (IIR) filters. Further, MPFs may also have two or more taps. For the two-tap FIR MPFs, there is not much freedom for adjustment of the shape of the filter response because of the limited number of taps.

So these are usually notch filters. The current research on the two-tap MPFs attempts to increase the FSR tuning range and notch rejection using simple structures. Because of the interference of the light signals, incoherent operation of the filter is also an important consideration. Previous work has presented designs of some coherence-free two-tap MPFs; however, the tunability in these designs is limited [17-22]. In this context, fibre Bragg grating is a simple passive component for wavelength selection. It has the properties of tunable reflection band, wavelength dependent time delay, and tunable differential group delay (DGD). By using these properties of FBGs and improving upon some previous schemes [23, 24], three coherence-free two-tap tunable MPF designs are proposed and experimentally demonstrated here with improved tunability, stability and notch rejection.

For multi-tap FIR MPFs, we have more freedom in adjusting the filter response through controlling the number and weight of the taps. To increase the number of taps, a laser array, sliced wideband source or a multi-wavelength laser can be used as the light source in the MPF [25-33]. Since a wavelength tunable laser array is usually costly and a sliced wideband source has the problems of low output power and tap weight control, a multi-wavelength laser is chosen in our research. Because each wavelength corresponds to one tap, the multi-tap FIR MPFs are inherently coherence-free. The challenging part for the multi-wavelength laser-based MPF is the tuning of the FSR which requires change in the wavelength spacing of the laser source and control of the power profile of the multi-wavelength laser (tap weight control). It is also important to build a theoretical model of the filter and set up a filter design procedure so that the key parameters of the MPF can be designed to meet the requirements.

The electrical IIR filters are based on feedback loops. IIR MPFs can be realized using RDL. Previous work has used 2x2 optical couplers and has demonstrated various functions such as convolution, correlation, frequency filtering, pulse compression, and even lattice signal processing [34-38]. Since 3x3 optical couplers are commercially available now, these can be used to replace the 2x2 optical couplers for improving the performance of IIR MPFs.

Currently, the photonic assisted microwave frequency measurement techniques can be divided into three categories: scanning receiver, optical channelizer receiver and IFM receiver [16, 39-51]. Scanning receiver cannot meet the ‘instantaneous’ requirement of the EW systems because of the scanning time. Optical channelizer receivers can perform instantaneous measurement but the measurement resolution is limited by the channel bandwidth and the number of channels. Thus the IFM receivers have attracted more attention from researchers. As described in the previous section, the IFM techniques can be further divided into two types: frequency-to-time mapping and frequency-to-power mapping. The frequency-to-time mapping receivers can measure multiple signals at the same time but require a costly real time high speed oscilloscope [44, 45]. So the previous work has mainly considered frequency-to-power mapping. The basic idea of frequency-to-power mapping is to construct an ACF by comparing two measured microwave amplitudes or powers or optical powers from the two arms of the system. The unknown microwave signal frequency can be estimated through the monotonic ACF. To construct an ACF, multiple laser sources, electro-optical modulators, PDs or power meters, and/or other passive components are used. The use of multiple laser sources may have problems such as wavelength-spacing drift and relative power fluctuation, which may increase the measurement error. The use of multiple modulators may also increase measurement errors since the modulators may

not be perfectly matched. When a Mach-Zehnder modulator (MZM) is used, the drift of the bias will also contribute to measurement errors. Because of the special area of application of IFM, i.e. EW systems, it is important to reduce the measurement error and system complexity, while maintaining the instantaneous measurement feature.

1.3 Major contributions of the thesis

Major contributions of the thesis are described below briefly.

Three types of MPFs and four designs for IFM have been investigated. In all cases, the performance has been predicted theoretically and measured results that match well the theoretical ones have been obtained.

A. FBG based two-tap filters

1. Using the DGD of a high-birefringence (Hi-Bi) FBG and by applying mechanical stress, continuous FSR tuning of 1.11 GHz with a notch rejection of 32 dB has been achieved.
2. Using a pair of polarization beam splitters (PBSs) to form two arms, putting a Hi-Bi linearly chirped fibre Bragg grating (LCFBG) in one of the arms, and by varying the wavelength of operation, more than 5 GHz FSR tuning and 40 dB notch rejection have been achieved. The filter response is very stable because of the high extinction ratio of the PBSs and polarization maintaining structure.
3. Incorporating a nonlinearly chirped fibre Bragg grating (NLCFBG), an MPF has been realized based on the CD induced power fading of double sideband (DSB) signals. An ultra high notch rejection of 45 dB with 4.7 GHz FSR tuning has been realized by varying the working wavelength.

B. Multi-tap FIR filter based on multi-wavelength laser

The window-method of the digital filter design has been extended to the design of FIR MPFs. A theoretical model has also been setup which is not only suitable for the multi-wavelength laser but also for the MPFs based on a laser array or a sliced wideband source. With the help of this theoretical model, we have built a windowed Fabry-Pérot filter-based multi-wavelength laser which has Blackman power profile and has wavelength spacing tunability. Using this laser and phase modulation, a bandpass MPF is realized with a 3 GHz passband centre frequency tuning range and more than 25 dB out-of-band rejection.

C. IIR MPF based on 3x3 coupler

A 3x3 collinear optical fibre coupler-based reflective double recirculating delay line MPF has been realized. A higher Q value compared to a 2x2 reflective recirculating delay line (RRDL) filter and both passbands and notches at the same time have been demonstrated. By adjusting the time delays of the two loops, the response of the proposed filter can be tailored.

D. IFM receiver designs

1. The first design utilizes the asymmetric non-linear group delay profile of a NLCFBG and requires only one tunable laser source and one MZM. Similar to other early IFM receivers, two low pass responses have been used to obtain the ACF. The measurement range is limited to 8.22 GHz - 9.82 GHz.
2. To extend the measurement range and increase the measurement accuracy, the second IFM receiver design is based on two complementary transfer functions. The resultant ACF varies from negative infinity to positive infinity on a log scale.

Thus the measurement range and accuracy improve. One laser source and two modulators are required. A wide measurement range from DC to 13.8 GHz with a measurement error less than ± 0.3 GHz has been achieved. This design also has a potential advantage of fast calculation of microwave frequency because the ACF has a very simple form.

3. The third IFM receiver design further reduces the number of modulators and measurement error. With the help of phase to intensity conversion and DGD induced power fading, the intensity modulator in the second IFM design can be removed. The proof of concept experiment shows 1.7 GHz to 12.2 GHz measurement range with a reduced measurement error of ± 0.07 GHz.
4. All the previous IFM receiver designs use two PDs or power meters. In the fourth IFM receiver design, for the first time, the number of PDs has been reduced to one by using the concept of an IIR filter to replace the ACF. The current measurement range is from 6.9088 to 6.9190 GHz and is limited by the long loop length. By using integrated solutions, wideband measurement is expected.

1.4 Thesis organization

Chapter 1 of the thesis introduces the background and motivation for the present photonic techniques for microwave signal processing and frequency measurement. The scope of the research is clearly stated, and the major contributions are elaborated, followed by the organization of the thesis. Chapter 2 gives a review of modulators and the calculation of the signal modulation, propagation and detection in the fibre systems. Several important ideas and equations presented in this chapter are used throughout the thesis. Chapter 3 describes three different FBG-based MPFs with improved tunability and stability. Chapter 4 focuses on the increase of the number of

filter taps by using a multi-wavelength laser. High performance multi-wavelength fibre ring laser source is constructed and the corresponding MPF shows bandpass response with centre frequency tunability and good out-of-band rejection ratio. A RDL-based IIR filter, which shows potential for achieving a high Q-factor band pass filter with deep notches at stop bands, is presented in Chapter 5. Chapter 6 describes designs for instantaneous frequency measurement receivers. Four configurations have been proposed with improved measurement resolution, reduced system complexity and measurement error. The thesis ends with the conclusion and plan for future work in Chapter 7.

Chapter 2 MODULATORS FOR MICROWAVE PHOTONICS

In order to process microwave signals in optical domain, electrical to optical conversion is essential. There are mainly two kinds of method to modulate the microwave signal onto the optical carrier: direct modulation of a laser diode (LD) and by using an external modulator. For direct modulation, the modulating electrical signal is applied to the LD directly which modulates the driving current of the LD and thus the output optical power [52]. The reported highest 3-dB bandwidth of direct modulation is about 30 GHz for laser diodes in the 1550 nm communication band [53, 54]. Although this bandwidth is adequate for many MWP applications, the frequency chirping effect, which is the laser relaxation oscillation caused by the change in refractive index with carrier density, will cause wavelength variation of the optical carrier [55-57]. Due to this wavelength variation, the optical signal will have an extra undesired broadening effect when propagating through a dispersive medium. Besides, direct modulation provides intensity modulation; it cannot provide phase modulation which is required for some applications.

The chirp parameter, which is defined as the ratio of the changes in the real to the imaginary parts of the refractive index, is typically between 4 to 6 for semiconductor lasers under direct modulation so external modulation is usually used to minimize this effect although the system complexity and cost may be higher [56, 58, 59]. This is the main reason that external modulation is preferred to direct modulation in microwave photonics applications and high-speed transmission systems.

This chapter reviews external modulators and the calculation of -signal modulation, propagation and detection in the fibre systems. Several important ideas and equations presented in this chapter are used throughout the thesis.

2.1 Review of external modulators

Currently, several types of modulators are available. These are Lithium Niobate (LiNbO_3) modulators, semiconductor MZMs, semiconductor electro-absorption modulators (EAMs), and polymer modulators [59, 60]. Based on the Franz-Keldysh effect in bulk semiconductor active layer [61-63] or quantum-confined Stark effect in semiconductor quantum-wells [64], the optical absorption coefficient in the material changes in the presence of an electric field. These are denoted as electro-absorption effects and directly result into optical intensity modulation in a single optical waveguide. Currently, bit rates of up to 40 Gb/s (30 GHz of 3 dB bandwidth) under reverse bias with an extinction ratio of 20 dB or more can be realized in commercial EAMs [65, 66]. Although the EAMs have enough bandwidth, small enough chirp, and the advantage of easy integration with semiconductor lasers, these can only realize intensity modulation.

LiNbO_3 modulators, semiconductor MZMs and polymer modulators are all based on linear electro-optic effect, which is defined as the change of material refractive index under the presence of an electric field. With an external modulating voltage, the phase of the optical carrier is changed, thereby causing phase modulation or intensity modulation in a Mach-Zehnder interferometer configuration. The EOMs usually employ a travelling-wave scheme with a long electrical-optical interaction region for increasing the modulation efficiency and bandwidth. When a III-V semiconductor material, such as GaAs/AlGaAs, is used, low V_π occurs at short wavelengths (1.15

μm). V_π is the voltage required to change the output light intensity from its maximum value to its minimum value for the intensity modulator, or the voltage that causes π phase shift of the optical signal for the phase modulator. In order to operate at the conventional communication optical wavelength band ($1.5 \mu\text{m}$), the V_π will increase to 10-20 V [67, 68].

Organic polymer materials have many advantages such as low dispersion, good match between the RF and optical indices, and the possibility of using a spin-coating fabrication technique. These advantages make the design and fabrication of EOM much easier [59, 69]. Polymer traveling wave modulators have demonstrated 40 GHz bandwidth [69, 70], however, thermal aging and facet cleaving problems make polymer modulators still far from practical use [59].

Among all the kinds of modulators mentioned above, LiNbO_3 modulators are the most widely commercialized modulators. Bandwidths as high as 75 GHz have been demonstrated [71, 72]. So in our work, we have mainly used LiNbO_3 modulators.

Besides the above modulators, micro-ring modulator, EAM and MZM have been realized using silicon photonics technologies [73]. Future high performance silicon modulators could also possibly be used in MWP applications.

2.2 Lithium Niobate phase modulator

The optical field without the consideration of phase and amplitude noise can be expressed as

$$e_{in}(t) = e_0 \cos(\omega_0 t) \quad (2.1)$$

where e_0 and ω_0 are the amplitude and the angular frequency of the electric field. Assume that the modulating microwave signal $f(t)$ is a single frequency cosine function with amplitude V and angular frequency ω_m

$$f(t) = V \cos \omega_m t \quad (2.2)$$

The output optical signal from the phase modulator can be expressed as

$$e_{PM}(t) = e_0 \cos(\omega_0 t + m_f \cos \omega_m t) \quad (2.3)$$

where $m_f = \pi V / V_\pi$ is the phase modulation index.

Using Bessel functions of the first kind, Eq. (2.3) can be rewritten as

$$e_{PM}(t) = e_0 \sum_{n=-\infty}^{+\infty} J_n(m_f) \cos\left[(\omega_0 + n\omega_m)t + \frac{1}{2}n\pi\right] \quad (2.4)$$

By using Fourier transformation of Eq. (2.4), the frequency spectrum can be derived.

We can see from Eq. (2.4) that the phase modulation generates a series of sidebands with amplitude coefficients determined by the Bessel functions. If this signal is directly measured by a PD, only a DC would be generated.

Under small signal modulation condition, only the first order sidebands need to be considered. With the property of Bessel functions $J_{-n}(m_f) = (-1)^n J_n(m_f)$, Eq. (2.4) can be simplified as

$$\begin{aligned} e_{PM}(t) \approx & e_0 J_0(m_f) \cos(\omega_0 t) \\ & + e_0 J_1(m_f) \cos\left[(\omega_0 + \omega_m)t + \frac{\pi}{2}\right] \\ & - e_0 J_1(m_f) \cos\left[(\omega_0 - \omega_m)t - \frac{\pi}{2}\right] \end{aligned} \quad (2.5)$$

When the modulated optical signal propagates through a section of single-mode fibre (SMF), the signal undergoes distortion because different frequency components travel

at different velocities. Such a dispersion mechanism is called intramodal dispersion or chromatic dispersion (CD).

When a signal centered at the frequency ω_0 propagates through a dispersive medium, such as a section of SMF, the group velocity v_g is defined as [74]

$$v_g = \left(\frac{d\beta}{d\omega} \right)^{-1} \quad (2.6)$$

where β is the propagation constant at ω_0 and the group delay τ_g after propagating a distance L is given by

$$\tau_g = \frac{L}{v_g} = \frac{d\beta}{d\omega} L \quad (2.7)$$

For a signal containing a small spread of frequencies around a carrier frequency ω_0 , the propagation constant $\beta(\omega)$ is a function of wavelength. Expanding $\beta(\omega)$ in a Taylor series

$$\begin{aligned} \beta(\omega) = & \beta(\omega_0) + \left. \frac{d\beta}{d\omega} \right|_{\omega=\omega_0} (\omega - \omega_0) + \frac{1}{2} \left. \frac{d^2\beta}{d\omega^2} \right|_{\omega=\omega_0} (\omega - \omega_0)^2 \\ & + \frac{1}{6} \left. \frac{d^3\beta}{d\omega^3} \right|_{\omega=\omega_0} (\omega - \omega_0)^3 + \dots \end{aligned} \quad (2.8)$$

Then the phase shift after a distance L is expressed as

$$\theta(\omega) = \beta_0 L + \beta_1 (\omega - \omega_0) L + \frac{1}{2} \beta_2 (\omega - \omega_0)^2 L + \dots \quad (2.9)$$

where β_m , denotes the m th derivative of the propagation constant with respect to the frequency ω_0 . In Eq. (2.9), the factor $\beta_1 L$ produces a group delay and β_2 is known as

the group velocity dispersion (GVD) or CD parameter. The higher dispersion terms are very small and can be neglected.

By using $\Delta\omega = (-2\pi c / \lambda^2) \cdot \Delta\lambda$ and neglecting high order dispersion, CD can be calculated as

$$\Delta\tau_D = DL\Delta\lambda \quad (2.10)$$

where $D = -2\pi c\beta_2 / \lambda^2$ is designated as the dispersion coefficient and is expressed in units of ps/nm/km.

As shown in Eq. (2.5), the phase modulation will introduce sidebands to the optical carrier. When the phase modulated signal propagates through a dispersive medium, chromatic dispersion will change the phase relation between the frequency components

$$\begin{aligned} e_{PM}(t) \approx & e_0 J_0(m_f) \cos(\omega_0 t - \theta_0) \\ & + e_0 J_1(m_f) \cos\left[(\omega_0 + \omega_m)t + \frac{\pi}{2} - \theta_{+1}\right] \\ & - e_0 J_1(m_f) \cos\left[(\omega_0 - \omega_m)t - \frac{\pi}{2} - \theta_{-1}\right] \end{aligned} \quad (2.11)$$

where θ_0 , θ_{+1} and θ_{-1} are the phase shifts experienced by the carrier, the upper and lower sidebands, respectively. By using Eq. (2.9) and (2.10), we have

$$\begin{cases} \theta_0 = \beta_0 L \\ \theta_{-1} = \beta_0 L - \beta_1 \omega_m L + \frac{1}{2} \beta_2 L \omega_m^2 \\ \theta_{+1} = \beta_0 L + \beta_1 \omega_m L + \frac{1}{2} \beta_2 L \omega_m^2 \end{cases} \quad (2.12)$$

When an optical signal is detected by a PD, the output photo current of the PD can be written as

$$i_{PD} = R \cdot \left\langle |e(t)|^2 \right\rangle \quad (2.13)$$

where R is a coefficient which includes the responsivity of the PD and optical loss.

By substituting Eq. (2.11) into Eq. (2.13) and ignoring the DC and higher-order harmonics, we have [75, 76]

$$\begin{aligned} i_{PD} &\propto RJ_0(m_f)J_1(m_f)\sin\left(\theta_0 - \frac{\theta_{+1} + \theta_{-1}}{2}\right)\cos\left(\omega_m t - \frac{\theta_{+1} - \theta_{-1}}{2}\right) \\ &= RJ_0(m_f)J_1(m_f)\sin\left(\frac{DL\lambda_c^2\omega_m^2}{4\pi c}\right)\cos[\omega_m(t - \tau_0)] \\ &= RJ_0(m_f)J_1(m_f)\sin\left(\frac{\pi DL\lambda_c^2 f^2}{c}\right)\cos[2\pi f(t - \tau_0)] \end{aligned} \quad (2.14)$$

where $\tau_0 = \beta_1 \omega_m L$ is the optical signal group delay, λ_c is the optical carrier wavelength and f is the modulating microwave frequency. From Eq.(2.14) we can see that the CD can convert a phase modulated signal to an intensity modulated signal (the intensity modulation will be introduced in the next section). This is the so-called chromatic dispersion induced phase modulation to intensity modulation (PM-IM) conversion. The sine term describes the amplitude of the detected microwave signal, which is the transfer function of the PM-IM conversion.

$$H_{PM-IM}(f) \propto \sin\left(\frac{\pi DL\lambda_c^2 f^2}{c}\right) \quad (2.15)$$

Fig. 2.1 Calculated frequency response of CD induced power fading.

Fig. 2.1 shows a calculation of $P_{PM-IM}(f)$ using the parameters $L = 15$ km, $D = 16.75$ ps/nm/km, and $\lambda_c = 1550$ nm. We can see that the detected microwave power is decided by the overall dispersion DL , and the detected microwave power is zero at the frequencies $\frac{\pi DL \lambda_c^2 f^2}{c} = k\pi$, $k = 0, 1, 2, \dots$. This is known as the CD-induced power fading.

2.3 Lithium Niobate Mach-Zehnder Modulator

The MZM can be modeled as two phase modulators in parallel and the amplitudes of the microwave drive signals applied to the electrodes of the two phase modulators (two arms) are equal. Then the output optical field is [77]

$$e_{MZM}(t) = \frac{\sqrt{2}e_0}{2} \left\{ \cos(\omega_0 t + \phi + m_f \cos \omega_m t) + \cos[\omega_0 t + m_f \cos(\omega_m t + \theta)] \right\} \quad (2.16)$$

where θ is the phase difference between the microwave signals applied to the two arms and $\phi = \pi \frac{V_{DC}}{V_\pi}$ is the DC bias induced phase difference. By expanding Eq. (2.16)

using Bessel functions, we have

$$e_{MZM}(t) = \frac{\sqrt{2}e_0}{2} \left\{ \begin{array}{l} J_0(m_f)(\cos\phi + 1)\cos(\omega_0 t) \\ -J_0(m_f)\sin\phi\sin(\omega_0 t) \\ -J_1(m_f)\cos\phi[\sin(\omega_0 - \omega_m)t + \sin(\omega_0 + \omega_m)t] \\ -J_1(m_f)\sin\phi[\cos(\omega_0 - \omega_m)t + \cos(\omega_0 + \omega_m)t] \\ -J_1(m_f)\{\sin(\omega_0 - \omega_m)t - \theta\} + \sin(\omega_0 - \omega_m)t + \theta \\ + \dots \end{array} \right\} \quad (2.17)$$

For a single electrode MZM, $\theta = (2k+1)\pi$, the modulation format is DSB modulation.

The operating point of the MZM is decided by ϕ which can be adjusted by varying the DC bias. When $\phi = 2k\pi$, the terms representing optical carrier in Eq. (2.17) reach the maximum value $2J_0(m_f)\cos(\omega_0 t)$. This is the maximum transmission point of the MZM. When $\phi = (2k+1)\pi$, the MZM works under null or minimum transmission point where the first two terms in Eq. (2.17) are zero.

When $\phi = (2k+1)\pi/2$, the MZM works at quadrature point (linear point). In such a case Eq. (2.17) becomes

$$e_{MZM}(t) = \frac{\sqrt{2}e_0}{2} \left\{ \begin{array}{l} J_0(m_f)\cos(\omega_0 t) - J_0(m_f)\sin(\omega_0 t) \\ -J_1(m_f)[\cos(\omega_0 - \omega_m)t + \cos(\omega_0 + \omega_m)t] \\ +J_1(m_f)[\sin(\omega_0 - \omega_m)t + \sin(\omega_0 + \omega_m)t] \\ + \dots \end{array} \right\} \quad (2.18)$$

Under small signal condition and following the same procedure as for Eq. (2.11) to Eq. (2.13), we have

$$H_{IM}(f) \propto \cos\left(\frac{\pi DL\lambda_c^2 f^2}{c}\right) \quad (2.19)$$

2.4 Summary

Microwave photonic systems process the microwave signals in the optical domain, so the modulation of microwave signals onto the optical carrier is an essential step. This can be achieved by direct modulation of a LD but it has limited bandwidth and causes carrier frequency chirping effect. Furthermore, direct modulation is only applicable to LD but not to other light sources such as a wideband source or a fibre laser etc. Therefore external modulation is preferred in microwave photonics applications.

There are several types of external modulators. Based on Franz-Keldysh effect or Stark effect, an EAM can be realized. The current commercialized EAMs have a bandwidth up to 30 GHz and high extinction ratio of 20 dB. However, an EAM can only realize intensity modulation. LiNbO₃ modulators, semiconductor MZMs and polymer modulators are based on linear electro-optic effect. The semiconductor MZMs usually require high V_π around 10-20 V, while, organic polymer modulators may be affected by thermal aging and facet cleaving problems. So the LiNbO₃ modulators become the most widely commercialized modulators and we mainly use LiNbO₃ modulators in our work.

Based on their structure and working principle, LiNbO₃ modulators can be divided into two types: phase modulators and MZM intensity modulators. Expressions for the modulated signal using LiNbO₃ phase and intensity modulators, including the effect of the CD of the optical fibre link, and for the detected microwave signal, have been presented in this chapter.

Chapter 3 TWO-TAP MICROWAVE PHOTONIC

FILTERS

3.1 Introduction

A typical microwave photonic filter has been described in Section 1.1.1. As can be seen from Fig. 1.1 and Eq. (1.1), optical interference may occur at the photodetector because of the polarized nature of light. The coherent summing of optical signals is sensitive to polarization fluctuation caused by environmental perturbation. Therefore the incoherent approach is more attractive in practice for stable operation. To satisfy the requirement of incoherent summing, either the laser coherence time is needed to be shorter than the optical delay time, or the states of polarization (SOPs) of the optical signals after division needs to be orthogonal. The first approach can be implemented by using a laser array [25, 26] or a sliced wideband source [27]. Usually, this approach provides a high number of taps and the flexibility of controlling the tap weight but the complexity of the system is higher. If a single wavelength laser source is used, either long time delay lines are required [20] which restricts the FSR, or optical mixing effect is used [78] which increases the system cost. So, an implementation based on a single wavelength laser source is not preferred.

For the second approach, the coherence time of the laser does not need to be considered. Although the number of taps is limited, usually to 2, the complexity is lower than the first approach. In this chapter, we mainly consider the orthogonal polarization approach.

Fig. 3.1 Experimental setup of the orthogonal polarization MPF using Hi-Bi LCFBG and Hi-Bi coupler [24]. EOM: electro-optical modulator; PC: polarization controller; PBS: polarization beam splitter; LCFBG: linearly chirped fibre Bragg grating; PD: photodetector.

Tunable MPFs using orthogonal polarization have been demonstrated in [21] and [22] by using polarization maintaining structures. However, [21] has 50% of power loss because of the 3 dB Hi-Bi coupler. In [22] a good FSR tuning of about 4 GHz was achieved but the adjustable DGD element is costly and the tuning is step tuning. By using FBGs, the MPF structure can be simpler and the cost can be lower [20, 23, 24]. A normal FBG was used in [20]. The optical path difference was kept longer than the coherence length of the tunable laser. So the FSR was limited to MHz range. By combining orthogonal polarization with FBGs, this problem can be solved. In [23], Hi-Bi FBGs were used and a 1 GHz FSR step tuning was shown. Continuous tunability was achieved in an MPF which consisted of a Hi-Bi linearly chirped fibre Bragg grating (LCFBG), a PBS and a Hi-Bi coupler [24] as shown in Fig. 3.1. However, the FSR tuning range achieved by applying uniform strength to the grating was relative small – in the range of MHz. Also, the notch rejection may be affected for this configuration, with a LCFBG inside a fibre loop, because the limited reflectivity of the grating causes signal crosstalk. Further, the Hi-Bi fibre coupler also

causes 6 dB loss. In order to overcome these limitations, several new MPF configurations are proposed here.

3.2 Continuously tunable microwave photonic filter based on high-birefringence linearly chirped grating

Our first proposed MPF configuration is based on a Hi-Bi LCFBG as the tuning component. The filter response tunability is realized through a change in the differential group delay of the Hi-Bi LCFBG by applying gradient tension or by adjusting the operating wavelength. Free spectral range tuning by 1.11 GHz with about 40 dB notch rejection is achieved [79].

The experimental setup of the proposed MPF is shown in Fig. 3.2. Linearly polarized continuous wave (CW) light from a tunable laser source (TLS, Anritsu MG9638A) is intensity modulated by a MZM (EOSpace AX-0K1-12-PFU-PFU) which is driven by an RF signal from a vector network analyzer (VNA, Anritsu 37369C). The output of the MZM is fed into a Hi-Bi LCFBG through a half-wave plate and a polarization maintaining (PM) circulator. The half-wave plate is used to excite two orthogonal and equal power linear SOP components along the two axes of the PM pigtail of the PM circulator. Since the whole structure is polarization maintaining, the two orthogonal polarization components propagate along the two axes, get reflected by the Hi-Bi LCFBG, are detected by a photodetector (New Focus Model 1544) and the frequency response is measured by a vector network analyzer.

Fig. 3.2 Schematic of the Hi-Bi LCFBG DGD based MPF. TLS: tunable laser source; MZM: Mach-Zehnder Modulator; PM: polarization maintaining; LCFBG: linearly chirped fibre Bragg grating; VNA: vector network analyzer; RF: radio frequency.

For an FBG, the strongest mode coupling occurs at the Bragg wavelength, λ_B , which is the centre wavelength of the input light that can be back reflected from the grating. The Bragg wavelength is a function of the grating pitch $\Lambda(z)$ [80]

$$\lambda_B = 2n_{eff}\Lambda(z) \quad (3.1)$$

where n_{eff} is the effective refractive index of the fibre. A linearly chirped grating has a variable pitch which is a linear function of the distance along the grating so that the Bragg wavelength and reflection location along the grating form a linear relation. A Hi-Bi LCFBG acts like two gratings corresponding to the two orthogonal polarization axes with the same pitch function. Since the effective refractive indices of the two axes are different, the propagation speed of light along the two axes, slow and fast, are also different. This results in a shift between the grating reflection spectra for the two axes in the wavelength domain. When the two orthogonally polarized components are reflected by the Hi-Bi LCFBG, they are delayed differently and a certain amount of DGD occurs between the signals along the fast and slow axes. Since the two components are orthogonal, incoherent summing is achieved at the photodetector. The

resulting MPF is a two-tap transversal filter. Its normalized transfer function can be derived based on Eq. (1.1) letting $n = 2$, as

$$\begin{aligned} H(f) &= \frac{1}{2} \left(1 + e^{-j2\pi f \Delta T} \right) = \frac{1}{2} e^{-j\pi f \Delta T} \left(e^{j\pi f \Delta T} + e^{-j\pi f \Delta T} \right) \\ &= e^{-j\pi f \Delta T} \cos(\pi f \Delta T) \end{aligned} \quad (3.2)$$

where $f = \frac{\omega}{2\pi}$ is the radio frequency and ΔT is the DGD induced time delay difference between the two axes. The coefficient “ $\frac{1}{2}$ ” is to make the total power of the two taps equal to 1. Thus the amplitude response of the filter is

$$|H(f)| = |\cos(\pi f \Delta T)| \quad (3.3)$$

Since the chirp rate of the LCFBG is not high, the dispersion of the LCFBG is not considered while deriving Eq. (3.2) and (3.3). If the dispersion cannot be neglected, the term $\cos\left(\frac{\pi D_{total} \lambda_B^2 f^2}{c}\right)$ should be included which is the dispersion induced power fading. In this cosine term, D_{total} is the total dispersion value of the LCFBG. The complete derivation is given in Appendix A starting from Eq. (A.10), the part after phase to intensity modulation.

The FSR of the MPF can be described by

$$FSR = \frac{1}{\Delta T} \quad (3.4)$$

Fig. 3.3 (a) Cantilever beam for tuning the Hi-Bi linearly chirped fibre Bragg grating and the schematic reflectivity and group delay when the spectrum (b) broadens and (c) narrows.

The total DGD contains two parts: the Hi-Bi LCFBG DGD (τ_g), and the fixed DGD (τ_0) which comes from the polarization maintaining fibre (PMF) pigtails of the

circulator and the Hi-Bi LCFBG. The FSR is tuned by applying gradient strength to the Hi-Bi LCFBG to cause a change of the pitch. Fig. 3.3 (a) shows the cantilever beam structure used for this purpose [81]. The grating is attached in a tilted manner onto the lateral side of a right triangular cantilever beam. For a single mode fibre LCFBG, when applying a force or displacement on the free end of the cantilever beam, half of the grating is under tension while the other half is under compression. The strain at the centre of the grating is zero. As a result, the grating spectrum broadens or narrows with a fixed centre wavelength which further changes the group delay slope. So our MPF has a much bigger FSR tuning range compared with [24] in which uniform stress does not change the group delay slope. For a Hi-Bi LCFBG, the reflection spectra for both slow and fast axes change simultaneously. Although the slopes of the group delay curves for both axes are always the same, the DGD changes with the reflection spectra as shown in Fig. 3.3 (b) and (c). By using this property, the FSR tuning is achieved through varying the DGD ($\Delta\tau_g$) of the Hi-Bi LCFBG at a certain wavelength. Then, Eq. (3.4) can be rewritten as

$$FSR = \frac{1}{\tau_0 + \tau_g + \Delta\tau_g} \quad (3.5)$$

where $\Delta\tau_g$ denotes the change of the DGD of the Hi-Bi LCFBG.

Fig. 3.4 FSR tuning range vs. differential group delay of the Hi-Bi linearly chirped fibre Bragg grating.

From Eq. (3.5) we can see that the FSR can also be tuned by varying τ_g . This can be achieved through adjusting the working wavelength because different working wavelengths result in different optical signal reflection positions in the Hi-Bi LCFBG which further changes τ_g . The PM pigtail length also has an effect on FSR. However, since the axes of the Hi-Bi LCFBG are aligned with the axes of the PM pigtail and τ_0 is much smaller than τ_g for several meters of PMF, we mainly consider the DGD induced by the Hi-Bi LCFBG. Fig. 3.4 shows the calculated results of the relationship between FSR tunability and the DGD of the Hi-Bi LCFBG. In the calculation, a $\Delta\tau_g$ of 30 ps is used. From the curve we can see that in order to increase the FSR tuning range, the DGD of the Hi-Bi LCFBG needs to be reduced.

Fig. 3.5 (a) Hi-Bi grating reflection spectrum with tuning and (b) corresponding measured frequency responses

The Hi-Bi LCFBG used in the experiment is fabricated by exposing a hydrogen-loaded PM fibre to a 244 nm UV laser beam through a linearly chirped phase mask. Fig. 3.5 (a) shows the measured reflection spectra for the fast and slow axes. It is clearly seen that the spectral width changes with a fixed centre wavelength of 1554.35 nm. 1553.8 and 1554.9 nm are used as the operating wavelengths to compare the FSR tuning through gradient stress and wavelength adjustment. As shown in Fig. 3.5 (b), for 1553.8 nm, the two FSRs for broadening and narrowing are 4.84 GHz and 5.07GHz. For 1554.9 nm, the corresponding FSRs are 4.41 GHz and 5.52 GHz. Thus,

an FSR tuning range of 1.11 GHz can be achieved by tuning the grating at the fixed wavelength of 1554.9 nm. The FSR tuning range for 1553.8 nm is smaller because the long wavelength port of the Hi-Bi LCFBG is connected to the PM circulator which makes the grating DGD for 1553.8 nm smaller than that for 1554.9 nm as indicated in Fig. 3.4. If one only changes the operating wavelength, the maximum tunability of FSR is 0.45 GHz. This is because the DGD induced by the wavelength tuning only changes the group delay value while the mechanical tuning contributes an additional DGD change.

3.3 Wide range continuously tunable microwave photonic filter using high-birefringence linearly chirped fibre Bragg grating and polarization beam splitters

The filter response obtained in section 3.2 is caused by the DGD of the Hi-Bi LCFBG. We know that the DGD value depends on the difference between the refractive index for the slow and fast axes of the PMF. However, when the two orthogonal polarized optical signals are propagating separately in two arms, a much greater change in time delay can be achieved by changing the length of one of the arms. Based on this observation, the tuning range can be extended further [82].

The experimental setup of the proposed MPF is shown in Fig. 3.6. CW light from a TLS (Anritsu MG9638A) is intensity modulated by a MZM (EOSpace AX-0K1-12-PFU-PFU) which is driven by a RF signal from a VNA (Anritsu 37369C). The output of the MZM is fed to a half-wave plate which is used to excite two orthogonal and equal power linear SOP components along the slow and fast axes of the PMF. These two components are then split by the PBS I. The component along the slow axis goes to the upper arm, gets reflected by a Hi-Bi LCFBG and reaches the PBS II. The

component along the fast axis propagates in the lower arm and goes to PBS II directly. In this setup, only the slow axis of the LCFBG comes in use. The reason for using a Hi-Bi grating here is to maintain the linear SOP. When the two components reach the PBS II, these are recombined and output from port 1. Since the connections of the PBS II are just the reverse for those of the PBS I, the two components at PBS II port 1 are orthogonal. Thus, incoherent summing is achieved at the photodetector.

Fig. 3.6 Configuration of the wide range tunable MPF using PBSs. TLS: tunable laser source; MZM: Mach-Zehnder Modulator; PBS: polarization beam splitter; PM: polarization maintaining; LCFBG: linearly chirped fibre Bragg grating; RF: radio frequency.

The resulting MPF is a two-tap transversal filter and its normalized transfer function is the same as given by Eq. (3.3). However, here ΔT is the total time delay difference between the two arms.

Let the fibre length of the upper and lower arms be L_1 (without including the length of the Hi-Bi LCFBG) and L_2 , and the distance from the grating input to the reflection point be L_g . ΔT can be expressed as

$$\Delta T = \frac{(L_1 + 2L_g)n_s - L_2n_f}{c} \quad (3.6)$$

where n_s and n_f are the effective refractive indices of the slow and fast axes of PMF.

Since $n_s \approx n_f \equiv n_g$, Eq. (3.6) becomes

$$\Delta T \approx \Delta L n_g / c + 2L_g n_g / c = \tau_0 + \tau_g \quad (3.7)$$

where $\Delta L = L_1 - L_2$ is the fixed difference between arm lengths, $\tau_0 = \Delta L n_g / c$ and $\tau_g = 2L_g n_g / c$ are the fixed and tunable time delay differences, respectively, and n_g is the average refractive index. In Eq. (3.6) and (3.7), the time delay difference caused by the birefringence of the fibre pigtails of port 1 of the two PBSs is omitted because the pigtail length is only tens of centimeters. The FSR of the filter is described by

$$FSR = \frac{1}{\Delta T} = \frac{1}{\tau_0 + \tau_g} \quad (3.8)$$

From Eq. (3.7) and (3.8), we can see that the FSR can be tuned by varying τ_g through changing the operating wavelength because L_g is a linear function of wavelength. The FSR tuning range depends on the fixed arm length difference ΔL or τ_0 . Fig. 3.7 shows the calculated relation between ΔL and FSR tuning range when L_g is 10 cm.

Fig. 3.7 FSR tuning range vs. difference in arm lengths

Fig. 3.8 Measurement of (a) frequency response for different operating wavelengths and calculated response for 1555.2 nm with a fixed arm length difference of 3.1 cm, and (b) comparison between the measured and calculated FSRs vs. wavelength.

In our experiments, the Hi-Bi LCFBG is the same as the one in section 3.2. The slow axis of the Hi-Bi LCFBG is used, with the operating wavelength ranging from 1554.1 nm to 1555.2 nm. The grating short wavelength port is connected to the PM circulator port 2, so a shorter operating wavelength corresponds to a higher FSR.

The measured frequency response of the proposed filter is shown in Fig. 3.8 (a). When the fixed arm length difference ΔL is 3.1 cm, the measured FSR at operating wavelengths 1554.1 nm, 1554.5 nm, 1555 nm and 1555.2 nm is 6.65 GHz, 2.88GHz, 1.67 GHz, and 1.48 GHz, respectively. More than 5 GHz FSR tuning range is achieved. A calculated response based on 1555.2 nm operating wavelength is also given. Good agreement can be seen. The calculated FSR tuning curve versus operating wavelength with $\Delta L = 3.1$ cm is shown in Fig. 3.8 (b). It also agrees well with the measurement. As indicated by Fig. 3.7, the FSR tuning range is also determined by ΔL . If we add a polarization maintaining time delay line in one of the arms for adjusting τ_0 , FSR tunability can be extended further.

The frequency response of the proposed filter is very stable. This is because the polarization maintaining structure greatly reduces the SOP fluctuation caused by environmental perturbation. Another major reason is the application of PBSs. PBS has a high extinction ratio between ports 2 and 3, e.g., 30 dB in our case. By using the PBSs as beam splitting and combining components, the crosstalk between the slow and fast axes at the second PBS port 1 is suppressed. Fig. 3.9 (a) shows the results of measurements for one hour with an interval of 10s between each measurement when the operating wavelength is 1555.2 nm. Very good stability can be seen. The fluctuation of the peaks is less than 1 dB and all the notches remain deeper than 40 dB. For comparison, similar measurements are carried out with PBS II replaced by a

PM coupler (extinction ratio 21 dB). In this case, Fig. 3.9 (b) shows that the peak variation is about 3.5 dB and the notch depths range from 25 dB to 50 dB. The small FSR in Fig. 3.9 (b) is due to the pigtails of PM coupler which increase the fixed arm length difference.

(a)

(b)

Fig. 3.9 Stability measurements at 1555.2 nm over one hour with 10s interval for (a) the proposed filter structure and (b) the structure using a PM coupler to replace PBS II.

3.4 Nonlinearly chirped grating based continuously tunable high notch rejection microwave photonic filter

The two previously proposed MPFs in this chapter use orthogonal polarizations in order to achieve incoherent operation. One disadvantage of such schemes is that the

SOP of the optical signal needs to be controlled carefully so as to excite equal power along the two axes of the PMF; the notch rejection is decided by the power ratio along the two axes. In this section, we introduce a continuously tunable high notch rejection MPF which is based on a NLCFBG and a circulator. Different from the previous filter structures, the filter response of the proposed structure is based on the CD of the two sidebands in a DSB modulated signal. Because of the nature of the DSB, the powers of the two sidebands are exactly the same which results in a very high notch rejection. Since there is only one optical signal, a stable operation is ensured [83].

Fig. 3.10 Schematic of the NLCFBG filter. TLS: tunable laser source; MZM: Mach-Zehnder modulator; VNA: vector network analyzer.

The experimental setup of the proposed MPF is shown in Fig. 3.10. CW light from a TLS (Anritsu MG9638A) is intensity modulated by an x-cut Mach-Zehnder modulator (EOSpace AX-0K1-12-PFU-PFU) driven by a microwave signal from a VNA (VNA, Anritsu 37369C). A direct-current bias is applied to make the MZM operate at the quadrature point. The output of the MZM is fed to an NLCFBG through a circulator. After getting reflected, the optical signal passes through the circulator again, is detected by a photodetector and the frequency response is measured by the VNA.

In our filter configuration, double sideband modulation is used. When the modulated optical signal gets reflected by the NLCFBG, the chromatic dispersion of the NLCFBG causes a certain time delay difference between the upper and lower sidebands of the optical signal. Thus, the mechanism for achieving the filter action is the power fading of the double side band modulated signal induced by the chromatic dispersion of the NLCFBG. Since the MZM is biased at the quadrature point and is operated under small signal condition, the power levels for the two first order sidebands at the output of the MZM are almost the same and the higher order sidebands can be neglected. This results in a very high level of notch rejection. The stable operation of this filter configuration can be guaranteed because the two sidebands propagate through the same optical path and they experience the same ambient environmental disturbance.

The grating pitch of the NLCFBG, which is designed and fabricated by the author, is a second order polynomial function of grating position

$$\Lambda(z) = \Lambda_0 + az + bz^2 \quad -L/2 < z < L/2 \quad (3.9)$$

where the coefficients are $\Lambda_0 = 1068.97 \text{ nm}$, $a = 1.38 \times 10^{-8} \text{ nm/cm}$, $b = 1.284 \times 10^{-7} \text{ nm/cm}^2$, and the grating length $L = 10 \text{ cm}$. From Eq. (3.1) and (3.9), we can find a second order function relating the Bragg wavelength and group delay, and a linear relation between Bragg wavelength and chromatic dispersion.

The filter power response can be expressed as

$$|H(f)| = \cos^2 \left(\pi D_{NLCFBG} \lambda_B^2 f^2 / c \right) \quad (3.10)$$

where f is the modulating radio frequency and D_{NLCFBG} is the NLCFBG-induced dispersion. The notch frequencies in the filter response are given by

$$f = \sqrt{\frac{(2k+1)c}{2D_{NLCFBG}\lambda_B^2}} \quad k = 0, 1, 2, \dots \quad (3.11)$$

Eq. (3.9) to (3.11) show that the grating dispersion and the filter response are a function of the optical wavelength. So the filter response can be tuned by changing the laser source wavelength. To use the grating dispersion effectively, an x-cut Mach-Zehnder modulator is used here for its zero chirp [59].

The NLCFBG used in the experiments was fabricated by exposing a hydrogen-loaded fibre to a 244 nm UV laser beam through a nonlinearly chirped phase mask. Fig. 3.11 shows the measured reflection spectrum of the NLCFBG. It is seen that the grating has a reasonably flat reflection spectrum over the wavelength range 1556.9 nm to 1557.7 nm. This is the range of wavelengths used in the experiment.

Fig. 3.11 Measured reflection spectrum of the nonlinearly chirped grating.

Fig. 3.12 Measured frequency response of the nonlinearly chirped fibre Bragg grating based microwave photonic filter for different optical wavelengths. Calculated response at 1556.89 nm is also included.

The corresponding measured filter frequency response for different wavelengths is shown in Fig. 3.12. The measurement results show that a 4.7 GHz notch frequency tuning range with notch rejection more than 45 dB has been achieved. The calculated response at 1556.89 nm in Fig. 3.12 is based on the calculated NLCFBG dispersion using Eq. (3.11). Good agreement can be seen. The tuning range of the notch frequency is determined by the flat reflection region of the NLCFBG. If the reflection region is not flat, there occurs a power imbalance between the two sidebands which in turn results in reduction of the notch rejection. By improving the grating fabrication process and enlarging the grating flat reflection region, the tuning range of the notch frequency can be extended further.

3.5 Summary

MPFs usually require summing of the optical taps at the PD. The coherent summing of optical signals is sensitive to polarization fluctuation caused by environmental perturbation, so the incoherent summing is more attractive in practice for stable operation. To achieve incoherent operation, either the SOP of the optical signals needs

to be orthogonal, or the light source coherence time needs to be shorter than the optical delay time. In this chapter, we mainly utilize the first approach. Because two orthogonal polarizations are used, the number of taps is two. Therefore, we cannot adjust the shape of the filter response and the main focus is on the tunability.

With the understanding of previous work reported in the literature, we propose an MPF which utilizes the time delay difference between the slow and fast axes (DGD) of a Hi-Bi LCFBG and PMF pigtails. An FSR tuning of 1.11 GHz with 32 dB notch rejection is achieved by applying mechanical stress to the grating; these results match our theoretical expression.

To further increase the FSR tuning range, we modify the filter structure by using a pair of PBSs to form two arms and put a Hi-Bi LCFBG in one of the arms. More than 5 GHz FSR tuning with more than 40 dB notch rejection is achieved and these results agree well with the theoretical calculations. Furthermore, the high extinction ratio of the PBSs and polarization maintaining structure make the filter response very stable.

The previously described two filter designs use orthogonal polarization. In such designs, careful polarization adjustment is needed to excite equal power along the two axes of the PMF and the notch rejection is decided by the power ratio in between. For the third design reported here, the filter response of the NLCFBG-based MPF is actually the CD induced power fading of the DSB signals. So, the polarization control is avoided. The two sidebands can be considered as the two taps of the MPF. Because the sidebands have equal power, the notch rejection is more than 45 dB.

A comparison between the performance of the previous MPFs reported in the literature and our proposed MPFs is listed in Table 3.1. Generally, our designs have a higher notch rejection and a bigger FSR tuning range.

Table 3.1 Comparison of the FBG-based MPFs

MPF	Principle	Notch rejection (dB)	Maximum FSR tuning (GHz)
Ref. [23]	Time delay between slow and fast axes	30	1 (step tuning)
Ref. [24]	Time delay between slow and fast axes	28	0.063
MPF 1	Time delay between slow and fast axes	32	1.11
MPF 2	Time delay between two arms	40	5
MPF 3	CD induced power fading	45	4.7

Chapter 4 MULTI-TAP MICROWAVE PHOTONIC

FILTERS

The MPF configurations proposed in Sections 3.2 and 3.3 achieve incoherent operation based on the orthogonal polarizations of optical signals. Usually, the number of taps is limited to two in this approach. By cascading multiple PMF sections, the number of taps can be increased but multiple wavelengths with a certain wavelength and power relation are required [84]. Another approach for achieving incoherent operation in MPFs is to use a low coherence light source such as a laser array [25, 26], a sliced wideband source [27-30], or a multi-wavelength laser [31-33] in conjunction with the chromatic dispersion property of SMFs or FBGs. Besides achieving incoherent operation in this approach, the number of taps can be increased dramatically. However, the number of taps may be limited when a laser array is used because of the resultant high system cost. For a spectrum-sliced wideband source, e. g. amplified spontaneous emission (ASE) or super luminescent diode, using FBGs or optical filters, the power levels are usually not high and high insertion loss may be incurred. So, a multi-wavelength laser source is more attractive because it is more cost effective compared with a laser array and has a much higher output power as compared to spectrally-sliced wideband sources.

The number of taps for an MPF based on a multi-wavelength laser, sliced wideband source or a laser array is finite, so these filters belong to the category of FIR filters. This chapter describes our work on the FIR MPFs based on a multi-wavelength laser.

One of the most important parts for this kind of MPF is the multi-wavelength laser. For building such a fibre laser, currently semiconductor optical amplifiers (SOA) and

Erbium-doped fibre (EDF) are the commonly used gain media. Compared to the SOA-based multi-wavelength fibre lasers, multi-wavelength Erbium-doped fibre lasers have higher saturated power, low polarization dependent gain, and higher signal-to-spontaneous-noise ratio. Due to the EDF's homogeneous line broadening at room temperature and the cross gain saturation, it is difficult to achieve stable multi-wavelength generation. Cooling the EDF to 77° K with liquid nitrogen [85] or adding optical feedback and nonlinear gain in the optical fibre [86, 87] can suppress the homogeneous line broadening. But achieving 77° K is impractical in many applications, and feedback requires more components. Recently, multi-wavelength Erbium-doped fibre laser was reported, using a highly-nonlinear dispersion shifted fibre (HNDSF) in a ring cavity [88] or using nonlinear polarization rotation [89] to suppress the line broadening.

In this chapter, we first describe the steps to obtain a stable multi-wavelength Erbium-doped fibre ring laser and its use as the light source in the MPF to get intuitive understanding of multi-tap MPF operation. Then, a theoretical model for such an MPF is developed. Finally, a windowed Fabry-Pérot filter based multi-wavelength laser is setup and is used to obtain a bandpass MPF with the ability to tune the passband centre frequency.

4.1 Microwave photonic bandpass filter based on a multiple dual-wavelengths Erbium-doped fibre ring laser

A multiple dual-wavelengths Erbium-doped fibre ring laser is used as the light source for the proposed MPF. Its schematic diagram is shown in Fig. 4.1 [90]. The fibre laser ring cavity for multiple dual-wavelengths consists of 1 km long HNDSF, two polarization controllers (PCs), a fibre polarizer, a PM Fabry-Pérot filter (Micron

Optics FFP-I), an Erbium-doped fibre amplifier (EDFA, Keopsys KPS-BT2-C-33-PB-SP-FA), an optical isolator and a 10:90 optical coupler. Since the Fabry-Pérot filter is PM type, PC I is used to align the axis of the polarizer with one of the axes of the PM Fabry-Pérot filter to achieve maximum output power. PC II adjusts the output power profile of the multiple wavelengths. The gain of the fibre laser is provided by the EDFA. The isolator in the laser cavity is used to ensure the unidirectional operation of the ring laser.

Fig. 4.1 Experimental setup of the MPF based on multiple dual-wavelengths fibre ring laser. PC: polarization controller; EDFA: Erbium doped fibre amplifier; HNDSF: highly-nonlinear dispersion shifted fibre; SMF: single mode fibre

The PM Fabry-Pérot filter has a fixed FSR of 40 GHz which corresponds to 0.32 nm. The optical frequency response of the Fabry-Pérot filter is measured using a broadband source. Fig. 4.2 (a) shows that the Fabry-Pérot filter has a fairly flat response over a very wide wavelength range. The response consists of multiple dual-peaks which are caused by the birefringence of the PM Fabry-Pérot filter. When a polarizer is added and aligned with one of the Fabry-Pérot filter's axes, only one peak can appear as shown in Fig. 4.2 (b).

Fig. 4.2 Measured optical frequency response of the PM Fabry-Pérot filter: (a) flat response over 60 nm, dual-peaks correspond to the slow and fast axes of the filter; (b) only one peak appears when the SOP is aligned to one of the axes of the filter.

(a)

(b)

(c)

Fig. 4.3 Measurement results for (a) continuous 4 scans of the ring cavity without HNLDSE, (b) bell-shaped multi-wavelength laser with HNLDSE, and (c) bell-shaped multiple dual-wavelength laser.

The 1 km HNLSF has a nonlinear coefficient of $14 \text{ W}^{-1} \cdot \text{km}^{-1}$ with the zero dispersion wavelength at 1558 nm. The output power of the EDFA is 13 dBm. The HNLSF induces four-wave-mixing (FWM) effect in the ring cavity which can stabilize the output [88]. Fig. 4.3 (a) shows 4 continuous measurements when the HNLSF is disconnected. Both power and the number of wavelengths change with time. The corresponding MPF frequency response is dominated by noise. When the HNLSF is added, together with the gain competition effect of the EDFA, the output power profile of the multiple wavelengths becomes a stable bell shape after adjusting the PC II to a suitable position; this is shown in Fig. 4.3 (b). Since the PM Fabry-Pérot filter has birefringence, multiple dual-wavelength laser can be achieved by adjusting PC I to excite equal power components along the two birefringence axes of the Fabry-Pérot filter. Fig. 4.3 (c) plots the optical spectrum of the multiple dual-wavelengths and the inset shows that the power profile is also bell shape.

With the multi-wavelength fibre laser, multi-tap MPF can be constructed [91]. As shown in Fig. 4.1, through the 10:90 coupler and PC III, the output wavelengths are fed to a phase modulator (EOspace PM-0K5-12-PFU-UL). This polarization controller is to align the laser SOP with the axis of the phase modulator. The output phase-modulated optical signal is launched into a 50-km SMF which functions as a dispersive component, and then fed to a photodetector. Finally, the frequency response of the filter is measured by a network analyzer.

The measured filter response is shown in Fig. 4.4 when the multi-wavelength laser is adjusted to the condition in Fig. 4.3 (b). This frequency response has an FSR of 3.52 GHz with full width half maximum (FWHM) of 0.2 GHz and a Q value of around 18. We can see that more than 30 dB out-of-band suppression is achieved.

Fig. 4.4 Measured filter response when the power profile of the multi-wavelength laser is a bell shape.

The filter response is determined by the lasing wavelengths, power profile and total dispersion of the SMF. Through the adjustment of these factors, the filter response can be tuned. Fig. 4.5 shows the response when the laser power profile is changed to the condition in Fig. 4.3 (c), the FWHM of the frequency response broadens from 0.2 GHz to 0.26 GHz.

Fig. 4.5 Measured filter response of the power profile when the laser operates in the multiple dual-wavelength mode.

The power profile of the multi-wavelength laser is an important factor that impacts the MPF frequency response. If the power profile is not bell-shaped, sidelobes appear, as shown in Fig. 4.6.

(a)

(b)

Fig. 4.6 Measurement results (a) the laser power profile is not a bell shape, (b) filter response has sidelobes.

The FSR is determined by the line separation of the multi-wavelength laser and the dispersion of the SMF. So we can tune the FSR by adjusting the length of the SMF. As shown in Fig. 4.7, when we reduce the length of the SMF by half to 25 km, the

FSR increases to 7 GHz which is twice the previous value. The FWHM also doubles to 0.4 GHz. The output of the ring cavity here is the same as that shown in Fig. 4.3 (b).

Fig. 4.7 Measured frequency response when the length of the single mode fibre is reduced to 25 km.

4.2 Modeling of the multi-tap MPF

The multi-tap MPF described in the previous section behaves as an FIR filter due to a finite number of taps that arise from the limited gain bandwidth of the EDFA. According to digital filter (discrete-time electrical filter) design theory, FIR filters are almost entirely restricted to discrete-time implementations and can be designed using the window method [92]. The window method basically involves three steps: 1. taking the inverse Fourier transform of the desired filter transfer function, 2. truncating the sequence to a finite length, and 3. adding a window, such as Blackman, Hamming, Hanning, etc., to reduce the excessive ripple in the passband and poor attenuation in the stopband which result from abrupt discontinuities due to the truncation. The multi-tap MPF design uses a similar concept. The main differences from the design of

digital filters are that in the case of MPFs, there is a need to perform electrical-optical-electrical conversions and to introduce optical time delays between each tap.

A discrete-time filter can be described using Eq. (7.70) in [92]

$$H_d(e^{j\omega}) = \sum_{n=-\infty}^{\infty} h_d[n] e^{-j\omega n} \quad (4.1)$$

where $h_d[n]$ is the impulse response sequence which can be expressed as

$$h_d[n] = \frac{1}{2\pi} \int_{-\pi}^{\pi} H_d(e^{j\omega}) e^{jn\omega} d\omega \quad (4.2)$$

While, for an N-tap MPF, the frequency response is given by [1, 4, 93]

$$H(f) = \sum_{m=0}^{N-1} P_m e^{-jm2\pi f \Delta T} \quad (4.3)$$

where P_m is the tap weight, and ΔT is the time delay difference between two adjacent taps. By comparing Eq. (4.1) and (4.3), we find that the taps in the MPF have fixed time delay in between. This is analogous to the impulse response sequence in discrete-time electrical filters except that the FSR becomes

$$FSR = \frac{1}{\Delta T} \quad (4.4)$$

Therefore, the design of an N-tap MPF can use the concept of windowing with the following steps:

- a. Inverse Fourier transform

Since the window function and impulse response sequence are in time domain, the desired filter response has to be converted to time domain. Take ideal square response as an example, its inverse Fourier transform is a sinc function.

b. Sample the time domain function

The time domain function is represented by a set of discrete wavelengths from a multi-wavelength light source, such as a laser array, a sliced-wideband source or a multi-wavelength laser, as mentioned in the beginning of this chapter. The wavelengths usually have equal spacing. By controlling the power at each wavelength, the power profile of the light source can be made to match the time domain function. For the abovementioned ideal square filter, the power profile can be the multiplication of sinc function with a window function.

c. Time delay difference between taps

To realize the filter response, the ΔT in Eq. (4.3) has to be introduced. ΔT actually refers to the detected RF signals after photodetection. Since in MPFs, electrical-optical-electrical process is used, the time delay of the RF signals is generated by the optical sub-systems. In FIR MPF designs, the ΔT is usually generated by passing the multi-wavelength light through a dispersive medium such as SMF, chirped FBG, etc. Because of the CD, different wavelengths propagate with different speed. So the detected microwave signal corresponding to each optical wavelength has a time delay which is related to the optical frequency (or wavelength). This is the so called frequency-to-time mapping (optical signal in frequency domain to microwave

signal in time domain) [94]. If phase modulation is used, CD also serves the function of phase-to-intensity modulation conversion [76].

After photodetection, the overall response of a multi-tap MPF can be derived by the linear superposition of the detected photocurrent of each tap.

Fig. 4.8 Typical configuration of a multi-tap microwave photonic filter.
EOM: electro-optic modulator; PD: photodetector.

A typical configuration of a multi-tap MPF is shown in Fig. 4.8. It contains mainly a light source and dispersive components corresponding to the abovementioned steps b and c. The multi-wavelength source determines the number of taps and tap weights (window function). The CW light is then modulated by an EOM which is driven by the microwave signal to be filtered. The EOM can be a phase or intensity modulator. Different modulators result in different changes to the filter response as shown later in this section. To introduce a fixed time delay difference between adjacent taps, chromatic dispersion is an effective method. Chirped FBG or SMF can be the candidates as the dispersive medium. After photodetection, each tap of the modulated optical signal is converted back to microwave signal and the interference between the taps leads to the filter response.

For the case of phase modulation together with the use of an SMF, by modifying Eq. (2.14), we have the photocurrent for the n^{th} tap as

$$\begin{aligned}
 i_n(\lambda_n, \omega_m) &= I_n(\lambda_n) RJ_0(m_f) J_1(m_f) \sin\left(\frac{DL\lambda_n^2 \omega_m^2}{4\pi c}\right) \cos[\omega_m t - \varphi_m] \\
 &\propto I_n(\lambda_n) \sin\left(\frac{DL\lambda_n^2 \omega_m^2}{4\pi c}\right) \cos[\omega_m t - \varphi_m]
 \end{aligned} \tag{4.5}$$

where $I_n(\lambda_n)$ denotes the tap weight or optical power of the n^{th} wavelength, $RJ_0(m_f)J_1(m_f)$ is a constant related to the input RF signal power and PD responsivity, and phase delay $\varphi_m = \omega_m D\lambda_n L$ is induced by dispersion.

Since different taps correspond to different wavelengths that travel through a long fibre length, there is no optical interference at the PD. Then the overall detected photocurrent can be summed as

$$\begin{aligned}
 i_{total}(\omega_m) &\propto I_1(\lambda_1) d_{PM,1} \cos(\omega_m t - \varphi_1) + I_2(\lambda_2) d_{PM,2} \cos(\omega_m t - \varphi_2) \\
 &= + \dots + I_n(\lambda_n) d_{PM,N} \cos(\omega_m t - \varphi_N) \\
 &= I_1(\lambda_1) d_{PM,1} \cos \varphi_1 \cos \omega_m t + I_1(\lambda_1) d_{PM,1} \sin \varphi_1 \sin \omega_m t + \\
 &= I_2(\lambda_2) d_{PM,2} \cos \varphi_2 \cos \omega_m t + I_2(\lambda_2) d_{PM,2} \sin \varphi_2 \sin \omega_m t + \\
 &\quad \dots \\
 &= I_N(\lambda_N) d_{PM,N} \cos \varphi_N \cos \omega_m t + I_N(\lambda_N) d_{PM,N} \sin \varphi_N \sin \omega_m t \\
 &= \sum_{n=1}^N [I_n(\lambda_n) d_{PM,n} \cos \varphi_n] \cos \omega_m t \\
 &\quad + \sum_{n=1}^N [I_n(\lambda_n) d_{PM,n} \sin \varphi_n] \sin \omega_m t \\
 &= \sqrt{\kappa_{\cos}^2 + \kappa_{\sin}^2} \cos\left(\omega_m t - \arctan \frac{\kappa_{\cos}}{\kappa_{\sin}}\right)
 \end{aligned} \tag{4.6}$$

where $d_{PM,n} = \sin\left(\frac{DL\lambda_n^2 \omega_m^2}{4\pi c}\right)$, $\kappa_{\cos} = \sum_{n=1}^N [I_n(\lambda_n) d_{PM,n} \cos \varphi_n]$ and

$$\kappa_{\sin} = \sum_{n=1}^N [I_n(\lambda_n) d_{PM,n} \sin \varphi_n].$$

By comparing Eq. (4.6) with the input microwave signal, the filter transfer function is derived as

$$H_{PM}(\omega_m) \propto \sqrt{\kappa_{\cos}^2 + \kappa_{\sin}^2} = \sqrt{\left\{ \sum_{n=1}^N \left[I_n(\lambda_n) \sin\left(\frac{DL\lambda_n^2\omega_m^2}{4\pi c}\right) \cos(\omega_m D\lambda_n L) \right] \right\}^2 + \left\{ \sum_{n=1}^N \left[I_n(\lambda_n) \sin\left(\frac{DL\lambda_n^2\omega_m^2}{4\pi c}\right) \sin(\omega_m D\lambda_n L) \right] \right\}^2} \quad (4.7)$$

Because of the sine term (CD-induced power fading of the phase modulated signal), $|H_{PM}(\omega_m)| = 0$ at DC. This implies that the passband at DC, which usually exists in FIR filters with all-positive taps, is removed. Therefore, the obtained MPF is a bandpass filter and the centre frequency of its first passband is numerically equal to the FSR of the MPF. As usual, the FSR can be calculated as

$$FSR = \frac{1}{T} = \frac{1}{D\Delta\lambda L} \quad (4.8)$$

When an MZM is used, the sine term in Eq. (4.5) needs to be changed to cosine according to Eq. (2.19). As a result, the transfer function becomes

$$H_{IM}(\omega_m) = \sqrt{\left\{ \sum_{n=1}^N \left[I_n(\lambda_n) \cos\left(\frac{DL\lambda_n^2\omega_m^2}{4\pi c}\right) \cos(\omega_m D\lambda_n L) \right] \right\}^2 + \left\{ \sum_{n=1}^N \left[I_n(\lambda_n) \cos\left(\frac{DL\lambda_n^2\omega_m^2}{4\pi c}\right) \sin(\omega_m D\lambda_n L) \right] \right\}^2} \quad (4.9)$$

By comparing Eq. (4.9) with Eq. (4.7), we can see that the term representing the CD induced power fading becomes a cosine function. In this case, the passband at DC stays and the filter response is lowpass.

A similar model has also been established in [28]. However the idea there is different. In [28], each slice of a sliced light source or each output of the multi-wavelength laser is considered as one tap. If the spectrum width of the slice or the linewidth of the laser is wide and cannot be neglected, integration within each tap needs to be done before the summation of all the taps. This will make the calculation of the filter response relatively complex. In our model, we consider each wavelength as one tap for calculation. Then the transfer function will be simply the summation of all taps for narrow linewidth multi-wavelength laser, or integration with respect to the wavelength of the sliced light source or wide linewidth multi-wavelength laser. Our model does not need to divide the input optical spectrum into different parts and integrate separately which makes the calculation simpler with about the same accuracy.

4.3 Tunable multi-tap bandpass microwave photonic filter

Although the MPF in section 4.1 realized bandpass response, the tuning of the passband centre frequency is achieved by a change in the SMF length which is not practical. Therefore the centre frequency tuning is still challenging. For instance, a broad tuning range of about 3 GHz was shown in [31] but the minimum tuning step was 0.58 GHz, while continuous tunability was shown in [33] but the dispersion-based tuning was limited to only 0.35 GHz.

In this section, we propose and demonstrate a multi-wavelength laser-based MPF which has windowed tap weights and exhibits tunability of the centre frequency of the passband over a broad frequency range which is comparable to [31] but offers a much finer resolution [95].

Fig. 4.9 Schematic diagram of the proposed tunable multi-tap MPF. EDFA: Erbium-doped fibre amplifier; HNLF: highly-nonlinear fibre; FP: Fabry-Pérot; PC: polarization controller; EDFA: Erbium doped fibre amplifier; VNA: vector network analyzer; SMF: single mode fibre; PD: photodetector.

The experimental setup is shown in Fig. 4.9. A multi-wavelength laser with a ring cavity, which contains a PC for polarization optimization, an optical isolator for unidirectional operation of the ring, a windowed Fabry-Pérot filter for both wavelength selection and power-profile management, an EDFA (Opto-Link EDFA-MP) for providing optical gain, and a section of highly-nonlinear fibre (HNLF) for FWM to suppress line broadening and increase wavelength stability [88, 96], is used as the optical source for the proposed MPF. Another PC is used to optimize the polarization of the multi-wavelength laser before it is fed into a phase modulator (EOspace PM-0K5-12-PFU-UL). After the light is modulated by an RF signal from a VNA (Anritsu 37369C), it goes through a section of SMF as a dispersive device to realize phase-to-intensity modulation conversion [76] and also frequency-to-time mapping (optical signal in frequency domain to microwave signal in time domain) [94]. The output from the SMF is then fed into a 12 GHz PD (New Focus Model 1544), and the corresponding frequency response is characterized by the VNA.

In the experiments, the drive current of the EDFA is set to 340 mA, and the HNLF has a length of 2 km with a nonlinear coefficient of $10.5 \text{ (W}^{-1} \cdot \text{km}^{-1})^2$. The windowed Fabry-Pérot filter is constructed by programming a WaveShaper (Finisar WaveShaper

4000S) which is actually a reconfigurable wavelength selective switch [97]. The main lobe of the sinc function is chosen for setting the profile of the Fabry-Pérot filter since its Fourier transform approximates a square passband in the frequency domain. A 25 km length of the SMF is used keeping in view the bandwidth constraint of the PD.

(a)

(b)

(c)

Fig. 4.10 Optical and microwave spectra when the multi-wavelength laser has 40 lasing wavelengths: (a) ASE spectrum of the ring cavity without the use of windowed Fabry-Pérot filter, (b) measured multi-wavelength laser output, and (c) measured filter response.

We first measure the ASE spectrum of the ring cavity with the windowed Fabry-Pérot filter disconnected, which is shown in Fig. 4.10 (a). We then set the centre wavelength of the windowed Fabry-Pérot filter to the wavelength with peak power in the ASE spectrum, in order to obtain a relatively symmetric multi-wavelength lasing output. The measured multi-wavelength laser spectrum is shown in Fig. 4.10 (b). More than 40 lasing wavelengths are obtained. Due to the FWM from the HNLF, the powers of the wavelengths at either end of the spectrum reduce to zero gradually and remain stable. Curve fitting shows that the resultant multi-wavelength laser is equivalent to a 23-point Hamming window or a 29-point Blackman window. Thus, abrupt truncation is avoided and no extra window function has to be applied to the sinc function in contrast to the case of digital filter design. The FWM modifies the output multi-wavelength power profile in such a way as to act like a virtual window. The corresponding measured filter response is plotted in Fig. 4.10 (c). Good agreement can be seen between the measurement and calculations using the model in section 4.2. More than 25 dB out-of-band rejection ratio is achieved. In the MPF response, there exists a small peak at 9.65 GHz. This is caused by the asymmetry of the multi-wavelength distribution, as shown in the inset of Fig. 4.10 (b). We further find that the asymmetry is due to the uneven loss of the WaveShaper with respect to wavelength.

It is well-known that the full-width-half-maximum (FWHM) of the FIR filter is determined by the number of taps. Fig. 4.11 (a) shows the measured response when the number of taps is 15, 25, 35, 45; the corresponding FWHM are 0.52 GHz, 0.46 GHz, 0.42 GHz and 0.37 GHz, respectively. Clearly, the more the number of taps we set, the smaller the FWHM or the higher the Q-factor we can realize.

(a)

(b)

Fig. 4.11 Tuning of the filter response. (a) FWHM tuning by varying the number of wavelengths used, with a fixed wavelength spacing of 40 GHz, and (b) tuning of the passband centre frequency by setting the wavelength spacing to 40 GHz, 70 GHz, and 110 GHz, respectively.

As indicated by Eq.(4.8), the centre frequencies of the passbands can be tuned by changing the wavelength spacing with the help of the windowed Fabry-Pérot filter. Fig. 4.11 (b) shows the tuning of the filter response as the wavelength spacing is increased from 40 GHz to 110 GHz, with the number of taps fixed at 25. It is clearly seen that the FSR or the centre frequency of the first passband shifts from 7 GHz to 2.6 GHz. However, the corresponding peak power also reduces. This is because the suppression of the passband at DC also causes attenuation of the passband when close

to DC. Since the sine term in Eq. (4.7) is CD-induced power fading, it can be avoided by using single sideband (SSB) intensity modulation, but this would sacrifice the passband suppression at DC. A centre frequency tuning range of 3 GHz can be achieved if we limit the system to a maximum of 3 dB power drop from the peak. The tuning range can be increased further by increasing the bandwidth of the modulator and PD. The frequency tuning is in steps because the setting of the wavelength spacing of the Fabry-Pérot filter has a minimum resolution of 1 GHz. The corresponding frequency tuning of the passband is 0.175 GHz which is much smaller than the value 0.58 GHz in [31].

4.4 Summary

This chapter focuses on the MPFs that come under the second incoherent operation approach in which the light source coherence time is shorter than the optical delay time. As compared to different types of light sources, a multi-wavelength laser has been chosen because it has a lower price and complexity. Because the number of wavelengths is limited, this kind of filter belongs to finite impulse response multi-tap MPF.

First, a multiple dual-wavelengths Erbium-doped fibre ring laser has been built with stable multi-wavelength output at room temperature. Using phase modulation and an SMF as the dispersive medium, a multi-tap MPF response can be seen. This gave us an intuitive understanding of the FIR MPF working principle.

Building on the understanding of the MPF working principle and the window method of the digital filter design theory, the design procedures and theoretical model for the chromatic dispersion-based multi-tap MPFs have been given.

With the help of the theoretical model, a windowed Fabry-Pérot filter-based multi-wavelength tunable laser has been built. The maximum number of lasing wavelengths is more than 45 with a Blackman or Hamming power profile. By tuning the wavelength spacing of the laser, 3 GHz of the FSR tuning and 25 dB out-of-band rejection ratio have been realized. The measurement results agree well with those calculated from the theoretical model.

Chapter 5 INFINITE IMPULSE RESPONSE

MICROWAVE PHOTONIC FILTER

The previous two chapters described microwave photonic filters with finite number of taps; hence these filters come under the category of finite impulse response (FIR) filters. Another important category of multi-tap filters, the infinite impulse response (IIR) filters, can also be realized using microwave photonic technologies. The IIR MPFs are usually fibre-optic recirculating delay line (RDL) filters. The RDL filters have been demonstrated using a 2x2 coupler with one input and one output connected together to form a time delay loop [38, 98-100]. For an un-amplified 2x2 coupler RDL filter, the coupling ratio has to be adjusted to 33:67 to get optimum notch rejection [98, 99]. To get a high Q value bandpass filter, gain has to be added in the recirculating delay loop [98-100] to form a 2x2 coupler-based amplified RDL filters. If a reflective structure is added, as shown in [101] and indicated in Fig. 5.1, a better performance can be obtained. Such a structure can be considered as a filter cascaded with another filter which is the image of itself produced by the reflective component. Since the image filter is identical to the original one, the random optical interference problem caused by the environmental disturbance can be reduced dramatically compared with the conventional cascaded structures. However, an EDFA has to be added for achieving a good performance.

Fig. 5.1 Topology of the 2x2 reflective amplified RDL filter [101].

In our work, by using a 3x3 collinear optical fibre coupler, a reflective double recirculating delay line (RRDL) MPF is realized which has a higher Q value compared to 2x2 RRDL filter and exhibits both passbands and notches at the same time [102]. By adjusting the time delays of the two loops, the response of the proposed filter can be tailored. Actually, a 3x3 coupler- based filter has been analyzed back in 90's [103, 104] but no experimental demonstrations have been made. To the best of our knowledge, this is the first demonstration of a 3x3 coupler-based IIR filter.

5.1 Configuration and operating principle of a 3x3 coupler-based MPF

The proposed RRDL filter using a 3x3 collinear fibre coupler is shown in Fig. 5.2.

The inputs and outputs of the optical coupler are related by the transmission matrix $\mathbf{T}$

$$\begin{bmatrix} I_4 \\ I_5 \\ I_6 \end{bmatrix} = \mathbf{T} \begin{bmatrix} I_1 \\ I_2 \\ I_3 \end{bmatrix} \quad (5.1)$$

where

$$\mathbf{T} = \begin{bmatrix} A & B & A \\ B & C & B \\ A & B & A \end{bmatrix} \quad (5.2)$$

For a 3x3 collinear fibre coupler, $A = 1/4$, $B = 1/2$ and $C = 0$. If the signal is input from the coupler centre ports 2 or 5, the output power splitting ratio is 50:0:50; if the signal enters from other ports, the power splitting ratio is 25:50:25.

Fig. 5.2 Experimental setup of the reflective recirculating delay line filter. MZM: Mach-Zehnder Modulator, RF: radio frequency.

In Fig. 5.2, CW light from a broadband source (Opto-Link CL15-16ASE, C- and L-Band) is intensity modulated by a MZM (EOSpace AX-0K1-12-PFU-PFU) and is connected to port 2 of a 3x3 coupler through an optical circulator. The optical signal splits equally into two components which appear at coupler ports 4 and 6, as shown in Fig. 5.3, where P_0 is the input signal power, $s_i(t)$ represents the RF signal and the subscript “i = 4, 5 or 6” denotes the port where the signal is considered. When the signal arrives at port 2, it splits into two subcomponents with equal power and appears at port 4 and 6 as $\frac{1}{2}P_0s_4(t)$ and $\frac{1}{2}P_0s_6(t)$. After the first looping, each of the two subcomponents, delayed by T_1 and T_2 , respectively, splits into another 6

subcomponents: $\frac{1}{4} \times \frac{1}{2} P_0 s_4(t-T_1)$, $\frac{1}{2} \times \frac{1}{2} P_0 s_5(t-T_1)$, $\frac{1}{4} \times \frac{1}{2} P_0 s_6(t-T_1)$ from $\frac{1}{2} P_0 s_4(t)$ and $\frac{1}{4} \times \frac{1}{2} P_0 s_4(t-T_2)$, $\frac{1}{2} \times \frac{1}{2} P_0 s_5(t-T_2)$, $\frac{1}{4} \times \frac{1}{2} P_0 s_6(t-T_2)$ from $\frac{1}{2} P_0 s_6(t)$. After second loop, 12 new subcomponents with 3 different time delays, $2T_1$, $T_1 + T_2$ and $2T_2$ are generated. For the n^{th} looping, $n+1$ time delays are generated. The subcomponents appearing at port 5 are reflected back to the coupler by the fibre mirror and repeat the above splitting process. Finally, the subcomponents which reach port 2 of the coupler are received by the photodetector (New Focus Model 1544) and the filter transfer function is measured by a network analyzer (Anritsu 37369C). For deriving the analytical expression for the transfer function, one can use the infinite series indicated in Fig. 5.3. These expressions can be used for calculating the filter response.

Fig. 5.3 Signal splitting in the microwave photonic filter

For comparison, if we disconnect the fibre mirror and use the photodetector to measure the output at coupler port 5, the structure becomes a 3x3 forward RDL (FRDL) filter. Compared with the 2x2 coupler-based RRDL filter, which only generates one subcomponent and one time delay after each loop, the 3x3 FRDL filter generates more subcomponents and new time delays and the 3x3 RRDL generates even more of these. This enables the 3x3 coupler-based IIR RDL filter to have deeper notches than a 2x2 coupler-based filter.

Since there are two RDLs in Fig. 5.2, we can define z_1 and z_2 corresponding to the two different time delay T_1 and T_2 , respectively. Based on the subcomponent splitting graph in Fig. 5.3, the total power P output from port 5 of the 3x3 coupler is

$$P = P_0 \times \left[\begin{array}{l} \left(\frac{1}{2}\right)^2 z_1^{-1} + \left(\frac{1}{2}\right)^2 \frac{1}{4} z_1^{-2} + \left(\frac{1}{2}\right)^2 \left(\frac{1}{4}\right)^2 z_1^{-3} + \left(\frac{1}{2}\right)^2 \left(\frac{1}{4}\right)^3 z_1^{-4} + \left(\frac{1}{2}\right)^2 \left(\frac{1}{4}\right)^4 z_1^{-5} + \dots \\ + \left(\frac{1}{2}\right)^2 z_2^{-1} + \left(\frac{1}{2}\right)^2 \frac{2}{4} z_1^{-1} z_2^{-1} + \left(\frac{1}{2}\right)^2 3 \left(\frac{1}{4}\right)^2 z_1^{-2} z_2^{-1} + \left(\frac{1}{2}\right)^2 4 \left(\frac{1}{4}\right)^3 z_1^{-3} z_2^{-1} + \left(\frac{1}{2}\right)^2 5 \left(\frac{1}{4}\right)^4 z_1^{-4} z_2^{-1} + \dots \\ + \left(\frac{1}{2}\right)^2 \frac{1}{4} z_2^{-2} + \left(\frac{1}{2}\right)^2 3 \left(\frac{1}{4}\right)^2 z_1^{-1} z_2^{-2} + \left(\frac{1}{2}\right)^2 6 \left(\frac{1}{4}\right)^3 z_1^{-2} z_2^{-2} + \left(\frac{1}{2}\right)^2 10 \left(\frac{1}{4}\right)^4 z_1^{-3} z_2^{-2} + \dots \\ + \left(\frac{1}{2}\right)^2 \left(\frac{1}{4}\right)^2 z_2^{-3} + \left(\frac{1}{2}\right)^2 4 \left(\frac{1}{4}\right)^3 z_1^{-1} z_2^{-3} + \left(\frac{1}{2}\right)^2 10 \left(\frac{1}{4}\right)^4 z_1^{-2} z_2^{-3} + \dots \\ + \left(\frac{1}{2}\right)^2 \left(\frac{1}{4}\right)^3 z_2^{-4} + \left(\frac{1}{2}\right)^2 5 \left(\frac{1}{4}\right)^4 z_1^{-1} z_2^{-4} + \dots \\ + \left(\frac{1}{2}\right)^2 \left(\frac{1}{4}\right)^4 z_2^{-5} + \dots \end{array} \right] \quad (5.3)$$

where $z_1 = \exp(j2\pi T_1 f_{rf})$ and $z_2 = \exp(j2\pi T_2 f_{rf})$.

In Eq.(5.3), the n^{th} column represents all the subcomponents coming out from port 5 after circulating for n rounds. It can be found that each column is actually the expansion of the summation of two terms with power n , such as the third column

$$\left(\frac{1}{2}\right)^2 \left(\frac{1}{4}\right)^2 z_1^{-3} + \left(\frac{1}{2}\right)^2 3 \left(\frac{1}{4}\right)^2 z_1^{-2} z_2^{-1} + \left(\frac{1}{2}\right)^2 3 \left(\frac{1}{4}\right)^2 z_1^{-1} z_2^{-2} + \left(\frac{1}{2}\right)^2 \left(\frac{1}{4}\right)^2 z_2^{-3} \text{ is actually } \left[(4z_1)^{-1} + (4z_2)^{-1} \right]^3$$

By combining each column in Eq. (5.3) together, we have

$$P = P_0 \times \left\{ \left[(4z_1)^{-1} + (4z_2)^{-1} \right]^1 + \left[(4z_1)^{-1} + (4z_2)^{-1} \right]^2 + \left[(4z_1)^{-1} + (4z_2)^{-1} \right]^3 + \left[(4z_1)^{-1} + (4z_2)^{-1} \right]^4 + \dots \right\} \quad (5.4)$$

The series in Eq. (5.4) is a geometric series. By simplifying it and comparing the output with the input, we have the transfer function for the FRDL as

$$H_{forward} = \frac{(4z_1)^{-1} + (4z_2)^{-1}}{1 - \left[(4z_1)^{-1} + (4z_2)^{-1} \right]} \quad (5.5)$$

Because of the symmetry of the 3x3 coupler, when the reflective configuration is used, all the signals (terms in Eq.(5.3)) go through the same structure once again. So the resultant filter response of the RRDL is

$$H_{forward}(z) = H_{forward}^2(z) = \left(\frac{(4z_1)^{-1} + (4z_2)^{-1}}{1 - \left[(4z_1)^{-1} + (4z_2)^{-1} \right]} \right)^2 \quad (5.6)$$

5.2 Experimental results and discussion

In the experiments, a wide band optical source is used such that the source coherence time is much shorter than the delay time of the shorter loop to fulfill the condition of the incoherent operation. The MZM is biased at quadrature point. In order to achieve incoherent summation of the subcomponents of the ‘same order of circulation’, two polarization controllers are applied in the two loops to adjust the polarization state of the subcomponents to be orthogonal at the photodetector when they come from different loops.

The filter response for different time delay ratios between T_1 and T_2 has been measured. When the ratio is 1:1, the filter transfer function has no zero. So the notch rejection for 3x3 FRDL and RRDL are only 10 dB and 18 dB as shown in Fig. 5.4 (a) and (b).

Fig. 5.4 Measured and calculated filter response for the 3x3 (a) forward recirculating delay line filter and (b) reflective recirculating delay line filter, when the time delay ratio is 1:1.

When we change the time delay ratio to 1:2, the notch rejection of the 3x3 FRDL filter becomes more than 40 dB and the full width half maximum (FWHM) is smaller for the 3x3 RRDL filter as shown in Fig. 5.5. In Fig. 5.5, the measured filter response has two small shoulders at both sides which are not there in the calculated response.

These may be caused by the imperfect splitting ratio of the collinear 3x3 coupler. For the RRDL filter, the signals pass through the coupler two times, so the mismatch between the calculation and the measurement for the RRDL filter is bigger than that for the FRDL filter.

Fig. 5.5 Measured and calculated filter response for the 3x3 (a) forward recirculating delay line filter and (b) reflective recirculating delay line filter, when the time delay ratio is 1:2.

Fig. 5.6 Calculated filter response of the 2x2 RRDL filter and the measured filter response of the 3x3 RRDL filter for time delay ratios 1:2 and 1:1. RRDL: reflective recirculating delay line

Comparing the measured response for the 3x3 RRDL filter with that for a 2x2 un-amplified RRDL filter calculated according to [101], the proposed 3x3 RRDL filter has much deeper notch rejection when the time delay ratio is 1:2 as shown in Fig. 5.6. Even when the ratio is 1:1, the notch rejection for the 3x3 RRDL filter is still deeper than that for the 2x2 RRDL filter.

Fig. 5.7 Measured filter response for 3x3 RRDL filter and FRDL filter when the time delay ratio is 1:1.16. RRDL: reflective recirculating delay line; FRDL: forward recirculating delay line.

Fig. 5.7 shows the results when the time delay ratio is 1:1.16, i.e., when T_2 is not an integral multiple of T_1 . The overall response is a superposition of two responses: the FSR of the two adjacent peaks corresponds to the smaller of the time delay T_1 and T_2 while the FSR of the big envelope corresponds to the time difference between T_1 and T_2 . For this case, the notch rejection for the 3x3 RRDL filter is much deeper than that for the 3x3 FRDL filter; the 2x2 RRDL filter does not have this kind of response. In our setup, the length of the shorter delay loop is 2.441 m, which limits the FSR to 83.6 MHz. The long loop length is due to the fibre pigtails of polarization controllers and the 3x3 coupler and can be eliminated by implementing the filter in a waveguide structure.

5.3 Summary

In this chapter, we have demonstrated an IIR MPF using a 3x3 collinear coupler reflective double recirculating delay line structure. Different filter responses can be obtained by changing the loop length ratio and the measured results basically match with the theoretical model. When the ratio is 1:2, more than 40 dB notch rejection is achieved. This filter shows higher Q values compared to a 2x2 RRDL filter and exhibits both passbands and notches at the same time.

Chapter 6 INSTANTANEOUS FREQUENCY

MEASUREMENT USING PHOTONIC TECHNIQUES

6.1 Introduction

The demand for instantaneous measurement of microwave frequency comes from the defense applications. In many radar and other electronic warfare systems, instantaneous frequency measurement (IFM) of a microwave signal is required to enable scanning, identification and analysis of the microwave signal over a large frequency range with a high probability of interception. In these systems, a number of specialized receivers are jointly employed to reduce the processing load of a single receiver. Therefore, the carrier frequency of a microwave signal is needed to be measured instantaneously using an IFM receiver before passing it to a specialized receiver for further processing.

Based on the scheme of realization, the current photonic assisted microwave measurement techniques can be divided into three categories: scanning receiver, optical channelizer receiver, and IFM receiver. Using a scanning Fabry-Pérot etalon to scan the MZM modulated carrier-suppressed double sideband modulation signal, the microwave signal frequency can be obtained [39]. An Echelle diffractive grating can also be used to scan the centre wavelength of the upper or lower sideband, so that the microwave signal frequency can be calculated through the wavelength difference between the optical carrier and sideband [40]. Since the scanning needs a certain time span, this kind of receivers are not suitable for measuring pulsed microwave signals.

For pulsed signals, both optical channelizer and IFM receivers can be utilized. The key component for the optical channelizer receiver is a diffraction grating, an arrayed waveguide grating, or a diffraction grating based Fabry-Pérot etalon [41-43]. After the narrowband optical carrier is modulated by the unknown microwave signal, sidebands are generated which correspond to the microwave frequency. When this modulated optical signal propagates through one of the abovementioned key components, because of the diffraction or interference, the signals with different frequencies will be separated in space and enter into different detector units in a detector array. So the microwave frequency can be estimated using the diffraction angle and detector location. From the working principle of the optical channelizer it is clear that this kind of receivers can measure multiple microwave signals at the same time. However, the measurement bandwidth and resolution are limited (20 GHz and 1 GHz, respectively) by the diffraction or interference component, and size and number of the detectors.

Photonics-based IFM techniques can be further divided into two types: frequency-to-time mapping and frequency-to-power mapping. A frequency-to-time mapping receiver can also measure multiple signals [44, 45]. It is based on carrier-suppressed double sideband modulation. After the modulated signal passes through a dispersive medium, the upper and lower sidebands will have a time delay difference because of different group velocities. When measuring this signal using a PD, a step-like power-time curve can be observed. The step width is decided by the time delay difference. Since the dispersion and modulation frequency decide the time delay difference, the frequency of the modulating microwave signal can be estimated accordingly. For this technique, a real time high speed oscilloscope is required to measure the optical power in time domain, so the system cost is very high.

Recently, frequency-to-power mapping based IFM receivers have attracted considerable interest because these have the advantages of relatively simpler configuration, higher resolution and accuracy, and insensitivity to the type of modulation of the pulsed signal under measurement. Although this kind of receiver can only measure one input signal at one time, it can be combined and used with other types of receivers to mitigate this drawback [11].

Fig. 6.1 Schematic of an electronic instantaneous frequency measurement receiver

The schematic of a typical electronic IFM receiver is shown in Fig. 6.1 [14, 15]. It generates two out-of-phase signals and calculates the frequency through the comparison of these two signals. The microwave photonic frequency-to-power mapping IFM receiver also uses a similar comparison. A typical scheme for such a receiver was introduced in Section 1.1.2. The basic concept is to construct an ACF which is the ratio of two different optical or microwave power functions. In [16], two different dispersion induced power fading functions were used. ACF establishes the relation between the input microwave signal frequency and the output power. From this relation, a calibrated lookup table can be established for estimating the microwave frequency from the measured power. This principle is actually the same as the CD monitoring in a fibre link [105]. Similar techniques can also be found in [46-49], however, these techniques either use two laser sources [46, 47, 49] or use a special modulator [48], both of which increase the system cost and complexity. In order to simplify the receiver structure, we propose an IFM configuration using an

asymmetric non-linear group delay profile with only one laser source and one modulator [106].

6.2 Instantaneous microwave frequency measurement using an asymmetric non-linear group delay profile

Fig. 6.2 shows the experimental setup for verifying and demonstrating the principle of the proposed scheme. A tunable laser source (Anritsu MG9638A) is used to generate CW light with a 100-kHz linewidth. A 40-GHz broadband MZM (Avanex SD-40) is driven by an unknown microwave signal from an RF signal generator. A DC bias is applied to bias the MZM at its quadrature point. The modulated optical signal is then split into two parts by a 3 dB coupler. Each part is launched into one of the ports of a NLCFBG through two optical circulators. The modulated optical signal is then split into two parts by a 3 dB coupler. Each part is launched into one of the ports of a NLCFBG through two optical circulators.

Fig. 6.2 Schematic diagram of the photonic microwave frequency measurement system using NLCFBG. PD: photodetector.

The NLCFBG used here is the same as the one described in section 3.4. Because of the second order pitch function as in Eq.(3.9), it has a linear CD variation over the Bragg wavelengths, as shown in Fig. 6.3 (a). In this figure we can see that the CD values are different when an optical signal of a particular wavelength is launched into different ports of the NLCFBG. The difference of CD values increases when the operation wavelength is close to the edges of the reflection band. In comparison, the

measured reflection spectrum and calculated dispersion of a LCFBG is plotted in Fig. 6.3 (b). In this case, the CD value is constant within the reflection band.

Fig. 6.3 Measured reflection spectrum and calculated dispersion of (a) nonlinearly chirped fibre Bragg grating and (b) linearly chirped fibre Bragg grating when the optical input is from different ports.

In our frequency measurement configuration, DSB is used. When the modulated optical signal is reflected by the NLCFBG, the CD of the NLCFBG causes a certain time delay difference between the upper and lower sidebands of the modulated optical signal. This results in power fading of the microwave signal at the output of the two PDs. If we assume that the optical loss of the two arms and the responses of the PDs are identical, the ACF can be expressed as

$$ACF = \frac{\cos^2\left(\pi D_{long} \lambda_c^2 f^2 / c\right)}{\cos^2\left(\pi D_{short} \lambda_c^2 f^2 / c\right)} \quad (6.1)$$

where f is the frequency of the unknown microwave signal, D_{short} and D_{long} denote the dispersion of the NLCFBG when the optical signal is launched into short and long wavelength ports, respectively; λ_c and c represent the wavelength of optical carrier and the speed of light in vacuum, respectively. Using Eq. (6.1), the RF frequency can be calculated from the measured ACF.

Fig. 6.4 Measured frequency response for input from the short and long wavelength ports of (a) nonlinearly chirped fibre Bragg grating and (b) linearly chirped fibre Bragg grating

A proof-of-concept experiment is carried out. The responses for the two arms were measured by connecting the input and output ports of a vector network analyzer to the PDs and MZM, respectively. The measured response of the NLCFBG is shown in Fig. 6.4 (a). Since the dispersion curves for the short and long wavelength ports are different, the corresponding frequency responses are also different. For comparison, the NLCFBG in Fig. 6.2 was replaced by a linearly chirped FBG; the measurement results in Fig. 6.4 (b) show that in this case the responses for the two ports are almost the same and hence cannot be used for frequency measurement.

The measured and calculated ACF using NLCFBG for an optical wavelength of 1550 nm, are shown in Fig. 6.5 (a). Good agreement is observed. In Fig. 6.5 (a), we can see that the ACF value varies between two extremes located at 8.22 GHz and 9.82 GHz. This range is chosen as the microwave frequency measurement range because the steep slope of ACF gives better measurement accuracy. Based on Eq. (6.1), a look-up table is set up and the microwave frequency is estimated using the measured ACF. Fig. 6.5 (b) shows good agreement between the estimated and input frequency. The measurement errors are less than ± 0.08 GHz as shown in Fig. 6.5 (c). The measurement range is determined by the separation of the two extreme values which is decided by the carrier wavelength. After changing the carrier wavelength to 1550.15 nm, the corresponding measurement range is reduced by half; this agrees with the calculated CD variation in Fig. 6.3.

(a)

(b)

(c)

Fig. 6.5 (a) Measured and calculated ACFs; (b) estimated frequency as a function of input frequency, and (c) measurement error.

6.3 Instantaneous microwave frequency measurement using phase and intensity modulators

In [46-49] and our work in the previous section, the measurement range for high measurement resolution is limited to a few GHz in the vicinity of the lower frequency notch. To get high measurement accuracy while maintaining a wide measurement range, the laser wavelength has to be tuned or multiple wavelengths need to be used. Still, it is difficult to perform an accurate measurement at low frequencies because of a small ACF slope.

To achieve a wider measurement range without the need for wavelength tuning, two complementary DC voltage [50] or power functions [51, 107-110] were used to construct the ACF, to achieve an ACF with infinite power variation or a large slope over the entire range of measurement. The major limitation of these approaches is the high system complexity: multiple laser sources, and/or other passive components must be used. The use of multiple laser sources may have problems such as wavelength spacing drift and relative power fluctuations, which may increase the measurement error [50, 51, 107]. In [109, 110], the MZM was biased to suppress the optical carrier, but an incomplete carrier suppression would also contribute to measurement errors. In order to achieve a more stable measurement, we propose in the following an IFM configuration using a pair of phase and intensity modulators.

Fig. 6.6 Schematic diagram of the IFM receiver with two modulators. SMF: single mode fibre; PD: photodetector.

The system configuration for the proposed photonic microwave frequency measurement is shown in Fig. 6.6 [111]. A distributed feedback (DFB) laser source (YOKOGAWA AQ2200-111) operating at 1550 nm is used to generate CW laser with a linewidth of 5 MHz. The linearly polarized output light is split into two parts by a PM coupler and sent to an x-cut MZM (Avanex SD40) and an x-cut phase modulator (EOspace PM-0K5-12-PFU-UL), respectively. The PM coupler is used to ensure the alignment of the linear polarization of CW light with the axes of the subsequent modulators. The modulators are driven by the unknown RF signal and the output modulated signals are introduced into standard SMFs (G.652) with dispersion coefficient of 17 ps/nm/km. The MZM is biased at quadrature point to generate DSB modulation. After passing through the SMFs, the microwave powers are measured by two identical PDs with a bandwidth of 20 GHz. Because of the CD of the SMF, the RF signal experiences power fading.

By using Eq. (2.15) and (2.19), the power ratio between the two detected powers, referred to as the amplitude comparison function (ACF), is obtained as

$$ACF = \frac{P_{PM}}{P_{IM}} = \frac{R_2 \sin^2 \left(\frac{\pi D L_2 \lambda_c^2 f^2}{c} \right)}{R_1 \cos^2 \left(\frac{\pi D L_1 \lambda_c^2 f^2}{c} \right)} \quad (6.2)$$

where f is the frequency of the unknown microwave signal, $L_i, i=1,2$, represent the lengths of the SMFs for the upper and lower arms, D denotes the dispersion coefficient of the SMF, λ_c and c represent the wavelength of optical carrier and the speed of light in vacuum, and R_i are the total losses of the two arms including the insertion loss of coupler, modulators, SMFs and the responsivity of the PDs.

Through calibration and post-processing we can obtain $R_1 = R_2$. If the SMF lengths in the two arms are also chosen to be equal, the ACF further simplifies as

$$ACF = \tan^2 \left(\frac{\pi D L \lambda_c^2 f^2}{c} \right) \quad (6.3)$$

Using Eq. (6.3), the RF frequency can be calculated from the measured ACF. In Fig. 6.6, post-processing is required to determine ACF. This can be achieved by passing the detected RF signals through detector log-arithmetic video amplifiers and taking the difference between the outputs. Alternatively, one can use A/D convertors after PDs and calculate ACF.

Fig. 6.7 Calculated ACF for different single mode fibre lengths

Fig. 6.7 shows the plots of analytically calculated ACF as a function of input microwave frequency for different SMF lengths, for bandwidths of 20, 30 and 40 GHz. It is clearly seen that the bandwidth of microwave frequency measurement is determined by the length or the total CD of the SMF. Also, smaller lengths of SMF produce an ACF with a wider bandwidth. The measurement bandwidth can be extended by reducing the length of the SMF.

In Eq. (6.3), the fibre lengths of the upper and lower arms are assumed to be the same. In practice, some length difference between the upper and lower arms always exists. Fig. 6.8 (a) shows the comparison between the conditions of equal length and that of 2 meters length difference. The two ACF curves almost overlap; there is a slight difference only at the high frequency end, as shown in the inset. Fig. 6.8 (b) gives the measurement error when the 2 m length difference is taken into account. A maximum error of 0.017 GHz is observed; this corresponds to 0.0425% of the microwave frequency. Therefore, the proposed method is tolerant to several meters of SMF length difference between the two arms. Actually, a commercially available optical time domain reflectometer can easily measure the SMF length with a resolution of 1 meter. Further, the SMF length difference induced measurement error reduces with

the reduction of the measurement bandwidth. For instance, for 20 GHz bandwidth, the tolerance of fibre length difference can be increased to 15 meters for the error level mentioned earlier.

Fig. 6.8 (a) Comparison between the conditions of equal length single mode fibres with that of 2 meter length difference; (b) the corresponding measurement error.

The drift of the operating wavelength can also affect the ACF values. Simulation results show that the effect of 0.01 nm wavelength variation (which is indicated in the specifications of our laser source) is negligible.

A proof-of-concept experiment has been carried out using 20 km SMFs for the two arms keeping in view the bandwidth constraints of the phase modulator and PDs. The

corresponding responses for the two arms were measured by connecting the input and output ports of a VNA (Anritsu 37369C) to the PDs and modulators, respectively. The measured responses, measured ACF, and calculated ACF are shown in Fig. 6.9. Excellent agreement is observed. The frequency of the microwave signal estimated through measurement and the error with respect to the actual input frequency are presented in Fig. 6.10 (a) and (b), respectively. It is observed that the proposed approach yields a measurement error less than ± 0.3 GHz. The error sources contributing to this can be the noise of the PDs and the bias drift of the MZM which modifies the ACF. The measurement frequency range is also wide, extending from close to 0 GHz to 13.8 GHz. The measurement range can be further extended by reducing the length of SMFs and increasing the bandwidths of the modulators and PDs accordingly.

Fig. 6.9 Measured and calculated ACFs

Fig. 6.10 (a) Estimated frequency as a function of the input frequency; (b) measurement error vs. the input frequency.

6.4 Instantaneous microwave frequency measurement based on phase modulation

Although the CD based IFM in section 6.3 is quite stable, the use of multiple modulators may induce noise because the Lithium Niobate modulators are sensitive to the bias voltage and ambient temperature change. To improve the performance and

reduce the cost, we propose another IFM receiver which uses only one phase modulator [112].

Fig. 6.11 Schematic diagram of the IFM receiver based on phase modulation. PMF: polarization maintaining fibre; DCF: dispersion compensation fibre; PD: photodetector.

The experimental setup for verifying and demonstrating the proposed scheme is shown in Fig. 6.11. A DFB laser source is used, similar to that in the previous section. The linearly polarized output light is sent to a 12 GHz phase modulator (EOspace PM-0K5-12-PFU-UL) through a half-wave plate. The pigtails of the laser source and the phase modulator are made of PMF. By adjusting the axis of the half-wave plate to 45° with respect to the slow axis of the PMF pigtails, two orthogonally polarized CW light signals, polarized along the slow and fast axes of the PMF, are excited equally and introduced to the subsequent phase modulator. The modulator is driven by the unknown microwave signal. After phase modulation, two orthogonally polarized out-of-phase optical signals are generated [113, 114] which carry the frequency information of the unknown microwave signal. A polarization maintaining coupler splits these two optical signals into two parts with the same power and polarization. In the upper arm, a polarizer is placed with its polarization angle set at 45° with respect to the slow axis of the PMF. This optical signal passes through the PMF and is detected by the PD. The detected signal exhibits a low-pass frequency response due to

the power fading induced by DGD [22, 93]. This frequency response can be expressed as (Appendix A)

$$H_L(f) = \cos^2(\pi f \Delta \tau) \quad (6.4)$$

In Eq. (6.4), f is the modulating microwave frequency, and $\Delta \tau$ is the DGD value of the PMF.

In the lower arm, the polarization axis of the polarizer is aligned with the slow axis of the PMF. So, only the optical signal along the slow axis can pass through the polarizer. This signal is transmitted through the dispersion compensation fibre (DCF). The detected signal in this case exhibits bandpass frequency response due to typical power fading induced by chromatic dispersion as expressed by Eq. (2.15). So the ACF between the detected microwave powers from the two arms is expressed as

$$\text{ACF} = \frac{H_B}{H_L} = \frac{\sin^2\left(\frac{\pi D \lambda_c^2 f^2}{c}\right)}{\cos^2(\pi f \Delta \tau)} \quad (6.5)$$

Fig. 6.12 Power fading characteristics of the signals for the two arms in Fig. 6.11 and the corresponding ACF

The calculated individual power fading response of the two arms and the corresponding ACF are plotted in Fig. 6.12, under the conditions of $\lambda_c = 1550$ nm, $D = 292$ ps/nm, and $\Delta\tau = 41$ ps. We can see that the frequency measurement range starts at a relatively low frequency and similar to section 6.3, the ACF varies from negative infinity to positive infinity on a log scale; this ensures high measurement accuracy. The upper limit of the measurement range is determined by the notch position of the PMF induced low-pass response which is given by $BW = \frac{1}{2\Delta\tau}$. To achieve a high resolution measurement, H_B and H_L need to monotonically increase and decrease, respectively, within the BW. Thus the maximum transmission frequency of H_B which is given by $H_{B,\max} = \sqrt{\frac{c}{2D\lambda_c^2}}$, needs to be larger than the BW. Finally, the microwave frequency can be calculated from the measured ACF once the lengths of the DCF and PMF are fixed. Since the two frequency responses have notches at high frequency and zero frequency, respectively, the ACF varies monotonically from negative infinity to positive infinity. This ensures a steep change in ACF with respect to frequency over a wider frequency range. By varying the lengths of the dispersion compensation fibre (DCF) and PMF, the measurement range can easily be extended further.

Fig. 6.13 Measured power fading functions and measured as well as calculated ACF

A proof-of-concept experiment has been performed using a 32 m long PMF and a 4.1 km long DCF, keeping in view the bandwidth constraints of the phase modulator and PD. The corresponding responses were measured by connecting the input and output ports of a VNA (Anritsu 37369C) to the PD and phase modulator, respectively. The measured ACF agrees well with the one calculated theoretically, as shown in Fig. 6.13. Based on Eq. (6.5), a look-up table is set up and the microwave frequency was estimated using the measured ACF.

Fig. 6.14 (a) Estimated frequency as a function of input frequency; (b) measurement error vs. the input frequency.

The measurement results in Fig. 6.14 (a) and (b) show that the proposed approach yields high accuracy over a frequency range extending from 1.7 GHz to 12.2 GHz.

Over this frequency range, the measurement error remains within ± 0.07 GHz which is much smaller than the value ± 0.3 GHz in section 6.3.

6.5 Instantaneous microwave frequency measurement using a microwave photonic filter with an infinite impulse response

The IFM receiver in section 6.3 uses one laser source and two modulators. In section 6.4, the number of modulators is also reduced to one. Although the shape of the ACF is similar in these cases, the measurement error for the scheme in section 6.4 is greatly reduced. However, the major limitation is that it still requires two PDs for getting the ACF and the noise of the PDs contributes to the measurement errors.

In this section, we propose a further improvement in the photonic approach for microwave frequency measurement by using a microwave photonic filter with an IIR filter [115]. It is different from the previous approaches in that only one filter response is measured, so that only one PD is required. This lowers the expected measurement error; the complexity and cost of the measurement system are also considerably reduced. Moreover, theoretically the power variation of the response is infinite which is similar to the ACF in sections 6.3 and 6.4; therefore, the measurement resolution can be maintained.

The system configuration for the proposed photonic microwave frequency measurement is shown in Fig. 6.15. It consists of an electrical feedback IIR filter cascaded with a FIR filter. Linearly polarized CW light from a LD is sent to a dual-drive Mach-Zehnder modulator (DD-MZM) which is driven via one RF port by a microwave signal for which the frequency is to be measured. The other RF port is

connected to the output of the PD to form a recirculating loop. The net gain of the loop is adjusted to avoid oscillations.

Fig. 6.15 Schematic diagram of the IFM receiver based on IIR filter. LD: laser diode; FIR: finite impulse response; IIR: infinite impulse response; PD: photodetector.

Fig. 6.16 Block diagram of the infinite impulse response filter

With the help of the block diagram in Fig. 6.16, the amplitude transfer function, without considering the FIR filter, can be derived using z transform

$$H_1(z) = \frac{LRk}{1 - (1-k)Gz^{-1}} \quad (6.6)$$

where L is the total loss including the insertion loss of the RF power divider, the modulator, and the optical fibres, R is the responsivity of the PD, G is the gain of the RF amplifier, and k is the split ratio of the RF power divider. This transfer function has one pole and no zero as shown in Fig. 6.17.

Fig. 6.17 Calculated frequency response of H_1 , H_2 , and H for $k=0.5$, $L=1$ and $G=1.95$

To have an infinite power variation, a two-tap FIR notch filter, formed by two 3-dB optical couplers, is added to introduce a zero in the transfer function. Its amplitude transfer function is

$$H_2(z) = 1 + z^{-1} = \frac{1+z}{z} \quad (6.7)$$

Then the total amplitude transfer function of the cascaded IIR and FIR filters is

$$H(z) = H_1(z) \cdot H_2(z) = \frac{LRk(1+z)}{z - (1-k)G} \quad (6.8)$$

In deriving Eq. (6.8), the time delay difference between the two arms of the FIR filter is set equal to the loop time delay of the IIR filter. The total transfer function H is also plotted in Fig. 6.17. We can see that, on a log scale, it has infinite power variation, ensuring a high measurement resolution.

With $z = \exp(j2\pi Tf)$, Eq. (6.8) can be rewritten in frequency domain as

$$H(f) = \frac{LRk[1 + \exp(j2\pi Tf)]}{\exp(j2\pi Tf) - (1-k)G} \quad (6.9)$$

where T is the loop time delay and f is the RF frequency. As seen in Fig. 6.17, the monotonic region of the total transfer function from DC to the first notch can be used as the frequency measurement range. Based on Eq. (6.9), a calibrated look-up table can be established to extract the frequency of the input microwave signal from the output power of the RF coupler.

From Eq. (2.18) - (2.19) we know that the output power is actually related to the power of the input microwave signal

$$P_{out} \propto J_0^2(m) J_1^2(m) |H(f)|^2 \quad (6.10)$$

To ensure that the microwave frequency has a unique relationship with the output power, the input signal power should be normalized. This can be done by tapping a small amount of input power and using the tapped signal to calculate the input microwave power related $J_0^2(m) J_1^2(m)$ in the post-processing in Fig. 6.15.

A proof-of-concept experiment is carried out for the configuration in Fig. 6.15. Light wave at 1550 nm generated by a tunable laser source (Anritsu MG9638A) is sent to the dual port MZM (Fujitsu FTM7921ER 10 Gb/s). The IIR filter uses electrical feedback, so it is coherence free. To make the FIR filter also work incoherently, the coherence control of the laser source is turned on. The resultant laser linewidth is 50 MHz so that the laser coherence length is shorter than the FIR filter arm length difference.

In deriving Eq.(6.8), the time delays of the two filters are assumed identical. Although the time delay of the IIR filter is changed when the FIR filter is incorporated, only the shorter arm of the FIR filter would contribute to the time delay of the IIR filter.

Therefore, in the experiment we can match the two time delays by adjusting a tunable optical delay line that is inserted in the longer arm of the FIR filter.

A microwave signal generated by a VNA (Agilent E8364A) is applied to the MZM. The measured transfer function is shown in Fig. 6.18. The transfer function calculated based on Eq. (6.9) is also shown. A good agreement is observed. Because of the specifications of the available components, such as the RF amplifier (Avantek SA82-0431) which has a bandwidth from 8 GHz to 18.2 GHz, together with the bandwidths and loss profiles of the other components, the maximum RF gain occurs around 6.9 GHz.

Fig. 6.18 Measured system transfer function

The estimated frequency of the microwave signal and the error with respect to the actual input frequency is presented in Fig. 6.19 (a) and (b). The frequency measurement range chosen in our experiment is from 6.9088 to 6.9190 GHz. Note that since the effective loop length of the IIR filter including the optical fibre pigtails and RF cables is very long (9.34 m), a FSR of 21.8 MHz is observed. For practical applications, the loop length should be reduced, leading to an increased FSR thus an increased measurement range. Considerable reduction in loop length can be achieved

using integrated solutions. For instance, a 10 GHz measurement range can be realized with 1 cm effective loop length by using integrated optics with an EAM modulator.

Fig. 6.19 (a) Estimated frequency as a function of the input frequency; (b) measurement error vs. the input frequency.

The measurement resolution can be characterized by using the first-order derivative of the transfer function (in dB) as $d[H(\text{dB})]/df$. Based on our calculation, the minimum resolution of the proposed configuration is 2.67 dB/GHz for a 10 GHz measurement range, which is much higher than the resolution of 0.19 dB/GHz for a 6.5 GHz range in [46].

6.6 Summary

Generally, the frequency-to-power mapping based IFM receivers construct an ACF which is the comparison of two different transfer functions to convert the frequency information to the power information. In this respect, IFM can be considered as an important application of microwave photonic filters.

Several IFM receiver designs have been proposed keeping in view the measurement accuracy and the system cost and complexity as shown in Table 6.1. The NLCFBG based IFM receiver utilizes the asymmetric non-linear group delay profile of the NLCFBG. This receiver only needs one tunable laser source and one MZM, so the cost is low. Similar to other early IMF receivers, two low pass responses have been used. So, the steep variation of the ACF is still limited from 8.22 GHz to 9.82 GHz in the vicinity of the first notch frequency and cannot cover low frequencies.

Table 6.1 Comparison between frequency-to-power mapping based IFM receiver designs

IFM	Principle	No. of Lasers	No. of EOMs	No. of PDs	Meas. bandwidth (GHz)	Meas. error (GHz)
Ref. [16]	CD induced power fading	2	1	2	8-12 (tunable)	Not given
IFM 1	NLCFBG	1	1	2	8.22-9.82	± 0.08
Ref. [50]	Photonic Hilbert transform	3	2	2	1-10	Not given
Ref. [51]	RF mixing	2	2	2	4-19	Not given
Ref. [107]	Optical filter and power measurement	2	1	2	1-20	± 0.2
IFM 2	Phase and intensity modulation	1	2	2	0.5-13.5	± 0.3
IFM 3	Phase modulation	1	1	2	2-12	± 0.07
IFM 4	IIR response	1	1	1	6.90-6.92	± 0.25 (MHz)

To obtain an ACF that also covers the low frequency range, two complementary transfer functions are required. This usually requires multiple laser sources and/or

modulators. We have found that by using the power fading of the phase and intensity modulated signals, the number of laser source can be reduced to one and the ACF is simply the square of a tangent function. This property gives a potential advantage of fast calculation of the microwave frequency using inverse function rather than using an ACF look-up table. The proof-of-concept experiment shows that our design can achieve a wide measurement range from DC to 13.8 GHz with a measurement error less than ± 0.3 GHz.

The above phase and intensity modulators based IFM has relatively large measurement error because the Lithium Niobate modulators are sensitive to the bias voltage and ambient temperature change. In our next IFM receiver design which is based on phase modulation, phase to intensity modulation conversion and DGD induced power fading is used. So the intensity modulator can be avoided. The proof-of-concept experiment shows a 1.7 GHz to 12.2 GHz measurement range with a reduced measurement error of ± 0.07 GHz.

The previous designs use ACF to perform frequency estimation; therefore, two PDs have to be used. To further reduce the number of PDs, an IFM receiver based on IIR filter has been proposed. This IIR filter has a transfer function similar to previous ACF so the measurement resolution can be maintained with the number of PDs reduced to one. The current measurement range is from 6.9088 to 6.9190 GHz which is limited by the long loop length. By using integrated solutions, wideband measurements can be obtained.

Chapter 7 CONCLUSION AND FUTURE WORK

7.1 Achievements and Conclusion

In this thesis, developments of microwave signal processing techniques for microwave photonic filters (MPFs) and microwave signal instantaneous frequency measurement (IFM) using photonic means have been presented. Our work belongs to the subject of microwave photonics which processes microwave signals conveyed by an optical carrier directly in the optical domain. So the modulation of microwave signals onto the optical carrier is an essential step. Because of the advantages of the LiNbO_3 modulators over those of direct modulation and other types of modulators, we mainly use these in our work. In the first part of the thesis, detailed modeling and calculation of the LiNbO_3 modulator-based analog optical fibre link has been given; this forms the basis of the rest of the work.

The second part of the thesis presents the development of novel incoherent MPFs. Two schemes for realizing the incoherent operation have been addressed. Under the first scheme, using orthogonal polarization, we have developed three high-birefringence (Hi-Bi) fibre Bragg grating (FBG)-based continuously tunable MPFs. The first one uses the differential group delay (DGD) of a Hi-Bi linearly chirped fibre Bragg grating (LCFBG) as the tuning mechanism. By applying mechanical stress to the grating, a continuous free spectral range (FSR) tuning of 1.11 GHz with 32 dB notch rejection has been achieved [79]. The second MPF consists of two arms using two polarization beam splitters (PBSs) with a Hi-Bi LCFBG in one of the arms. By varying the length of the arm with LCFBG, more than 5 GHz FSR tuning with 40 dB notch rejection has been achieved. The filter response is very stable due to the high

extinction ratio PBSs and polarization maintaining structure [82]. The third MPF aims to achieve high notch rejection. By using the chromatic dispersion (CD) of the nonlinearly chirped fibre Bragg grating (NLCFBG), more than 45 dB notch rejection has been achieved with 4.7 GHz FSR tuning [83]. In all cases, the measured filter response matches the theoretical calculations.

The basic scheme described above uses orthogonal polarizations so the number of taps is restricted to two. The second scheme uses multi-wavelength light source to make its coherence time shorter than the optical time delay difference. Therefore, the MPFs under this scheme are multi-tap finite impulse response (FIR) filters and offer flexibility in adjusting the shape of the filter response also rather than just the FSR. Based on the window method of the digital filter design, a theoretical model for the CD-based multi-tap MPFs has been setup. With the help of this model, a windowed Fabry-Pérot filter-based multi-wavelength tunable laser has been built which has more than 45 wavelengths and has a Blackman or Hamming power profile. By tuning the wavelength spacing of the laser and using phase modulation, a bandpass filter with 3 GHz of the passband centre frequency tuning and more than 25 dB out-of-band rejection ratio has been realized [95].

We have also investigated an infinite impulse response (IIR) filter design. A 3x3 collinear coupler has been used to construct recirculating delay line type MPF. Different filter responses have been obtained by changing the ratio between the two recirculating delay loop lengths. The measured results match well the theoretical calculations. When the ratio of the loop lengths is 1:2, more than 40 dB notch rejection has been achieved [102]. This filter shows potential for achieving high Q-factor bandpass filter with deep notches at stopbands.

The third part of the thesis presents the development of the frequency-to-power mapping-based IFM techniques keeping in view the improvement of measurement accuracy and the reduction of the system cost and complexity. Our work can be divided in three stages. In the first stage, a simple and cost effective IFM design based on a NLCFBG has been proposed. It only uses one laser source, one modulator, but the measurement range is from 8.22 GHz to 9.82 GHz and the low frequency band cannot be covered. In the second stage, we realize that the amplitude comparison function (ACF) formed by complementary transfer functions can cover low frequencies. We first use the power fading curve of the phase and intensity modulated signals to construct an ACF. This leads to a wide measurement range from DC to 13.8 GHz with a measurement error less than ± 0.3 GHz. To further reduce the number of modulators and measurement error, phase modulation-based IFM design has been proposed which uses phase to intensity modulation conversion and DGD induced power fading. In this way the intensity modulator of the previous stage has been avoided. The proof-of-concept experiment has shown reduced measurement error of ± 0.07 GHz. Because of the ACF, all the previous IFM techniques employ two PDs. In the third stage, we have designed an IIR filter which has a response similar to ACF, so that the number of PDs is reduced to one. Although the current measurement bandwidth, from 6.9088 to 6.9190 GHz, is limited by the long loop length, this can be extended by using integrated solutions [115].

7.2 Future Work

Though many MPF and IFM designs have been proposed and demonstrated in this thesis, there are still a number of ways in which the present work may be extended. In the following, some of these possibilities are mentioned.

In chapter 4, the passband centre frequency tuning and tap weight control are decided by the windowed Fabry-Pérot filter of multi-wavelength laser which is implemented by a programmable WaveShaper. Although the WaveShaper is very flexible, its overall bandwidth will limit the number of wavelengths or taps generated. The cost of the WaveShaper is also high. These drawbacks can be improved by replacing the WaveShaper with other tunable wavelength selection filters such as Sagnac loop comb filter [116], the filter based on high order polarization mode dispersion and polarizer [117], or multi-channel FBGs. The current FIR MPF has all positive taps. If negative taps can be introduced, the shape of the filter response can be controlled further; for instance, a square-shape response can be realized.

Currently, the 3x3 collinear coupler based RRDL filter in chapter 5 has only realized deep notches. It is possible to incorporate amplifiers in the two recirculating loops. High Q factors with deep notches are expected.

In chapter 6, we have demonstrated an IIR filter-based IFM design with only one laser source, one modulator and one photodetector. However, the measurement range is limited by the long loop length. It has been demonstrated that an SOA can be integrated with micro-ring resonators using a vertically stacked asymmetric twin waveguide structure which comprises InGaAsP, InP layers and InP substrate [118]. By using such a photonic integration technique, the loop length can be reduced greatly and it should be possible to increase the measurement range to tens of GHz.

Recently, slow light has attracted significant attention. Stimulated Brillouin scattering (SBS)-based slow light has the ability to tune the time delay of the optical signals continuously by adjusting the pump light power [119]. So, SBS is a promising technique to achieve tunability in MPFs.

Most recently, complex coefficient [120-125] MPFs have been proposed. With the help of negative and complex coefficients, filters can be synthesized with the most general case and have the advantage of filter center frequency tuning without the change of its response shape and FSR. So realizing multi-tap coefficient MPF with simple and cost effective structures is a promising research direction to work on.

APPENDIX A: CALCULATION OF THE DGD

INDUCED POWER FADING

The upper arm in Fig. 6.11 is same as the above figure.

The input microwave signal can be expressed as

$$V = V_0 \sin \omega_m t \quad (\text{A.1})$$

where V_0 is the amplitude of the RF signal.

Since after the half waveplate, two orthogonal optical components with equal power are generated along the x- and y- axis and expressed as

$$\begin{bmatrix} E_x \\ E_y \end{bmatrix} = \frac{\sqrt{2}}{2} E_0 \begin{bmatrix} \exp(j\omega_0 t) \\ \exp(j\omega_0 t) \end{bmatrix} \quad (\text{A.2})$$

A phase modulator can actually be considered as a variable birefringent component. Its birefringence changes with external voltage applied. If the principal axis of the phase modulator aligns with x-axis, its Jones matrix can be written as

$$J_{PM} = \begin{bmatrix} \exp(i\alpha \sin \omega_m t + j\phi_0) & 0 \\ 0 & 1 \end{bmatrix} \quad (\text{A.3})$$

where $\alpha = \pi \frac{V_0}{V_\pi}$ is the modulation index, V_π is the half-wave voltages of the phase modulator, ω_m is the angular frequency of RF signal, ϕ_0 is the initial phase difference between the two axes of the phase modulator which can be changed by adjusting the modulator bias.

So the output E-field from phase modulator is

$$\begin{aligned} \begin{bmatrix} E_x \\ E_y \end{bmatrix} &= \frac{\sqrt{2}}{2} E_0 \begin{bmatrix} \exp(i\alpha \sin \omega_m t + j\phi_0) & 0 \\ 0 & 1 \end{bmatrix} \begin{bmatrix} \exp(j\omega_0 t) \\ \exp(j\omega_0 t) \end{bmatrix} \\ &= \frac{\sqrt{2}}{2} E_0 \begin{bmatrix} \exp(j\omega_0 t + j\alpha \sin \omega_m t + j\phi_0) \\ \exp(j\omega_0 t) \end{bmatrix} \end{aligned} \quad (\text{A.4})$$

If the polarizer axis makes an angle θ with respect to the x-axis, the output e-field is

$$E = \frac{\sqrt{2}}{2} E_0 \begin{bmatrix} \cos \theta \exp(j\omega_0 t + j\alpha \sin \omega_m t + j\phi_0) \\ + \sin \theta \exp(j\omega_0 t) \end{bmatrix} \quad (\text{A.5})$$

When $\theta = \pi / 4$, Eq. (A.5) becomes

$$\begin{aligned} E &= \frac{1}{2} E_0 [\exp(j\omega_0 t + j\alpha \sin \omega_m t + j\phi_0) + \exp(j\omega_0 t)] \\ &= \frac{1}{2} E_0 \exp(j\omega_0 t) \exp\left(j \frac{\alpha \sin \omega_m t + \phi_0}{2}\right) \\ &\quad \times \left[\exp\left(j \frac{\alpha \sin \omega_m t + \phi_0}{2}\right) + \exp\left(-j \frac{\alpha \sin \omega_m t + \phi_0}{2}\right) \right] \\ &= E_0 \exp(j\omega_0 t) \exp\left[j \frac{\alpha}{2} \sin \omega_m t + j \frac{\phi_0}{2}\right] \cos\left[\frac{\alpha}{2} \sin \omega_m t + \frac{\phi_0}{2}\right] \\ &= E_0 \exp(j\omega_0 t) \exp(j\delta) \cos\left[\frac{\alpha}{2} \sin \omega_m t + \frac{\phi_0}{2}\right] \end{aligned} \quad (\text{A.6})$$

where $\delta = \pi \frac{V_0}{2V_\pi} \sin \omega_m t + \frac{\phi_0}{2}$

The corresponding optical power is

$$\begin{aligned}
 E &= E^* E \\
 &= E_0^2 \cos^2 \left(\frac{\alpha}{2} \sin \omega_m t + \frac{\phi_0}{2} \right) \\
 &= P_0 \{1 + \cos[\alpha \sin \omega_m t + \phi_0]\}
 \end{aligned} \tag{A.7}$$

If $\theta = -\pi/4$, Eq. (A.5) becomes

$$\begin{aligned}
 E &= \frac{1}{2} E_0 \left[\exp \left(j\omega_0 t + j\pi \frac{V_0}{V_\pi} \sin \omega_m t + j\phi_0 \right) - \exp(j\omega_0 t) \right] \\
 &= E_0 \exp(j\omega_0 t) \exp(j\theta) \sin \left[\frac{\alpha}{2} \sin \omega_m t + \frac{\phi_0}{2} \right]
 \end{aligned} \tag{A.8}$$

The corresponding optical power is

$$E = E^* E = P_0 \{1 - \cos[\alpha \sin \omega_m t + \phi_0]\} \tag{A.9}$$

Eq. (A.7) and Eq. (A.9) shows that after the polarizer, the phase modulated signal is converted to intensity modulated signal. By compare Eq. (A.7) and Eq. (A.9) we can also see that they are complemented. This can be used to generate negative coefficient in microwave photonic filters.

Let's just use the intensity modulation when $\theta = \pi/4$ for calculating the first order polarization mode dispersion induced power fading. When $\phi_0 = \frac{\pi}{2}$, Eq. (A.6) becomes

$$\begin{aligned}
 E &= E_0 \exp(j\omega_0 t + j\theta) \cos\left(\frac{\alpha}{2} \sin \omega_m t + \frac{\pi}{4}\right) \\
 &= E_0 \exp(j\omega_0 t + j\theta) \left\{ \cos\left[\frac{\alpha}{2} \cos(\omega_m t)\right] - \sin\left[\frac{\alpha}{2} \cos(\omega_m t)\right] \right\}
 \end{aligned} \tag{A.10}$$

By using the relations

$$\cos(z \cos \omega t) = J_0(z) + 2 \sum_{n=1}^{\infty} (-1)^n J_{2n}(z) \cos 2n\omega t$$

$$\sin(z \cos \omega t) = -2 \sum_{n=1}^{\infty} (-1)^n J_{2n-1}(z) \cos(2n-1)\omega t$$

$$\cos A \cos B = \frac{1}{2} [\cos(A-B) + \cos(A+B)]$$

and real electrical field, Eq. (A.10) becomes

$$\begin{aligned}
 E &= E_0 \cos(\omega_0 t + \theta) \left\{ \begin{aligned} &J_0\left(\frac{\alpha}{2}\right) + 2 \sum_{n=1}^{\infty} (-1)^n J_{2n}\left(\frac{\alpha}{2}\right) \cos(2n\omega_m t) \\ &- 2 \sum_{n=1}^{\infty} (-1)^n J_{2n-1}\left(\frac{\alpha}{2}\right) \cos[(2n-1)\omega_m t] \end{aligned} \right\} \\
 &= E_0 \left\{ \begin{aligned} &J_0\left(\frac{\alpha}{2}\right) \cos(\omega_0 t + \theta) \\ &- 2J_2\left(\frac{\alpha}{2}\right) \cos(\omega_0 t + \theta) \cos(2\omega_m t) + 2J_4\left(\frac{\alpha}{2}\right) \cos(\omega_0 t + \theta) \cos(4\omega_m t) + \dots \\ &+ 2J_1\left(\frac{\alpha}{2}\right) \cos(\omega_0 t + \theta) \cos(\omega_m t) - 2J_3\left(\frac{\alpha}{2}\right) \cos(\omega_0 t + \theta) \cos(3\omega_m t) + \dots \end{aligned} \right\} \\
 &= E_0 \left\{ \begin{aligned} &J_0\left(\frac{\alpha}{2}\right) \cos(\omega_0 t + \theta) \\ &J_1\left(\frac{\alpha}{2}\right) [\cos(\omega_0 t + \omega_m t + \theta) + \cos(\omega_0 t - \omega_m t + \theta)] \\ &- J_2\left(\frac{\alpha}{2}\right) [\cos(\omega_0 t + 2\omega_m t + \theta) + \cos(\omega_0 t - 2\omega_m t + \theta)] + \dots \end{aligned} \right\}
 \end{aligned} \tag{A.11}$$

Under small signal modulation condition, Eq. (A.11) can be simplified as

$$E = E_0 \begin{bmatrix} J_0\left(\frac{\alpha}{2}\right) \cos(\omega_0 t + \theta) \\ + J_1\left(\frac{\alpha}{2}\right) \cos(\omega_0 t - \omega_m t + \theta) + J_1\left(\frac{\alpha}{2}\right) \cos(\omega_0 t + \omega_m t + \theta) \end{bmatrix} \quad (\text{A.12})$$

The PMF has only first order DGD. After two signals propagating through the PMF along the two axes, the phase difference induced by DGD is $\phi_{DGD} = \Delta\tau\Delta\omega$ between the two principle axes [22, 126-128].

Because the fast and slow axes of the PMF are the principle axes, and they are also aligned with x- and y- axes. After propagating through the PMF, the E-field can be expressed using complex forms as

$$E_x = \frac{\sqrt{2}}{2} E_0 \begin{bmatrix} J_0\left(\frac{\alpha}{2}\right) \exp(j\omega_0 t + j\theta - j\beta_0 L) \\ + J_1\left(\frac{\alpha}{2}\right) \exp\left(j\omega_0 t - j\omega_m t + j\theta - j\frac{\Delta\tau\omega_m}{2} - j\beta_0 L + j\beta_1\omega_m L - j\frac{1}{2}\beta_2 L\omega_m^2\right) \\ + J_1\left(\frac{\alpha}{2}\right) \exp\left(j\omega_0 t + j\omega_m t + j\theta + j\frac{\Delta\tau\omega_m}{2} - j\beta_0 L - j\beta_1\omega_m L - j\frac{1}{2}\beta_2 L\omega_m^2\right) \end{bmatrix} \quad (\text{A.13})$$

$$E_y = \frac{\sqrt{2}}{2} E_0 \left[\begin{aligned} & J_0\left(\frac{\alpha}{2}\right) \exp(j\omega_0 t + j\theta - j\beta_0 L) \\ & + J_1\left(\frac{\alpha}{2}\right) \exp\left(j\omega_0 t - j\omega_m t + j\theta + j\frac{\Delta\tau\omega_m}{2} - j\beta_0 L + j\beta_1\omega_m L - j\frac{1}{2}\beta_2 L\omega_m^2\right) \\ & + J_1\left(\frac{\alpha}{2}\right) \exp\left(j\omega_0 t + j\omega_m t + j\theta - j\frac{\Delta\tau\omega_m}{2} - j\beta_0 L - j\beta_1\omega_m L - j\frac{1}{2}\beta_2 L\omega_m^2\right) \end{aligned} \right] \quad (\text{A.14})$$

When deriving Eq. (A.13) and (A.14), $\Delta\omega = (\omega_0 + \omega_m) - \omega_0 = \omega_m$ for upper sideband and $\Delta\omega = (\omega_0 - \omega_m) - \omega_0 = -\omega_m$ for lower sideband. The phase delay caused by CD which is expressed in Eq. (2.12) is also considered.

Because E_x and E_y are orthogonal, the detected optical power can be expressed as

$$I = \bar{E} \cdot \bar{E}^* = \bar{E}_x \cdot \bar{E}_x^* + \bar{E}_y \cdot \bar{E}_y^* \quad (\text{A.15})$$

For the optical signal along x-axis we have

$$\bar{E}_x \cdot \bar{E}_x^* = \frac{1}{2} E_0^2 \left[\begin{aligned} & J_0\left(\frac{\alpha}{2}\right) \exp(j\omega_0 t + j\theta - j\beta_0 L) \\ & + J_1\left(\frac{\alpha}{2}\right) \exp\left(j\omega_0 t - j\omega_m t + j\theta - j\frac{\Delta\tau\omega_m}{2} - j\beta_0 L + j\beta_1\omega_m L - j\frac{1}{2}\beta_2 L\omega_m^2\right) \\ & + J_1\left(\frac{\alpha}{2}\right) \exp\left(j\omega_0 t + j\omega_m t + j\theta + j\frac{\Delta\tau\omega_m}{2} - j\beta_0 L - j\beta_1\omega_m L - j\frac{1}{2}\beta_2 L\omega_m^2\right) \end{aligned} \right]$$

$$\begin{aligned}
 & \times \left[\begin{aligned} & J_0\left(\frac{\alpha}{2}\right) \exp(-j\omega_0 t - j\theta + j\beta_0 L) \\ & + J_1\left(\frac{\alpha}{2}\right) \exp\left(-j\omega_0 t + j\omega_m t - j\theta + j\frac{\Delta\tau\omega_m}{2} + j\beta_0 L - j\beta_1\omega_m L + j\frac{1}{2}\beta_2 L\omega_m^2\right) \\ & + J_1\left(\frac{\alpha}{2}\right) \exp\left(-j\omega_0 t - j\omega_m t - j\theta - j\frac{\Delta\tau\omega_m}{2} + j\beta_0 L + j\beta_1\omega_m L + j\frac{1}{2}\beta_2 L\omega_m^2\right) \end{aligned} \right] \\
 & \approx \frac{1}{2} E_0^2 \left[\begin{aligned} & J_0^2\left(\frac{\alpha}{2}\right) + 2J_0\left(\frac{\alpha}{2}\right) J_1\left(\frac{\alpha}{2}\right) \exp\left(j\omega_m t + j\frac{\Delta\tau\omega_m}{2} - j\beta_1\omega_m L + j\frac{1}{2}\beta_2 L\omega_m^2\right) \\ & + 2J_0\left(\frac{\alpha}{2}\right) J_1\left(\frac{\alpha}{2}\right) \exp\left(-j\omega_m t - j\frac{\Delta\tau\omega_m}{2} + j\beta_1\omega_m L + j\frac{1}{2}\beta_2 L\omega_m^2\right) \\ & + J_0^2\left(\frac{\alpha}{2}\right) + 2J_0\left(\frac{\alpha}{2}\right) J_1\left(\frac{\alpha}{2}\right) \exp\left(-j\omega_m t - j\frac{\Delta\tau\omega_m}{2} + j\beta_1\omega_m L - j\frac{1}{2}\beta_2 L\omega_m^2\right) \\ & + 2J_0\left(\frac{\alpha}{2}\right) J_1\left(\frac{\alpha}{2}\right) \exp\left(j\omega_m t + j\frac{\Delta\tau\omega_m}{2} - j\beta_1\omega_m L - j\frac{1}{2}\beta_2 L\omega_m^2\right) \end{aligned} \right] \\
 & = E_0^2 \left[\begin{aligned} & J_0^2\left(\frac{\alpha}{2}\right) + 2J_0\left(\frac{\alpha}{2}\right) J_1\left(\frac{\alpha}{2}\right) \cos\left(\omega_m t + \frac{\Delta\tau\omega_m}{2} - \beta_1\omega_m L + \frac{1}{2}\beta_2 L\omega_m^2\right) \\ & + 2J_0\left(\frac{\alpha}{2}\right) J_1\left(\frac{\alpha}{2}\right) \cos\left(\omega_m t + \frac{\Delta\tau\omega_m}{2} - \beta_1\omega_m L - \frac{1}{2}\beta_2 L\omega_m^2\right) \end{aligned} \right] \\
 & = E_0^2 \left[J_0^2\left(\frac{\alpha}{2}\right) + 4J_0\left(\frac{\alpha}{2}\right) J_1\left(\frac{\alpha}{2}\right) \cos\left(\frac{1}{2}\beta_2 L\omega_m^2\right) \cos\left(\omega_m t + \frac{\Delta\tau\omega_m}{2} - \beta_1\omega_m L\right) \right] \tag{A.16}
 \end{aligned}$$

Similarly, we have

$$\bar{E}_y \cdot \bar{E}_y^* \approx E_0^2 \left[J_0^2\left(\frac{\alpha}{2}\right) + 4J_0\left(\frac{\alpha}{2}\right) J_1\left(\frac{\alpha}{2}\right) \cos\left(\frac{1}{2}\beta_2 L\omega_m^2\right) \cos\left(\omega_m t - \frac{\Delta\tau\omega_m}{2} - \beta_1\omega_m L\right) \right] \tag{A.17}$$

So the total optical power is

$$\begin{aligned}
 I &\approx E_0^2 \left[J_0^2\left(\frac{\alpha}{2}\right) + 4J_0\left(\frac{\alpha}{2}\right)J_1\left(\frac{\alpha}{2}\right)\cos\left(\frac{1}{2}\beta_2 L\omega_m^2\right)\cos\left(\omega_m t + \frac{\Delta\tau\omega_m}{2} - \beta_1\omega_m L\right) \right] \\
 &\quad + E_0^2 \left[J_0^2\left(\frac{\alpha}{2}\right) + 4J_0\left(\frac{\alpha}{2}\right)J_1\left(\frac{\alpha}{2}\right)\cos\left(\frac{1}{2}\beta_2 L\omega_m^2\right)\cos\left(\omega_m t - \frac{\Delta\tau\omega_m}{2} - \beta_1\omega_m L\right) \right] \\
 &\approx E_0^2 \left[2J_0^2\left(\frac{\alpha}{2}\right) + 8J_0\left(\frac{\alpha}{2}\right)J_1\left(\frac{\alpha}{2}\right)\cos\left(\frac{1}{2}\beta_2 L\omega_m^2\right)\cos\left(\frac{\Delta\tau\omega_m}{2}\right)\cos(\omega_m t - \beta_1\omega_m L) \right]
 \end{aligned} \tag{A.18}$$

Ignore the DC and constants, we have

$$I \approx \cos\left(\frac{1}{2}\beta_2 L\omega_m^2\right)\cos\left(\frac{\Delta\tau\omega_m}{2}\right)\cos(\omega_m t - \tau_0) \tag{A.19}$$

where $\tau_0 = \beta_1\omega_m L$ is the optical signal group delay.

By using the relations $\omega_m = 2\pi f$ and $D = -2\pi c\beta_2/\lambda^2$, and compare the detected signal with modulating microwave signal, the amplitude transfer function can be derived

$$H(f) \approx \cos\left(\frac{\pi D L \lambda_c^2 f^2}{c}\right)\cos(\pi f \Delta\tau) \tag{A.20}$$

From Eq. (A.20) we can see that the second cosine term is the DGD induced power fading. If the length of the PMF is not long, the CD induced power fading can be ignored and the microwave power transfer function is

$$|H(f)|^2 \approx \cos^2(\pi f \Delta\tau) \tag{A.21}$$

PUBLICATIONS

Journal papers:

- [1] **J. Q. Zhou**, S. Fu, F. Luan, S. Aditya, P. P. Shum, and K. E. K. Lee, "Tunable multi-tap bandpass microwave photonic filter using a windowed Fabry-Pérot filter based multi-wavelength tunable laser," *J. Lightwave Technol.*, accepted, 2011.
- [2] **J. Q. Zhou**, S. Aditya, P. Shum, and J. Yao, "Instantaneous Microwave Frequency Measurement Using a Photonic Microwave Filter with an Infinite Impulse Response," *IEEE Photon. Technol. Lett.*, vol. 22, pp. 682-684, May 2010.
- [3] **J. Q. Zhou**, S. Fu, S. Aditya, P. P. Shum, and C. Lin, "Instantaneous Microwave Frequency Measurement Using Photonic Technique," *IEEE Photon. Technol. Lett.*, vol. 21, pp. 1069-1071, Aug. 2009
- [4] **J. Q. Zhou**, S. Aditya, P. Shum, L. H. Cheng and B. P. Parhusip , "Microwave Photonic Bandpass Filter Using a Multi-wavelength Laser with a Bell-Shaped Power Profile", *Microwave Opt. Tech. Lett.*, vol. 51, pp. 1329-1332, May 2009.
- [5] **J. Q. Zhou**, S. Fu, P. P. Shum, S. Aditya, L. Xia, J. Li, X. Sun, and K. Xu, "Photonic measurement of microwave frequency based on phase modulation," *Opt. Express*, vol. 17, pp. 7217-7221, Apr. 2009.
- [6] **J. Q. Zhou**, S. Aditya, P. Shum, L. Xia, and B.P. Parhusip, "Wide Range Continuously Tunable Microwave Photonic Filter Using High-Birefringence Linearly Chirped Fiber Bragg Grating and Polarization Beam Splitters", *Opt. Eng.*, vol. 48, p. 010502, Jan 2009.
- [7] G. Ning, **J. Q. Zhou**, S. Aditya, P. Shum, Vincent Wong, and Desmond Lim, "Multiple Dual-Wavelengths Erbium-Doped Fiber Ring Laser Using a Polarization-Maintaining Fabry-Pérot Filter", *IEEE Photon. Technol. Lett.*, vol.20, pp. 1606-1608, Oct. 2008.
- [8] G. Ning, **J. Q. Zhou**, L. Cheng, S. Aditya, and P. Shum "Generation of different modulation formats using Sagnac fiber loop with One Electroabsorption Modulator", *IEEE Photon. Technol. Lett.*, vol. 20, pp. 297-299, Feb. 2008.

Conference papers:

- [1] **J. Q. Zhou**, S. Aditya, P. P. Shum, K. E. K. Lee, and V. Wong, "Continuously Tunable Microwave Photonic Filter Based on High-Birefringence Linearly Chirped Fiber Bragg Gratings," in *Asia-Pacific Microwave Photonics conference 2010 (APMP2010)* Hong Kong, China, 2010.
- [2] **J. Q. Zhou**, S. Fu, S. Aditya, P. P. Shum, C. Lin, V. Wong, and D. Lim, "Photonic Temporal Differentiator based on Polarization Modulation in a LiNbO₃ Phase Modulator," in *2009 International Topical Meeting on Microwave Photonics*, Valencia, Spain, Oct. 2009.
- [3] **J. Q. Zhou**, S. Fu, L. Xia, S. Aditya, P. P. Shum, C. Lin, V. Wong, and D. Lim, "Instantaneous microwave frequency measurement using an asymmetric non-linear group delay profile," in *2009 International Topical Meeting on Microwave Photonics*, Valencia, Spain, Oct. 2009.
- [4] **J. Q. Zhou**, L. Xia, S. Aditya, P. P. Shum, and B. P. Parhusip, "Nonlinearly Chirped Grating Based Continuously Tunable High Notch Rejection Microwave Photonic Filter," in *2009 Asia-Pacific Microwave Photonics Conference (APMP2009)* Beijing, China, Apr. 2009.
- [5] **J. Q. Zhou**, H. Dong, S. Aditya, P. Shum, L. Xia, B. P. Parhusip and X. L. Tian, "Reflective 3x3 coupler-based double recirculating delay line microwave photonic filter", *Photonics 2008* New Delhi, India, Dec. 2008.
- [6] **J. Q. Zhou**, H. Dong, S. Aditya, P. Shum, L. Xia, Songnian Fu, M. Tang, "Continuously Tunable Microwave Photonic Filter Based on High-Birefringence Linearly Chirped Grating", in *SPIE Asia-Pacific Optical Communications* Hangzhou, China, Oct. 2008.

REFERENCES

- [1] J. Capmany, B. Ortega, and D. Pastor, "A tutorial on microwave photonic filters," *J. Lightwave Technol.*, vol. 24, pp. 201-229, 2006.
- [2] J. Capmany and D. Novak, "Microwave photonics combines two worlds," *Nature Photon.*, vol. 1, pp. 319-330, Jun. 2007.
- [3] J. P. Yao, "Microwave Photonics," *J. Lightwave Technol.*, vol. 27, pp. 314-335, Jan-Feb 2009.
- [4] J. Capmany, B. Ortega, D. Pastor, and S. Sales, "Discrete-time optical Processing of microwave signals," *J. Lightwave Technol.*, vol. 23, pp. 702-723, Feb 2005.
- [5] K. Wilner and A. P. van den Heuvel, "Fiber-optic delay lines for microwave signal processing," *Proc. IEEE*, vol. 64, pp. 805-807, 1976.
- [6] C. T. Chang, J. A. Cassaboom, and H. F. Taylor, "Fibre-optic delay-line devices for r.f. signal processing," *Electron. Lett.*, vol. 13, pp. 678-680, 1977.
- [7] A. J. Seeds, "Microwave photonics," *IEEE Trans. Microw. Theory Tech.*, vol. 50, pp. 877-887, 2002.
- [8] R. A. Minasian, "Photonic signal processing of microwave signals," *IEEE Trans. Microw. Theory Tech.*, vol. 54, pp. 832-846, Feb 2006.
- [9] D. C. Schleher, *Introduction to electronic warfare*: Artech House, 1988.
- [10] A. E. Spezio, "Electronic warfare systems," *IEEE Trans. Microw. Theory Tech.*, vol. 50, pp. 633-644, 2002.
- [11] D. L. Adamy, *Introduction to electronic warfare modeling and simulation*: Artech House, Inc., 2003.
- [12] D. C. Schleher, *Electronic warfare in the information age*: Artech House, Inc., 1999.
- [13] W. G. James, "Instantaneous frequency measurement system," 3,939,411, 1976.
- [14] J. B. Y. Tsui, "Instantaneous frequency measurement receiver with digital processing," 4,633,516, 1986.
- [15] D. Lynch, *Introduction to RF Stealth*: SciTech Publishing Inc., 2004.
- [16] L. V. T. Nguyen and D. B. Hunter, "A photonic technique for microwave frequency measurement," *IEEE Photon. Technol. Lett.*, vol. 18, pp. 1188-1190, May 2006.
- [17] D. Norton, S. Johns, C. Keefer, and R. Soref, "Tunable microwave filtering using high dispersion fiber time delays," *IEEE Photon. Technol. Lett.*, vol. 6, pp. 831-832, 1994.
- [18] J. E. Bowers, S. A. Newton, and H. J. Shaw, "Fibre-optic variable delay lines," *Electron. Lett.*, vol. 18, p. 999, 1982.
- [19] W. Zhang, J. A. R. Williams, and I. Bennion, "Optical fibre delay line filter free of limitation imposed by optical coherence," *Electron. Lett.*, vol. 35, pp. 2133-2134, 1999.
- [20] W. Zhang, A. R. Williams, L. A. Everall, and I. A.-B. Bennion, I., "Fibre optic radio frequency notch filter with linear and continuous tuning by using a chirped fibre grating," *Electron. Lett.*, vol. 34, pp. 1770-1772, Sep. 1998.

- [21] G. X. Ning, P. Shum, S. Aditya, Y. D. Gong, and C. Lu, "Novel tunable microwave photonic notch filter with a variable polarization beamsplitter and a Hi-Bi coupler," *Opt. Eng.*, vol. 44, Oct 2005.
- [22] G. X. Ning, S. Aditya, P. Shum, and N. Liu, "Tunable microwave filter that uses a high-birefringent fiber and a differential-group-delay element," *J. Opt. soc. Am. A-Optics Image Science and Vision*, vol. 22, pp. 913-916, May 2005.
- [23] W. Zhang, J. A. R. Williams, and I. Bennion, "Polarization synthesized optical transversal filter employing high birefringence fiber gratings," *IEEE Photon. Technol. Lett.*, vol. 13, pp. 523-525, 2001.
- [24] X. K. Yi, C. Lu, X. F. Yang, W. D. Zhong, F. Wei, *et al.*, "Continuously tunable microwave-photonic filter design using high-birefringence linear chirped grating," *IEEE Photon. Technol. Lett.*, vol. 15, pp. 754-756, May 2003.
- [25] J. Mora, S. Sales, M. D. Manzanedo, R. Garcia-Olcina, J. Capmany, *et al.*, "Continuous tuning of photonic transversal filter based on the modification of tapped weights," *IEEE Photon. Technol. Lett.*, vol. 18, pp. 1594-1596, 2006.
- [26] V. Polo, B. Vidal, J. L. Corral, and J. Marti, "Novel tunable photonic microwave filter based on laser arrays and N x N AWG-based delay lines," *IEEE Photon. Technol. Lett.*, vol. 15, pp. 584-586, 2003.
- [27] J. Mora, B. Ortega, A. Diez, J. L. Cruz, M. V. Andres, *et al.*, "Photonic microwave tunable single-bandpass filter based on a Mach-Zehnder interferometer," *J. Lightwave Technol.*, vol. 24, pp. 2500-2509, 2006.
- [28] X. K. Yi and R. A. Minasian, "Dispersion induced RF distortion of spectrum-sliced microwave-photonic filters," *IEEE Trans. Microw. Theory Tech.*, vol. 54, pp. 880-886, Feb. 2006.
- [29] J. Mora, M. V. Andres, J. L. Cruz, B. Ortega, J. Capmany, *et al.*, "Tunable all-optical negative multitap microwave filters based on uniform fiber Bragg gratings," *Opt. Lett.*, vol. 28, pp. 1308-1310, Aug. 2003.
- [30] X. Yi and R. A. Minasian, "New Spectrum-Sliced Microwave Photonic Filter for High-Frequency Signal Processing," *IEEE Photon. Technol. Lett.*, vol. 21, pp. 230-232, 2009.
- [31] L. R. Chen and V. Page, "Tunable photonic microwave filter using semiconductor fibre laser," *Electron. Lett.*, vol. 41, pp. 1183-1184, 2005.
- [32] D. Liu, N. Q. Ngo, and S. C. Tjin, "Microwave photonic bandpass filter using a multiwavelength semiconductor-optical-amplifier ring laser," *Opt. Eng.*, vol. 46, 2007.
- [33] X. H. Feng, C. Lu, H. Y. Tam, and P. K. A. Wai, "Reconfigurable microwave photonic filter using multiwavelength erbium-doped fiber laser," *IEEE Photon. Technol. Lett.*, vol. 19, pp. 1334-1336, Sep-Oct 2007.
- [34] M. Tur, J. W. Goodman, B. Moslehi, J. E. Bowers, and H. J. Shaw, "Fiber-optic signal processor with applications to matrix-vector multiplication and lattice filtering," *Opt. Lett.*, vol. 7, pp. 463-465, 1982.
- [35] S. A. Newton, R. S. Howland, K. P. Jackson, and H. J. Shaw, "High-speed pulse-train generation using single-mode-fibre recirculating delay lines," *Electron. Lett.*, vol. 19, pp. 756-758, 1983.
- [36] R. L. Ohlhaber and K. Wilner, "Fiber optic delay-lines for pulse coding," *Electro-Optical Systems Design*, vol. 9, pp. 33-35, 1977.
- [37] K. P. Jackson, S. A. Newton, and H. J. Shaw, "1-GBit/s code generator and matched-filter using an optical fiber tapped delay-line," *Appl. Phys. Lett.*, vol. 42, pp. 556-558, 1983.

- [38] B. Moslehi, J. W. Goodman, M. Tur, and H. J. Shaw, "Fiber-optic lattice signal processing," *Proc. IEEE*, vol. 72, pp. 909-930, 1984.
- [39] S. T. Winnall and A. C. Lindsay, "A Fabry-Perot scanning receiver for microwave signal processing," *IEEE Trans. Microw. Theory Tech.*, vol. 47, pp. 1385-1390, 1999.
- [40] H. L. Guo, G. Z. Xiao, N. Mrad, and J. P. Yao, "Measurement of Microwave Frequency Using a Monolithically Integrated Scannable Echelle Diffractive Grating," *IEEE Photon. Technol. Lett.*, vol. 21, pp. 45-47, Jan. 2009.
- [41] W. S. Wang, R. L. Davis, T. J. Jung, R. Lodenkamper, L. J. Lembo, *et al.*, "Characterization of a coherent optical RF channelizer based on a diffraction grating," *IEEE Trans. Microw. Theory Tech.*, vol. 49, pp. 1996-2001, Oct. 2001.
- [42] S. T. Winnall, A. C. Lindsay, M. W. Austin, J. Canning, and A. Mitchell, "A microwave channelizer and spectroscope based on an integrated optical Bragg-grating Fabry-Perot and integrated hybrid fresnel lens system," *IEEE Trans. Microw. Theory Tech.*, vol. 54, pp. 868-872, Feb. 2006.
- [43] F. A. Volkening, "Photonic channelized RF receiver employing dense wavelength division multiplexing," US 7,245,833 B1, Jul. 17, 2007.
- [44] L. V. T. Nguyen, "Microwave frequency measurement utilising frequency to time mapping," in *International Topics Meeting on Microwave Photonics, 2008. Jointly held with the 2008 Asia-Pacific Microwave Photonics Conference (MWP/APMP 2008)*, Beijing, China, 2008, pp. 330-332.
- [45] L. V. T. Nguyen, "Microwave Photonic Technique for Frequency Measurement of Simultaneous Signals," *IEEE Photon. Technol. Lett.*, vol. 21, pp. 642-644, May. 2009.
- [46] X. Zou and J. Yao, "An Optical Approach to Microwave Frequency Measurement With Adjustable Measurement Range and Resolution," *IEEE Photon. Technol. Lett.*, vol. 20, pp. 1989-1991, Dec. 2008.
- [47] X. M. Zhang, H. Chi, S. L. Zheng, X. F. Jin, and J. P. Yao, "Instantaneous Microwave Frequency Measurement Using an Optical Phase Modulator," *IEEE Microw. Wireless Compon. Lett.*, vol. 19, pp. 422-424, Jun. 2009.
- [48] J. Q. Li, S. N. Fu, K. Xu, J. Q. Zhou, P. Shum, *et al.*, "Photonic-assisted microwave frequency measurement with higher resolution and tunable range," *Opt. Lett.*, vol. 34, pp. 743-745, Mar. 2009.
- [49] X. Zou and J. Yao, "Microwave frequency measurement with improved measurement range and resolution," *Electron. Lett.*, vol. 45, pp. 497-498, 2009.
- [50] H. Emami, N. Sarkhosh, L. A. Bui, and A. Mitchell, "Amplitude independent RF instantaneous frequency measurement system using photonic Hilbert transform," *Opt. Express*, vol. 16, pp. 13707-13712, Sep. 2008.
- [51] M. Attygalle and D. B. Hunter, "Improved Photonic Technique for Broadband Radio-Frequency Measurement," *IEEE Photon. Technol. Lett.*, vol. 21, pp. 206-208, Feb. 2009.
- [52] T. Ikegami and Y. Suematsu, "Resonance-like characteristics of the direct modulation of a junction laser," *Proc. IEEE*, vol. 55, pp. 122-123, 1967.
- [53] Y. Matsui, H. Murai, S. Arahira, S. Kutsuzawa, and Y. Ogawa, "30-GHz bandwidth 1.55- μ m strain-compensated InGaAlAs-InGaAsP MQW laser," *IEEE Photon. Technol. Lett.*, vol. 9, pp. 25-27, Jan 1997.
- [54] M. C. Wu, C. Chang-Hasnain, E. K. Lau, X. X. Zhao, and Osa, "High-speed modulation of optical injection-locked semiconductor lasers," in *2008 Conference on Optical Fiber Communication/National Fiber Optic Engineers*

- Conference, Vols 1-8*, ed Washington: Optical Soc America, 2008, pp. 1880-1882.
- [55] C. Henry, "Theory of the linewidth of semiconductor lasers," *IEEE J. Quantum Electron.*, vol. 18, pp. 259-264, 1982.
 - [56] T. L. Koch and J. E. Bowers, "Nature of wavelength chirping in directly modulated semiconductor lasers," *Electron. Lett.*, vol. 20, pp. 1038-1040, 1984.
 - [57] M. Osinski and J. Buus, "Linewidth broadening factor in semiconductor lasers--An overview," *IEEE J. Quantum Electron.*, vol. 23, pp. 9-29, 1987.
 - [58] L. A. Coldren and S. W. Corzine, *Diode Lasers and Photonic Integrated Circuits*. New York: John Wiley & Sons, Inc, 1995.
 - [59] G. L. Li and P. K. L. Yu, "Optical intensity modulators for digital and analog applications," *J. Lightwave Technol.*, vol. 21, pp. 2010-2030, 2003.
 - [60] I. Kaminow and T. Li, *Optical Fiber Telecommunications IV A Components*: Academic Press An Elsevier science Imprint, 2002.
 - [61] L. V. Keldysh, "The effect of a strong electric field on the optical properties of insulating crystals," *Sov. Phys.*, vol. 7, pp. 788-790, 1958.
 - [62] W. Franz, "Einfluss eines elektrischen feldes auf eine optische absorptionskante," *Zeitschrift Fur Naturforschung Part a-Astrophysik Physik Und Physikalische Chemie*, vol. 13, pp. 484-489, 1958.
 - [63] J. D. Dow and D. Redfield, "Electroabsorption in semiconductors: the excitonic absorption edge," *Phys. Rev. B*, vol. 1, pp. 3358-&, 1970.
 - [64] D. A. B. Miller, D. S. Chemla, T. C. Damen, A. C. Gossard, W. Wiegmann, *et al.*, "Electric field dependence of optical absorption near the band gap of quantum-well structures," *Phys. Rev. B*, vol. 32, p. 1043, 1985.
 - [65] G. P. Agrawal, *Fiber-optic Communications Systems, Third Edition*: John Wiley & Sons, Inc., 2002.
 - [66] (2008). *OM5653C-30B 40 Gb/s EA Modulator with Polarization Maintaining Fiber, Built-in Cooler*. Available: http://www.okisemiopto.com/pdf/fids_om5653c_30b.pdf
 - [67] R. G. Walker, "High-speed III-V semiconductor intensity modulators," *IEEE J. Quantum Electron.*, vol. 27, pp. 654-667, 1991.
 - [68] R. Spickermann, S. R. Sakamoto, M. G. Peters, and N. Dagli, "GaAs/AlGaAs travelling wave electro-optic modulator with an electrical bandwidth >40 GHz," *Electron. Lett.*, vol. 32, pp. 1095-1096, 1996.
 - [69] C. C. Teng, "Travelling-wave polymeric optical-intensity modulator with more than 40 GHz of 3-dB electrical bandwidth," *Appl. Phys. Lett.*, vol. 60, pp. 1538-1540, Mar 1992.
 - [70] C. Datong, D. Bhattacharya, A. Udupa, B. Tsap, H. R. Fetterman, *et al.*, "High-frequency polymer modulators with integrated finline transitions and low V_{π} ," *IEEE Photon. Technol. Lett.*, vol. 11, pp. 54-56, 1999.
 - [71] K. Noguchi, H. Miyazawa, and O. Mitomi, "75 GHz broadband Ti:LiNbO₃ optical modulator with ridge structure," *Electron. Lett.*, vol. 30, pp. 949-951, 1994.
 - [72] K. Noguchi, O. Mitomi, and H. Miyazawa, "Millimeter-wave Ti:LiNbO₃ optical modulators," *J. Lightwave Technol.*, vol. 16, pp. 615-619, 1998.
 - [73] H.-W. Chen, Y.-h. Kuo, and J. E. Bowers, "High speed silicon modulators," in *OptoElectronics and Communications Conference, 2009. OECC 2009. 14th*, 2009, pp. 1-2.

- [74] G. Keiser, *Optical fiber communications third edition*: The McGraw-Hill Companies, Inc., 2000.
- [75] F. Zeng and J. P. Yao, "All-optical bandpass microwave filter based on an electro-optic phase modulator," *Opt. Express*, vol. 12, pp. 3814-3819, Aug. 2004.
- [76] J. Yao, F. Zeng, and Q. Wang, "Photonic Generation of Ultrawideband Signals," *J. Lightwave Technol.*, vol. 25, pp. 3219-3235, 2007.
- [77] G. H. Smith, D. Novak, and Z. Ahmed, "Overcoming chromatic-dispersion effects in fiber-wireless systems incorporating external modulators," *IEEE Trans. Microw. Theory Tech.*, vol. 45, pp. 1410-1415, 1997.
- [78] J. Wang and J. Yao, "A tunable photonic microwave notch filter based on all-optical mixing," *IEEE Photon. Technol. Lett.*, vol. 18, pp. 382-384, 2006.
- [79] J. Q. Zhou, H. Dong, S. Aditya, P. Shum, L. Xia, *et al.*, "Continuously Tunable Microwave Photonic Filter Based on High-Birefringence Linearly Chirped Grating," in *SPIE Asia-Pacific Optical Communications (APOC2008)*, Hangzhou, China, 2008.
- [80] T. Erdogan, "Fiber grating spectra," *J. Lightwave Technol.*, vol. 15, pp. 1277-1294, 1997.
- [81] X. Y. Dong, P. Shum, N. Q. Ngo, C. C. Chan, J. H. Ng, *et al.*, "Largely tunable CFBG-based dispersion compensator with fixed center wavelength," *Opt. Express*, vol. 11, pp. 2970-2974, Nov. 2003.
- [82] J. Q. Zhou, S. Aditya, P. Shum, L. Xia, and B. P. Parhusip, "Wide-range continuously tunable microwave photonic filter using high-birefringence linearly chirped fiber Bragg grating and polarization beamsplitters," *Opt. Eng.*, vol. 48, p. 010502, Jan. 2009.
- [83] J. Q. Zhou, L. Xia, S. Aditya, P. S. Ping, and B. P. Parhusip, "Nonlinearly chirped grating based continuously tunable high notch rejection microwave photonic filter," in *Proceedings of 2009 2nd IEEE International Conference on Broadband Network and Multimedia Technology (IEEE IC-BNMT2009)*, Beijing, China, 2009, pp. 913-915.
- [84] H. Dong, J. Q. Zhou, S. Aditya, Y. D. Gong, and P. Shum, "Stable operation of microwave photonic filters constructed with two cascaded Hi-Bi fibers," *Microwave Opt. Tech. Lett.*, vol. 52, pp. 207-212, Jan 2010.
- [85] J. Chow, G. Town, B. Eggleton, M. Ibsen, K. Sugden, *et al.*, "Multiwavelength generation in an erbium-doped fiber laser using in-fiber comb filters," *Photonics Technology Letters, IEEE*, vol. 8, pp. 60-62, 1996.
- [86] Y. Zhao, C. Shu, S. P. Li, H. Ding, and K. T. Chan, "Multiple wavelength operation of a unidirectional Er-doped fiber ring laser with optical feedback," in *Lasers and Electro-Optics, 1997. CLEO '97., Summaries of Papers Presented at the Conference on VO - 11, 1997*, p. 396.
- [87] G. J. Cowle and D. Y. Stepanov, "Multiple wavelength generation with Brillouin/erbium fiber lasers," *IEEE Photon. Technol. Lett.*, vol. 8, pp. 1465-1467, 1996.
- [88] X. Liu and C. Lu, "Self-stabilizing effect of four-wave mixing and its applications on multiwavelength erbium-doped fiber lasers," *IEEE Photon. Technol. Lett.*, vol. 17, pp. 2541-2543, 2005.
- [89] X. H. Feng, H. Y. Tam, and P. K. A. Wai, "Stable and uniform multiwavelength erbium-doped fiber laser using nonlinear polarization rotation," *Opt. Express*, vol. 14, pp. 8205-8210, Sep. 2006.
- [90] G. Ning, J. Q. Zhou, S. Aditya, P. Shum, V. Wong, *et al.*, "Multiple Dual-Wavelengths Erbium-Doped Fiber Ring Laser Using a Polarization-

- Maintaining Fabry-Perot Filter," *IEEE Photon. Technol. Lett.*, vol. 20, pp. 1606-1608, Oct. 2008.
- [91] J. Q. Zhou, S. Aditya, P. Shum, L. H. Cheng, B. P. Parhusip, *et al.*, "Microwave photonic bandpass filter using a multi-wavelength laser with a bell-shaped power profile," *Microwave Opt. Tech. Lett.*, vol. 51, pp. 1329-1332, May 2009.
- [92] A. V. Oppenheim and R. W. Schaffer, *Discrete-time signal processing*: Prentice-Hall International, Inc., 1989.
- [93] Q. Wang and J. Yao, "Multitap Photonic Microwave Filters With Arbitrary Positive and Negative Coefficients Using a Polarization Modulator and an Optical Polarizer," *IEEE Photon. Technol. Lett.*, vol. 20, pp. 78-80, 2008.
- [94] M. H. Khan, H. Shen, Y. Xuan, L. Zhao, S. J. Xiao, *et al.*, "Ultrabroad-bandwidth arbitrary radiofrequency waveform generation with a silicon photonic chip-based spectral shaper," *Nature Photon.*, vol. 4, pp. 117-U30, Feb. 2010.
- [95] J. Q. Zhou, S. Fu, F. Luan, S. Aditya, P. P. Shum, *et al.*, "Tunable multi-tap bandpass microwave photonic filter using a windowed Fabry-Pérot filter based multi-wavelength tunable laser," *J. Lightwave Technol.*, Accepted.
- [96] G. Ning, P. Shum, C. Lu, J. Q. Zhou, V. Wong, *et al.*, "Multiple dual-wavelengths erbium-doped fiber laser," in *2008 Conference on Optical Fiber Communication/National Fiber Optic Engineers Conference (OFC/NFOEC2008)*, Vols 1-8, 2008, pp. 403-405.
- [97] M. A. F. Roelens, S. Frisken, J. A. Bolger, D. Abakoumov, G. Baxter, *et al.*, "Dispersion Trimming in a Reconfigurable Wavelength Selective Switch," *J. Lightwave Technol.*, vol. 26, pp. 73-78, 2008.
- [98] B. Moslehi and J. W. Goodman, "Novel amplified fiber-optic recirculating delay line processor," *J. Lightwave Technol.*, vol. 10, pp. 1142-1147, 1992.
- [99] B. Moslehi, "Fibre-optic filters employing optical amplifiers to provide design flexibility," *Electron. Lett.*, vol. 28, pp. 226-228, 1992.
- [100] B. Vizoso, C. Vazquez, R. Civera, M. A.-L.-A. Lopez-Amo, M., and M. A. A.-M. Muriel, M.A., "Amplified fiber-optic recirculating delay lines," *J. Lightwave Technol.*, vol. 12, pp. 294-305, 1994.
- [101] E. H. W. Chan and R. A. Minasian, "Reflective Amplified Recirculating Delay Line Bandpass Filter," *J. Lightwave Technol.*, vol. 25, pp. 1441-1446, 2007.
- [102] J. Q. Zhou, H. Dong, S. Aditya, P. Shum, L. Xia, *et al.*, "Reflective 3x3 coupler-based double recirculating delay line microwave photonic filter," in *Photonics 2008*, New Delhi, India, 2008.
- [103] Y. H. Ja, "A Vernier fiber double-ring resonator with a 3x3 fiber coupler and degenerate two-wave mixing," *IEEE Photon. Technol. Lett.*, vol. 4, pp. 743-745, 1992.
- [104] J. Capmany, "Amplified double recirculating delay-line using a 3X3 coupler," *J. Lightwave Technol.*, vol. 12, pp. 1136-1143, Jul 1994.
- [105] G. Rossi, T. E. Dimmick, and D. J. Blumenthal, "Optical performance monitoring in reconfigurable WDM optical networks using subcarrier multiplexing," *J. Lightwave Technol.*, vol. 18, pp. 1639-1648, Dec. 2000.
- [106] J. Q. Zhou, S. Fu, X. Li, S. Aditya, P. P. Shum, *et al.*, "Instantaneous microwave frequency measurement using an asymmetric non-linear group delay profile," in *International Topical Meeting on Microwave Photonics, 2009 (MWP '09)*, Valencia, Spain, 2009, pp. 1-3.

- [107] H. Chi, X. Zou, and J. Yao, "An Approach to the Measurement of Microwave Frequency Based on Optical Power Monitoring," *IEEE Photon. Technol. Lett.*, vol. 20, pp. 1249-1251, Jul. 2008.
- [108] S. Pan and J. P. Yao, "Instantaneous photonic microwave frequency measurement with a maximized measurement range," presented at the 2009 International Topical Meeting on Microwave Photonics (MWP 2009), Valencia, Spain, 2009.
- [109] X. H. Zou, H. Chi, and J. P. Yao, "Microwave Frequency Measurement Based on Optical Power Monitoring Using a Complementary Optical Filter Pair," *IEEE Trans. Microw. Theory Tech.*, vol. 57, pp. 505-511, Feb. 2009.
- [110] M. V. Drummond, P. Monteiro, and R. N. Nogueira, "Photonic RF instantaneous frequency measurement system by means of a polarization-domain interferometer," *Opt. Express*, vol. 17, pp. 5433-5438, Mar. 2009.
- [111] J. Q. Zhou, S. Fu, S. Aditya, P. P. Shum, and C. Lin, "Instantaneous Microwave Frequency Measurement Using Photonic Technique," *IEEE Photon. Technol. Lett.*, vol. 21, pp. 1069-1071, Aug. 2009.
- [112] J. Q. Zhou, S. Fu, P. P. Shum, S. Aditya, L. Xia, *et al.*, "Photonic measurement of microwave frequency based on phase modulation," *Opt. Express*, vol. 17, pp. 7217-7221, Apr. 2009.
- [113] E. Ciaramella, A. D'Errico, R. Proietti, and G. Contestabile, "WDM-POLSK Transmission Systems by Using Semiconductor Optical Amplifiers," *J. Lightwave Technol.*, vol. 24, pp. 4039-4046, 2006.
- [114] T.-Y. Kim, C. K. Oh, S.-J. Kim, and C.-S. Park, "Tunable Photonic Microwave Notch Filter With Negative Coefficient Based on Polarization Modulation," *IEEE Photon. Technol. Lett.*, vol. 19, pp. 907-909, 2007.
- [115] J. Q. Zhou, S. Aditya, P. P. Shum, and J. Yao, "Instantaneous Microwave Frequency Measurement Using a Photonic Microwave Filter With an Infinite Impulse Response," *IEEE Photon. Technol. Lett.*, vol. 22, pp. 682-684, May 2010.
- [116] X. Fang, H. Ji, C. T. Allen, K. D. A.-K. Demarest, and L. P. A.-L. Pelz, "A compound high-order polarization-independent birefringence filter using Sagnac interferometers," *IEEE Photon. Technol. Lett.*, vol. 9, pp. 458-460, 1997.
- [117] H. Gouraud, P. Di Bin, L. Billonnet, B. Jarry, E. Lacroizier, *et al.*, "Reconfigurable and tunable microwave-photonics bandpass-slicing filter using a dynamic gain equalizer," *Microwave Opt. Tech. Lett.*, vol. 48, pp. 562-567, Mar. 2006.
- [118] V. M. Menon, W. Tong, and S. R. Forrest, "Control of quality factor and critical coupling in microring resonators through integration of a semiconductor optical amplifier," *IEEE Photon. Technol. Lett.*, vol. 16, pp. 1343-1345, May 2004.
- [119] A. E. Willner, B. Zhang, L. Zhang, L. Yan, and I. Fazal, "Optical Signal Processing Using Tunable Delay Elements Based on Slow Light," *Selected Topics in Quantum Electronics, IEEE Journal of*, vol. 14, pp. 691-705, 2008.
- [120] M. Sagues, A. Loayssa, and J. Capmany, "Multitap Complex-Coefficient Incoherent Microwave Photonic Filters Based on Stimulated Brillouin Scattering," *IEEE Photon. Technol. Lett.*, vol. 19, pp. 1194-1196, 2007.
- [121] M. Sagues, R. G. Olcina, A. Loayssa, S. Sales, and J. Capmany, "Multi-tap complex-coefficient incoherent microwave photonic filters based on optical single-sideband modulation and narrow band optical filtering," *Opt. Express*, vol. 16, pp. 295-303, Jan 2008.

- [122] Y. Yan and J. Yao, "A Tunable Photonic Microwave Filter With a Complex Coefficient Using an Optical RF Phase Shifter," *IEEE Photon. Technol. Lett.*, vol. 19, pp. 1472-1474, 2007.
- [123] Y. Dai and J. Yao, "Nonuniformly Spaced Photonic Microwave Delay-Line Filters and Applications," *IEEE Trans. Microw. Theory Tech.*, vol. 58, pp. 3279-3289, 2010.
- [124] X. Yi, T. X. H. Huang, and R. A. Minasian, "Tunable and Reconfigurable Photonic Signal Processor With Programmable All-Optical Complex Coefficients," *IEEE Trans. Microw. Theory Tech.*, vol. 58, pp. 3088-3093, 2010.
- [125] X. Li, J. Dong, Y. Yu, and X. Zhang, "A Tunable Microwave Photonic Filter Based on an All-Optical Differentiator," *IEEE Photon. Technol. Lett.*, vol. 23, pp. 308-310, 2011.
- [126] C. D. Poole and R. E. Wagner, "Phenomenological approach to polarisation dispersion in long single-mode fibres," *Electron. Lett.*, vol. 22, pp. 1029-1030, Sep. 1986.
- [127] B. L. Heffner, "Automated Measurement of Polarization Mode Dispersion Using Jones Matrix Eigenanalysis," *IEEE Photon. Technol. Lett.*, vol. 4, pp. 1066-1069, Sep. 1992.
- [128] R. M. Jopson, L. E. Nelson, and H. Kogelnik, "Measurement of second-order polarization-mode dispersion vectors in optical fibers," *IEEE Photon. Technol. Lett.*, vol. 11, pp. 1153-1155, Sep. 1999.