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All-Optical Microwave Filters Based on Optical Phase Modulation

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ALL-OPTICAL MICROWAVE FILTERS BASED ON OPTICAL PHASE MODULATION

By

Jun Wang

A thesis submitted in partial fulfillment of the
requirements for the degree of

Master of Applied Science

Ottawa-Carleton Institute of Electrical and Computer Engineering
School of Information Technology and Engineering
Faculty of Engineering
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ABSTRACT

Implementation of all-optical microwave filters based on optical phase modulation is investigated in this thesis. Compared with optical intensity modulation, optical phase modulation has an inherent feature that the two first-order sidebands are π out of phase. This feature provides some interesting applications in all-optical microwave signal processing.

Most of the all-optical microwave filters proposed so far are based on intensity modulation under incoherent operation. There are two main limitations: first, for many applications, only a narrow-linewidth optical source is used, such as in a radio-over-fiber system, the strong coherence of the light source will result in strong interferences among the time-delayed optical signals, leading to a very unstable frequency response. A solution to this problem is to convert the RF signal from the optical domain to the electrical domain, and then using a laser array or an incoherence source to implement photonic microwave filtering. Since optical-electrical and electrical-optical conversions are required, the system is very complicated and costly; second, to achieve bandpass filtering, microwave filters with negative coefficients are needed. For all-optical microwave filters operating under incoherent condition, only optical intensity can be manipulated, which restricts the filter to have all positive coefficients. All-optical microwave filters with only positive coefficients can only function as a lowpass filter.

To overcome the limitations, in this thesis three different microwave filter architectures based on optical phase modulation are proposed and demonstrated.

The first photonic microwave filter consists of an electro-optic phase modulator (EOPM), a length of high birefringence (Hi-Bi) fiber, a 25-km single-mode fiber and a narrow linewidth laser source. Different time delays are achieved when the two orthogonal polarization modes are traveling along the Hi-Bi fiber. The baseband resonance is eliminated by use of the EOPM in combination with the 25-km single-mode fiber serving as a dispersive device. The proposed filter is immune to optical interference because of the orthogonality of the two polarization modes. A two-tap all-optical microwave bandpass filter with a null-to-null bandwidth of 8.7 GHz and a notch rejection level greater than 30 dB implemented in the 25-km radio-over-fiber link is demonstrated.

In the second and the third photonic microwave filters, we focus on the technique to obtain bipolar coefficients based on phase modulation to intensity modulation (PM-IM) conversion. In the second filter, chirped fiber Bragg gratings (CFBGs) are used as PM-IM conversion devices. Positive and negative coefficients are obtained through PM-IM conversion, by passing the phase modulated optical carriers through the CFBGs having group delay responses with positive and negative slopes. A two-tap transversal microwave filter with one negative coefficient is experimentally implemented.

In the third filter, the negative coefficients are obtained by locating the optical carriers at the opposite slopes of the transfer function of an optical filter, to convert the phase-

modulated signals to intensity-modulated signals, with phase inversion of the RF modulating signals. Based on this scheme, a two-tap microwave bandpass filter with one negative coefficient is demonstrated.

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LIST OF ACRONYMS

A

ASE	Amplified Spontaneous Emission
AWG	Arrayed Waveguide Grating

B

BPSK	Binary Phase Shift Keying
------	---------------------------

C

CFBG	Chirped Fiber Bragg Grating
------	-----------------------------

D

DPSK	Differential Phase Shift Keying
------	---------------------------------

E

E/O	Electrical to Optical
EDFA	Erbium-Doped Fiber Amplifier
EMI	Electromagnetic Interference
EOIM	Electro-Optic Intensity Modulator
EOPM	Electro-Optic Phase Modulator

F

FBG	Fiber Bragg Grating
FM	Frequency Modulation
FSR	Free Spectral Range
FWHM	Full Width Half Maximum

H

Hi-Bi	High Birefringence
-------	--------------------

I

IM	Intensity Modulation
----	----------------------

L

LD	Laser Diode
LCFBG	Linearly Chirped Fiber Bragg Grating

M

MTI	Moving Target Identification
MSR	Mainlobe to Sidelobe Ratio
MSW	Magnetostatic Wave
MZI	Mach-Zehnder Interferometer

O

O/E	Optical to Electrical
OOK	On-Off Keying
OSA	Optical Spectrum Analyzer

P

PC	Polarization Controller
PM	Phase Modulation
PM-IM	Phase Modulation to Intensity Modulation
PMD	Polarization Mode Dispersion

S

SAW	Surface Acoustic Wave
SDL	Superconducting Delay-Line
SOA	Semiconductor Optical Amplifier

X

XGM	Cross Gain Modulation
-----	-----------------------

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LIST OF PUBLICATIONS

1. J. Wang, F. Zeng, and J. P. Yao, "All-optical microwave bandpass filters implemented in a radio-over-fiber link," *IEEE Photonics Technology Letters*, vol. 17, no. 8, pp. 1737-1739, Aug. 2005.
2. F. Zeng, J. Wang, and J. P. Yao, "All-optical microwave bandpass filter with negative coefficients based on an electro-optic phase modulator and linearly chirped fiber Bragg gratings," *Optics Letters*, vol. 30, no. 17, pp. 2203-2205, Sept. 2005.
3. J. Wang, F. Zeng, and J. P. Yao, "All-optical microwave filters with negative coefficients based on PM-IM conversion," *IEEE Photonics Technology Letters*, vol. 17, no. 10, pp. 2176-2178, Oct. 2005.
4. J. Wang, and J. P. Yao, "Tunable photonic microwave filters based on all-optical mixing," *Proceedings of SPIE*, vol. 5971, Sept. 2005.

Chapter 1

INTRODUCTION

1.1 Background review

Applications in the fields such as moving target identification (MTI) radar systems, radio-over-fiber communication systems have been calling for signal processing techniques of high speed, broad bandwidth and wide dynamic range [1] [2]. Analog signal processing and digital signal processing techniques widely used nowadays though are very effective at low frequencies; processing signals having bandwidths of many gigahertz can be a real challenge for them. For instance, the surface acoustic wave (SAW) transversal filters fabricated with planar processing techniques can only operate at frequencies up to several hundred megahertz [3] [4]. Magnetostatic-wave (MSW) devices, using the propagation of slow, dispersive spin waves in low-loss ferromagnetic materials, can operate at frequencies in the range of 2-12 GHz with bandwidths on the order of 1 GHz [3]. Superconducting delay-line (SDL) filters, which make use of niobium transmission lines and proximity coupler taps, promise to offer low-loss devices with bandwidths to 20 GHz [5]. Meanwhile, the speed of digital signal processing is also less than several gigahertz at present [6] [7] [8], which is limited by the fact that the required sampling speed increases in direct proportion to the bandwidth of the signal to be processed. In addition, the electronic bottleneck is not the only source

of limitation on current signal processing techniques, electromagnetic interference (EMI) and frequency dependent losses could also result in important impairments. All-optical microwave signal processing, with several significant advantages over the approaches discussed above, such as low loss, low dispersion, light weight, high time bandwidth products, and immunity to EMI [1] [9], has been recognized as one of the promising candidates to process high frequency and wideband signals.

At the same time, with the wide deployment of digital optical communication systems having minimum channel rates of 10 Gb/s and the evolution of the Ethernet standard to encompass a transmission rate of 10 Gb/s, it is expected that microwave photonic techniques will be utilized in optical communication systems, and fiber-radio access networks will become a commercial reality in the near future [10]. Consequently, the ability of processing microwave signals directly in the optical domain, without the need of inefficient and costly intermediate conversions to and from the optical and electrical domains, can be of great practical value for future communication networks.

Motivated by above interests, many research groups have been working on this subject over the last 30 years. The first work on fiber delay-line microwave signal processing can be traced back to the seminal paper of Wilner and Van de Heuvel [9], who noted that the low loss and high modulation bandwidth of optical fibers are ideal for broadband signal processing. Following it, several experimental investigations on photonic microwave signal processing using multimode fibers were performed during 1970s [11] [12]. Between 1980 and 1990, an intensive theoretical and experimental

research work using single-mode fiber delay lines was carried by researchers at the University of Stanford [13] [14]. With the advent of some key optical components, including optical amplifiers [15-22], variable couplers [23-29], high-speed modulators [30-32] and electro-optic switches [33], more flexible structures employing these components have been put forth. Yet, the availability of fiber Bragg gratings (FBGs) [34-58] and arrayed waveguide gratings (AWGs) [59-62] has opened a new perspective toward the implementation of fully reconfigurable and tunable all-optical microwave filters.

Among all the above configurations [11-62], an electro-optic intensity modulator (EOIM) is usually used to modulate an RF signal onto an optical carrier, which is partially due to the fact that intensity modulation is a mature technology widely used in optical communication systems based on on-off-keying (OOK) formats. However, with the development of optical modulation techniques, phase modulation (PM) has become a hot topic with the renaissance of the differential-phase-shift-keying (DPSK) technique [63-65], which gives rise to the significance of exploring the optical signal processing techniques based on optical phase modulation. From the point-of-view of output spectrum, phase modulation is different from intensity modulation. For intensity modulation, the two first-order optical sidebands at the output are in phase. For phase modulation, if the modulation depth is low, only two first-order optical sidebands are needed to be considered. The two optical sidebands at the output of an electro-optic phase modulator (EOPM) are π out of phase. The beating between the optical carrier and the +1 order sideband will cancel completely the beating between the optical carrier

and the -1 order sideband. No RF signal would be detected at a photodetector. To recover the RF signal, phase modulation to intensity modulation (PM-IM) conversion is thus required. This property provides two new functions for all-optical microwave filters based on phase modulation. First, a bandpass-equivalent filter is easy to achieve because the beating between the carrier and the +1 order sideband always cancels the beating between the carrier and the -1 sideband at dc, no matter what kind of PM-IM conversion methods is employed [66-68]. A notch at dc is generated, which eliminates the lowpass resonance, leading to a bandpass filter. Second, by use of the feature that the upper and lower sidebands of a phase modulated signal are out of phase, the phase modulated optical signal can be converted to intensity modulated optical signal, with the converted RF modulating signal either in phase or out of phase with respect to the modulating RF signal. This could be used to implement microwave bandpass filters with bipolar coefficients [69] [70].

1.2 Objectives of the research

All-optical microwave filters can operate in two regimes: coherent regime and incoherent regime. For a delay-line optical filter, if the time delayed optical carriers are combined at a photodetector coherently, the filter is then said operating in the coherent regime. Although in a coherent system optical phase can be manipulated to achieve negative coefficients [71] [72], the coherent interference of the time-delayed carriers at the photodetector is extremely sensitive to environment variations, which leads to a very unstable filter response. Therefore, for most of the all-optical microwave filters,

incoherent operation is employed to obtain stable frequency responses. Under incoherent operation, however, only the intensity of the optical signal can be manipulated and hence negative taps are difficult to obtain. It is known that delay-line filters with all-positive coefficients can only function as a lowpass filter. For many applications, such as radio-over-fiber systems, bandpass filters are required. In addition, to avoid optical interference, incoherent optical sources such as broadband optical sources or an array of laser diodes are usually used. All-optical microwave filters using an incoherent light source cannot be directly deployed in a radio-over-fiber link where a telecommunication-type laser source with narrow linewidth is often employed; additional optical to electrical (O/E) and electrical to optical (E/O) conversions are required.

The objectives of this research are to find novel architectures to implement all-optical bandpass microwave filters based on optical phase modulation to overcome the two main limitations of incoherent all-optical microwave filters, i.e., the optical source must be incoherent and the coefficients are all positive. Specifically,

(1) we will investigate and implement an all-optical microwave bandpass filter operating under the incoherent regime with a coherent source (a laser diode), and the filter is expected to be directly incorporated into a radio-over-fiber system without extra O/E and E/O conversions;

(2) we will explore novel techniques to implement all-optical microwave filters with bipolar coefficients. All-optical microwave filters with bipolar coefficients can function as not only lowpass filters, but also bandpass filters with improved frequency response.

1.3 Major contribution

In this research work, three major contributions have been achieved:

1. Based on a detailed investigation of phase modulation, a generic model of all-optical microwave filters based on optical phase modulation is presented; the filter transfer function is derived. It provides the basis for the investigation of specific filter architectures based on optical phase modulation.
2. An all-optical bandpass equivalent microwave filter using a high-coherent optical source but immune to coherent interference is proposed and demonstrated. The proposed filter consists of an EOPM, a length of Hi-Bi fiber, a 25-km single-mode fiber and a narrow linewidth laser source. A two-tap all-optical microwave bandpass filter with a null-to-null bandwidth of 8.7 GHz and a notch rejection level greater than 30 dB implemented in the 25-km radio-over-fiber link is demonstrated. The filter is suitable for direct deployment in a radio-over-fiber link.
3. Two techniques to obtain bipolar coefficients based on PM-IM conversion are proposed and two all-optical bandpass microwave filters based on the two techniques are experimented. In the first approach, positive and negative coefficients are obtained by passing the phase modulated optical carriers through chirped fiber Bragg gratings

(CFBGs) having group delay responses with positive and negative slopes. In the second technique, bipolar coefficients are obtained by locating the optical carriers at the opposite slopes of the transfer function of an optical filter, to convert the phase-modulated signals to intensity-modulated signals, with phase inversion of the recovered RF modulating signals. Both of the approaches have been experimentally demonstrated with a two-tap transversal microwave filter with one negative coefficient.

1.4 Organization of this thesis

The thesis consists of six chapters. In Chapter 1, a brief review of the background of optical microwave signal processing, especially all-optical microwave filters is first presented, then the objectives and major contributions of this research are summarized. In Chapter 2, key optical components such as EOPMs, EOIMs and photodetectors are introduced and a comparison between intensity modulation and phase modulation is made. The general structure and transfer function of all-optical microwave filters based on intensity modulation and phase modulation are described in this chapter. An all-optical bandpass microwave filter, which employs a narrow linewidth laser source but without coherent limitation, is presented in Chapter 3. In Chapter 4, a review on bipolar microwave filters is performed. A novel approach to obtaining negative coefficients is presented. Theoretical analysis and experimental results of a bipolar microwave filter using CFBGs with positive and negative dispersions as PM-IM conversion devices are given. In Chapter 5, a second technique to obtain bipolar coefficients using an optical filter is presented. In this approach, bipolar coefficients are obtained by locating the

optical carriers at the negative or positive slopes of the optical filter. Lastly, a conclusion is drawn in Chapter 6 with recommendations for future work.

Chapter 2

ALL-OPTICAL MICROWAVE FILTERS – A REVIEW

In this chapter, a review of key components including EOPMs, EOIMs and photodetectors used in all-optical microwave filters is first presented. Then, a general architecture of all-optical microwave filters based on optical intensity modulation is discussed. The difference between intensity modulation and phase modulation is studied, followed by a general architecture of all-optical microwave filters based on optical phase modulation and its corresponding transfer function.

2.1 Key components

2.1.1 Electro-optic phase modulator

An EOPM is based on the electro-optic effect: an externally electrical field E applied to an optical crystal leads to a refractive index change in the crystal [73] [74]. Lithium Niobate (LiNbO_3) is one of the crystals with such an effect. The refractive index change with respect to the applied electrical field E is given

$$\delta_n(E) = -\frac{1}{2}rn^3E, \quad (2.1)$$

where $\delta_n(E)$ is the refractive index variation which is proportional to the electrical field E , r is the electro-optic effect coefficient, and n is the effective refractive index of this medium.

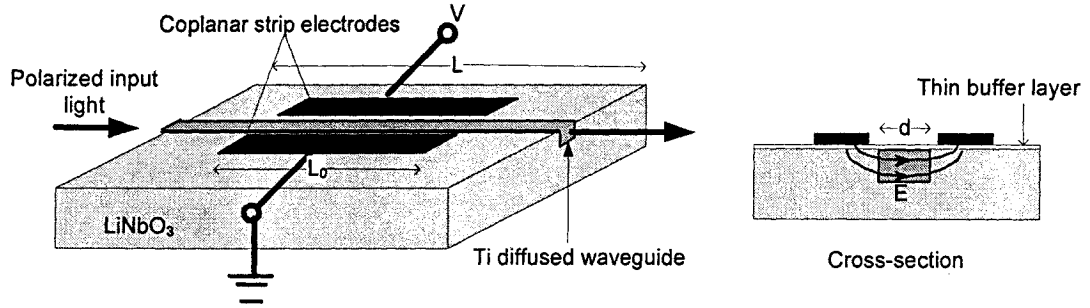


Figure 2.1 Block diagram of a LiNbO3 EOPM.

Fig. 2.1 shows a general structure of a LiNbO₃ EOPM. The modulating signal is applied to the EOPM via the electrodes, which leads to the refractive index change because of the electro-optic effect of the LiNbO₃ material. Suppose that the applied electrical field is E , the total phase shift Φ induced to the light propagating through the EOPM is given by

$$\Phi = \Phi_o + \Phi_E = \frac{2\pi n L}{\lambda_o} + \frac{2\pi \delta_n(E) L_0}{\lambda_o}, \quad (2.2)$$

where L is the total length of the EOPM, L_0 is the length of the LiNbO₃ crystal to which the electrical field E is applied, λ_o is the wavelength of the incident light,

$\Phi_o = \frac{2\pi n L}{\lambda_o}$ is the phase shift of the light after propagating through length L and Φ_E is

the phase shift induced by the refractive index change within L_0 . Substituting Eq. (2.1) into Eq. (2.2), we have the phase shift Φ ,

$$\Phi = \Phi_o - \pi \frac{rn^3 EL_0}{\lambda_o}. \quad (2.3)$$

Assuming that the applied modulating signal is

$$V(t) = V_e \cos(\omega_e t), \quad (2.4)$$

where V_e and ω_e are the amplitude and angular frequency of the modulating signal.

Using $E = -V/d$ (d is the distance separating the two faces of the crystal across which the electrical field is applied), Eq. (2.3) can be expressed in terms of the modulating signal $V(t)$ as

$$\Phi = \Phi_o + \pi \cdot \frac{V(t)}{V_\pi}, \quad (2.5)$$

where $V_\pi = \frac{d}{L_0} \cdot \frac{\lambda_o}{rn^3}$. V_π is an important parameter related to the EOPM, namely half-

wave voltage, i.e. a voltage at which Φ_E equals to π .

If we denote the electrical field of the input optical signal as

$$E_{in}(t) = E_o \cos(\omega_o t), \quad (2.6)$$

where E_o and ω_o are the amplitude and angular frequency of the optical carrier, then, the modulated optical signal at the output of the EOPM is

$$E_{out}(t) = E_o \cos[\omega_o t + \Phi_o + \frac{V_e \cos(\omega_e t)}{V_\pi} \cdot \pi]. \quad (2.7)$$

Usually, the fixed phase shift Φ_o can be ignored and only the phase shift caused by the modulating signal will be considered. In this situation, the output optical electrical field of the modulator can be simplified as

$$E_{out}(t) = E_o \cos[\omega_o t + \frac{V_e \cos(\omega_e t)}{V_\pi} \cdot \pi]. \quad (2.8)$$

2.1.2 Electro-optic intensity modulator

An EOIM can be achieved by use of an EOPM in one arm of a Mach-Zehnder interferometer (MZI), as shown in Fig. 2.2.

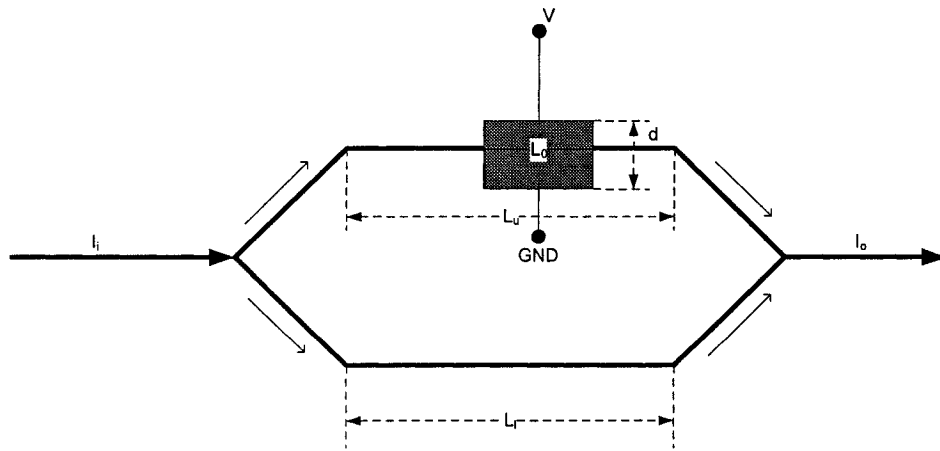


Figure 2.2 A Mach-Zehnder interferometer based EOIM.

Assuming that the EOPM is located at the upper branch, when the applied modulating signal is $V(t) = V_e \cos(\omega_e t)$, according to Eq. (2.5), the light propagating through the upper arm will experience a phase shift Φ_u expressed by

$$\Phi_u = \Phi_{uo} + \Phi_E = \frac{2\pi n L_u}{\lambda_o} + \pi \cdot \frac{V(t)}{V_\pi}, \quad (2.9)$$

where L_u is the total length of the upper arm, and Φ_{uo} is the phase shift of the light after propagating along the upper arm.

The phase shift Φ_l induced by the lower arm is

$$\Phi_l = \frac{2\pi n L_l}{\lambda_o}, \quad (2.10)$$

where L_l is the total length of the lower arm.

If the input light described by Eq. (2.6) is distributed equally into the two arms of the modulator, the electrical field E_{out} at the output of the modulator is

$$\begin{aligned} E_{out} &= \frac{1}{2} E_o \cos(\omega_o t + \Phi_u) + \frac{1}{2} E_o \cos(\omega_o t + \Phi_l) \\ &= E_o \cos(\omega_o t + \frac{\Phi_u + \Phi_l}{2}) \cos(\frac{\Phi_u - \Phi_l}{2}) \\ &= E_o \cos(\omega_o t + \frac{\Phi_u + \Phi_l}{2}) \cos[\frac{\pi n}{\lambda_o} (L_u - L_l) + \frac{\pi}{2} \cdot \frac{V(t)}{V_\pi}] \end{aligned} \quad (2.11)$$

For clarity, we denote the phase difference $\Phi_u - \Phi_l$ between the two arms as Φ and

the fixed term $\frac{2\pi n}{\lambda_o}(L_u - L_l)$ as Φ_0 . Therefore, Eq. (2.11) can be simplified as

$$\begin{aligned} E_{out} &= E_o \cos(\omega_o t + \frac{\Phi_u + \Phi_l}{2}) \cos(\frac{\Phi}{2}) \\ &= E_o \cos(\omega_o t + \frac{\Phi_u + \Phi_l}{2}) \cos[\frac{\Phi_0}{2} + \frac{\pi}{2} \cdot \frac{V(t)}{V_\pi}] \end{aligned} \quad (2.12)$$

Correspondingly, the light intensity I_{out} at the output of the modulator is

$$I_{out} = I_{in} \cos^2(\frac{\Phi}{2}). \quad (2.13)$$

The transmittance of this modulator, $T(V)$, defined as the ratio between the output optical intensity and input optical intensity, is given

$$T(V) = \cos^2(\frac{\Phi}{2}) = \cos^2(\frac{\Phi_0}{2} + \frac{\pi}{2} \cdot \frac{V(t)}{V_\pi}). \quad (2.14)$$

Fig. 2.3 shows the transmittance versus the applied voltage. It can be found that when $\Phi_0 = \pi/2$, the modulator is operating at the linear region around $T(V) = 0.5$.

When the modulator is operating at the linear region, i.e. around point B in Fig. 2.3, the device acts as a linear intensity modulator, then Eq. (2.13) can be approximated as

$$I_{out} = \frac{I_{in}}{2} [1 + m_i \cdot V(t)], \quad (2.15)$$

where $m_i \approx -\frac{\pi}{V_\pi}$ is the intensity modulation index.

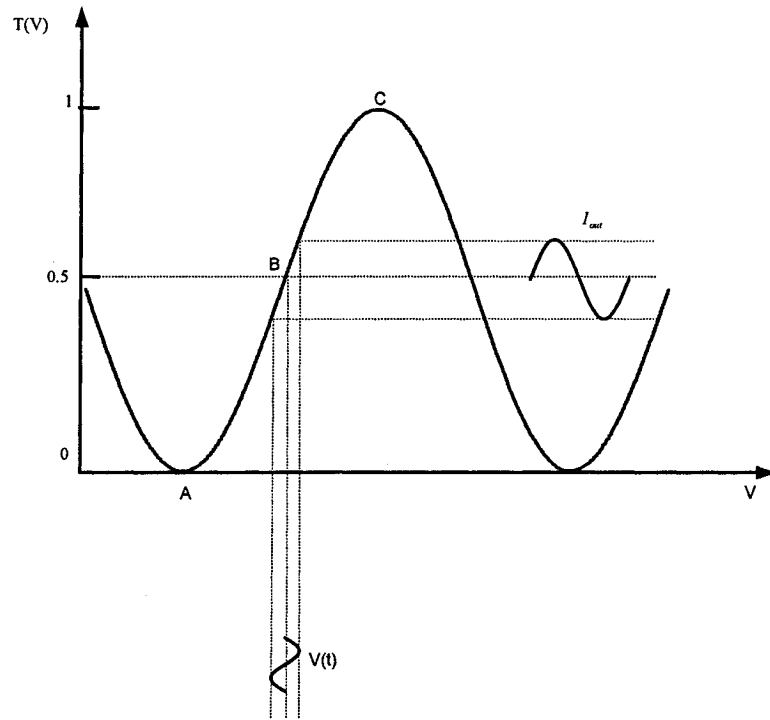


Figure 2.3 Transmittance of an EOIM.

It is worth pointing out that in Eq. (2.5) and Eq. (2.15) the values of V_π and m_i are obtained for a monochromatic source with a wavelength λ_o . For photonic microwave filters, a laser array or a broadband light source are usually used to avoid coherent interference; in this case, the light source consists of many carrier frequency components, the corresponding values of V_π and the intensity modulation index m_i for different carrier components are slightly different. However, since the carrier

components are only different from λ_o by a very small fraction of λ_o , V_π and m_i can be assumed constant.

2.1.3 Photodetector

Fig. 2.4 shows a basic structure of a PIN photodetector [73] [74]. It consists of an intrinsic semiconductor layer sandwiched between p-doped and n-doped layers. The photodetector is reversely biased to increase the thickness of the depletion region, which results in a strong internal electrical field. When a photon is incident on the photodetector, and if the photon energy is equal to the band gap of the semiconductor material, it can be absorbed to generate an electron-hole pair, photo current is thus produced.

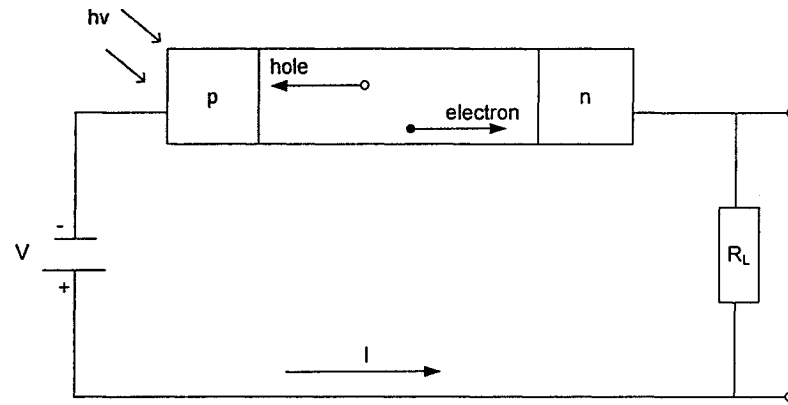


Figure 2.4 Structure of a PIN photodetector.

The photo current i_p can be expressed as a function of the input optical power P_i by

$$i_p = \frac{\eta \cdot q}{h\nu} \cdot P_i = \mathfrak{R} \cdot P_i, \quad (2.16)$$

where η is the quantum efficiency, h is the plank constant, ν is the frequency of carrier photon, and q is the charge of an electron. $\mathfrak{R} \equiv \frac{\eta q}{h\nu}$ is defined as the responsivity of the photodetector.

For example, when an optical signal described by Eq. (2.12) with $V(t) = V_e \cos(\omega_e t)$ is incident on the photodetector, the output electrical signal, either voltage or current, may be expressed as

$$y(t) = \frac{\int_{\Delta t} |E_{out}(t)|^2 dt}{\Delta t} \cdot \mathfrak{R} = A_0 + A \cdot \cos(\omega_e t), \quad (2.17)$$

where Δt is the response time of the photodetector with $2\pi/\omega_e \gg \Delta t \gg 2\pi/\omega_o$; A_0 is the dc component and A is the amplitude of the recovered RF signal. From Eq. (2.17), it is clear to see that the photodetector plays a role of an envelope detector.

2.2 All-optical microwave filters based on intensity modulation

2.2.1 General structure

A general structure of an all-optical microwave transversal filter based on intensity modulation is shown in Fig. 2.5. It consists of an optical source, usually incoherent, an EOIM, a tapped delay-line device, and a photodetector. The optical carrier, after being modulated by the input RF electrical signal $x(t)$ at the EOIM, is sent to the tapped delay-line device, where it is split into several channels with different time delays and

attenuations and then summed incoherently at a photodetector. The filtered electrical signal $y(t)$ is obtained at the output of the photodetector. Provided that the nonlinear effects in the system are small and negligible, the entire system can be considered as a linear, time-invariant system. We will show in Section 2.2.2 that the system is functioning as a microwave filter, in which the output $y(t)$ is a convolution of the input electrical signal $x(t)$ with the impulse response of the filter. The time delay unit T , which represents the time delay difference between two adjacent taps, will determine the free spectral range (FSR) of the all-optical microwave filter, and the attenuations a_k of the taps will determine the coefficients of the transfer function.

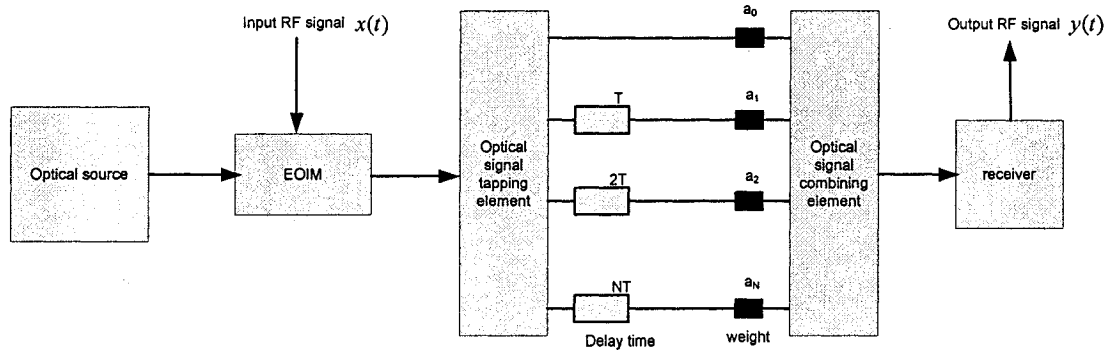


Figure 2.5 General architecture of an all-optical microwave transversal filter.

It is worth pointing out that Fig. 2.5 is not the only architecture of all-optical microwave filters. Actually, in the reported filter structures [11-62], the optical source could be a single optical source (coherent or incoherent), a laser array, or a sliced broadband source, the tapping device could be splitters, uniform FBGs or AWGs, and the time delay component could be fiber delay lines, CFBGs or a length of dispersive fiber.

Furthermore, in Chapter 1, we have mentioned that all-optical microwave filters could operate under either coherent regime or incoherent regime. It is helpful to give the definitions of coherent operation and incoherent operation here, in terms of the relationship between the coherence time τ_c of the light source used in the filter and the unit time delay T . In general, the coherence time τ_c of a light source is given by

$$\tau_c = l_c / c \approx \lambda^2 / \Delta\lambda, \quad (2.18)$$

where c is the velocity of light in free space, l_c is the coherence length, λ and $\Delta\lambda$ are the center wavelength and spectral bandwidth of the light source, respectively. For an all-optical microwave filter with a unit time delay T , if $\tau_c \gg T$ or $l_c \gg l$ (l is the length difference between any two adjacent delay lines), the filter is considered working under coherent regime and the phase of the taps plays a predominant role in the overall time and frequency response. Theoretically, in this situation, it is possible to implement negative and complex coefficients by using coherent detection. However, the optical phases of the tapped signals are highly sensitive to environmental changes, such as temperature variations, and the interferences between the time-delayed optical signals with random phase variations at the photodetector will make the filter extremely unstable. On the contrary, if $\tau_c \ll T$ or $l_c \ll l$, the filter is operating under incoherent regime. In this situation, the time-delayed optical signals will not interfere at the photodetector. The optical power at the photodetector input is simply the sum of the optical powers of all the time delayed optical signals. All-optical microwave filters

operating under this regime are free of environmental effects and thus are very stable. One severe problem related to incoherent operation is that the time-delayed microwave signals obtained at the output of the photodetector is proportional to the optical intensity, which is always positive. Therefore, all-optical microwave filters operating under incoherent regime have only positive coefficients. It is known based on signal processing theory [8] that delay-line filters with all-positive coefficients can only function as lowpass filters. For many applications, however, bandpass filters are required. The objectives of this thesis are to investigate all-optical microwave filters based on phase modulation with bandpass functionalities.

2.2.2 System transfer function

A general system transfer function of all-optical microwave filters based on intensity modulation can be obtained based on the architecture shown in Fig. 2.5. Here we only consider the case that the light source is incoherent and the filter works under incoherent regime.

Without loss of generality, we assume that the optical power of the light source is P_{in} , the EOIM is operating at the linear region. Based on Eq. (2.15), the output optical power of the EOIM, P_{EOM} , is

$$P_{EOM} = \frac{1}{2} P_{in} [1 + m_i x(t)], \quad (2.19)$$

where $x(t)$ is the modulating signal applied to the EOIM. In most of the cases, the modulated optical signal is equally split into $N + 1$ channels, and each channel has an optical power of $\frac{1}{N + 1} P_{EOM}$. The optical signals will then experience different time delays and different attenuations. Since this system is supposed to operate under incoherent regime, the optical signal power P_{com} at the photodetector is simply the sum of optical powers from the $N + 1$ channels,

$$\begin{aligned} P_{com} &= \sum_{k=0}^N \frac{1}{2} \cdot \frac{1}{N + 1} \cdot P_{in} \cdot [1 + m_i x(t - kT)] \cdot a_k \\ &= \sum_{k=0}^N \frac{P_{in} \cdot a_k}{2(N + 1)} + \sum_{k=0}^N \frac{P_{in} \cdot a_k}{2(N + 1)} \cdot m_i \cdot x(t - kT) \end{aligned} \quad (2.20)$$

where T is the time delay unit, and a_k is the attenuation index of the k -th tap.

Using Eq. (2.16), the electrical signal at the output of the photodetector is

$$y(t) = \Re \cdot \frac{P_{in} \cdot m_i}{2(N + 1)} \cdot \sum_{k=0}^N a_k \cdot x(t - kT). \quad (2.21)$$

Here only the ac component is considered and the dc component at the output of the photodetector is ignored in Eq. (2.21). In practice, the dc component can be removed by a dc blocker. Applying the Fourier transform on both side of Eq. (2.21), the system transfer function is then obtained

$$H(\omega) = K \cdot \sum_{k=0}^N a_k \cdot e^{-j\omega kT}, \quad (2.22)$$

Note that $K = \Re \cdot \frac{P_{in} \cdot m_i}{2(N+1)}$ is only a scale factor that does not affect the shape of the filter response.

Eq. (2.22) identifies a transfer function with a periodic spectral characteristic, as shown in Fig. 2.6. The frequency period is known as the FSR of the filter, which is inversely proportional to the time delay unit T . The filter selectivity measured by its quality or Q factor is given

$$Q = \frac{FSR}{\Delta\Omega}, \quad (2.23)$$

where $\Delta\Omega$ is the full width half maximum (FWHM) of the filter.

Fig. 2.6 shows the transfer functions of three different filters with different parameters. Fig. 2.6 (a) shows the transfer function of a five-tap microwave filter with identical coefficients $[1 \ 1 \ 1 \ 1 \ 1]$, and a time delay unit of $T = 75$ ps; The transfer function shown in Fig. 2.6 (b) is also a five-tap filter with identical coefficients $[1 \ 1 \ 1 \ 1 \ 1]$, but the time delay unit is $T = 100$ ps. Since the time delay unit is larger, a smaller FSR is observed. Fig. 2.6 (c) shows the transfer function of a five-tap filter with a time delay unit of $T = 100$, but with different coefficients $[0.46 \ 0.81 \ 1 \ 0.81 \ 0.46]$, which forms a Gaussian window. Comparing the simulation results, it is easy to conclude that the value of T determines the FSR, and the weight distribution, i.e. the values of a_k , forms a window function. Based on signal processing theory [8], the filter with identical coefficients can be considered as a rectangular window function, which has a poor mainlobe to sidelobe

ratio (MSR), as shown in Fig. 2.6(a) and (b); on the contrary, the filter in Fig. 2.6(c) uses a Gaussian window, which is helpful to suppress the sidelobes. We should note that the improved MSR is gained at the cost of a reduced Q factor.

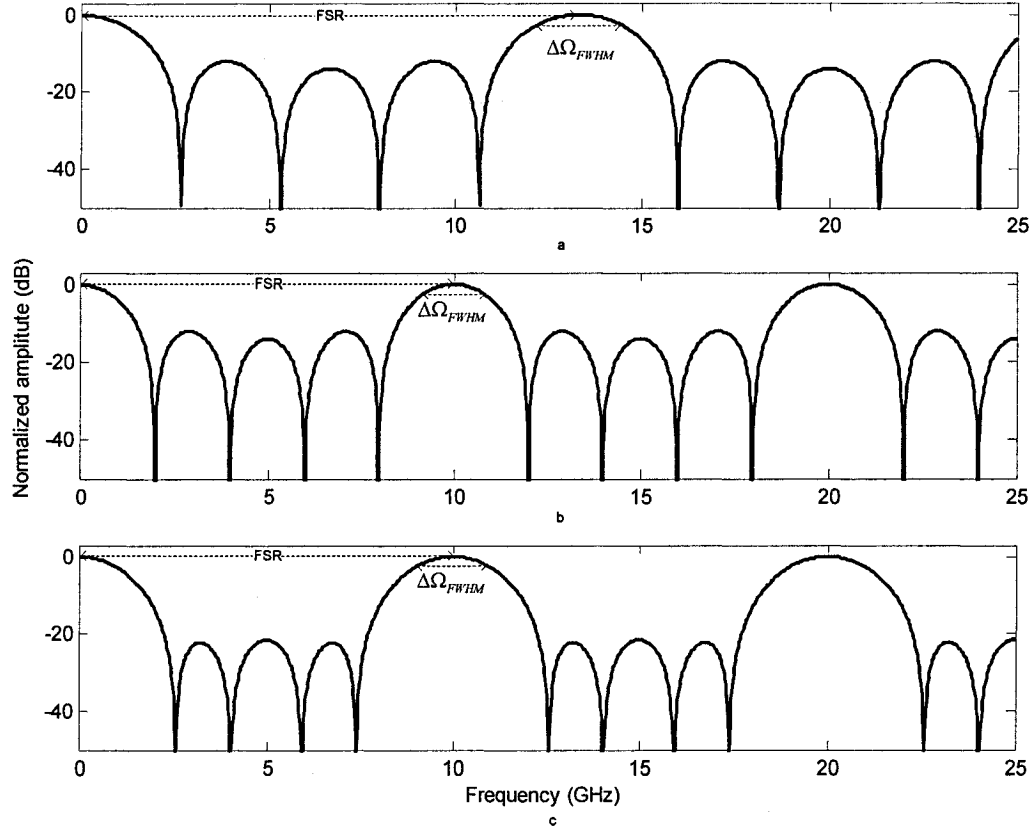


Figure 2.6 Transfer functions of three different all-optical microwave filters. (a) Coefficients: [1 1 1 1 1], $T = 75\text{ps}$; (b) Coefficients: [1 1 1 1 1], $T = 100\text{ps}$; (c) Coefficients: [0.46 0.81 1 0.81 0.46], $T = 100\text{ps}$.

2.3 All-optical microwave filters based on phase modulation

Based on the analysis in Section 2.2, it can be seen that for an all-optical microwave filters based on intensity modulation, the filtering function is mainly decided by the time

delays and attenuations applied to the channels. Actually, this is the basis for various transversal filters. However, depending on the modulation and detection techniques employed in a specific optical microwave filtering system, the filtering function may be changed.

Basically, to recover a modulated microwave signal from the optical carrier, direct detection by use of a photodetector is the simplest technique in all-optical microwave signal processing. However, a variety of methods can be employed to modulate a microwave signal onto an optical carrier. Generally, a microwave signal can be converted from the electrical domain to the optical domain by using direct modulation or external modulation. Many approaches have been proposed to realize direct intensity modulation (IM) [75], direct phase modulation (PM) [76], and direct frequency modulation (FM) [77]. In a system using direct modulation of a laser diode, three modulation products (IM, FM and PM) usually co-exist at the output of the laser diode; the system performance using direct modulation is degraded, especially for systems operating at high microwave frequencies. The use of external modulation can solve this problem. For analog applications, there are two types of external modulators: EOIMs and EOPMs. To date, all-optical microwave filters using EOIM have been explored by many researchers and many filter architectures with different functionalities have been proposed and demonstrated [11-62]. To implement all-optical microwave filters using an EOPM has recently been proposed by Zeng and Yao [66], in which an all-optical microwave filter with bandpass functionality was demonstrated. The key difference between intensity modulation and phase modulation is that the two first-order sidebands

of a phase modulated optical signal are out of phase. This unique feature makes all-optical microwave signal processing based on phase modulation an area with many new and interesting applications.

2.3.1 Phase modulation versus intensity modulation

For simplicity, in the following analysis the electrical and optical sources are considered sinusoidal. Fig. 2.7 shows a scheme of optical phase modulation with an input light source having an electrical field $E_{in}(t)$ and an electrical drive signal $V(t)$.

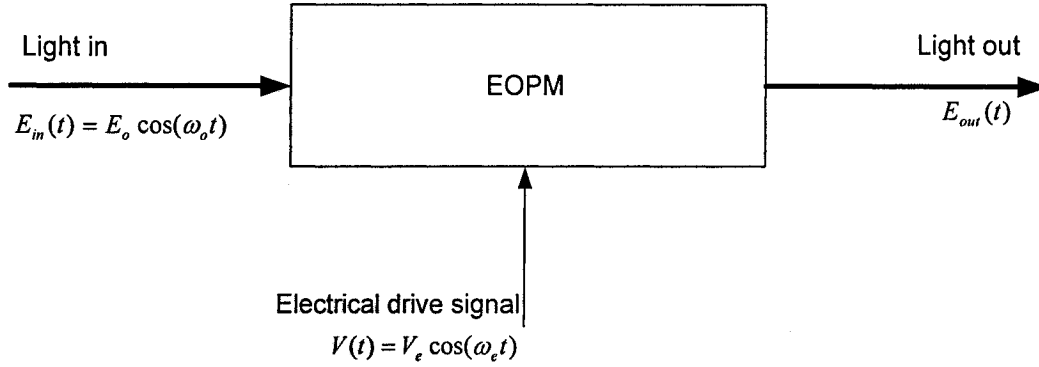


Figure 2.7 Schematic diagram of optical phase modulation.

Using Eq. (2.8), the optical signal at the output of the EOPM, $E_{out}(t)$, is given by

$$\begin{aligned}
 E_{out}(t) &= E_o \cos\left[\omega_o t + \frac{V_e \cos(\omega_e t)}{V_\pi} \cdot \pi\right] \\
 &= E_o \sum_{n=-\infty}^{\infty} J_n(\beta) \cos\left[(\omega_o + n\omega_e)t + n \cdot \frac{\pi}{2}\right]
 \end{aligned} \tag{2.24}$$

where $J_n(\beta)$ is the Bessel function of the first kind of order n with argument of β .

$\beta = \frac{\pi}{V_\pi} \cdot V_e$ is related to the phase modulation depth.

Considering the case of small signal modulation, in which the value of β is small, the higher-order sidebands in Eq. (2.24) are very small and can be neglected, therefore, Eq. (2.24) can be approximated as

$$\begin{aligned}
 E_{out}(t) &\approx E_o \{ J_0(\beta) \cos(\omega_o t) \\
 &+ J_1(\beta) \cos[(\omega_o + \omega_e)t + \frac{\pi}{2}] + J_{-1}(\beta) \cos[(\omega_o - \omega_e)t - \frac{\pi}{2}] \} . \\
 &= E_o \{ J_0(\beta) \cos(\omega_o t) \\
 &+ J_1(\beta) \cos[(\omega_o + \omega_e)t + \frac{\pi}{2}] + J_1(\beta) \cos[(\omega_o - \omega_e)t + \frac{\pi}{2}] \}
 \end{aligned} \tag{2.25}$$

The corresponding spectrum is

$$\begin{aligned}
 E_{out}(\omega) &= E_o \{ \frac{J_0(\beta)}{2} [\delta(\omega - \omega_o) + \delta(\omega + \omega_o)] \\
 &+ \frac{J_1(\beta)}{2} [j \cdot \delta(\omega - (\omega_o + \omega_e)) - j \cdot \delta(\omega + (\omega_o + \omega_e))] , \\
 &+ \frac{J_1(\beta)}{2} [j \cdot \delta(\omega - (\omega_o - \omega_e)) - j \cdot \delta(\omega + (\omega_o - \omega_e))] \}
 \end{aligned} \tag{2.26}$$

On the contrary, using Eq. (2.12), the electrical field at the output of an EOIM, as illustrated in Fig. 2.8, is given

$$E_{out} = E_o \cos(\omega_o t + \frac{\Phi_u + \Phi_l}{2}) \cos(\frac{\Phi_0}{2} + \frac{\pi}{2} \cdot \frac{V(t)}{V_\pi}). \quad (2.27)$$

Without loss of generality, the constant phase shift of the optical carrier is often neglected and Eq. (2.27) can be expanded in the form of Bessel functions as

$$\begin{aligned} E_{out}(t) = & E_o \cos(\frac{\Phi_0}{2}) \{J_0(\beta) + 2 \sum_{n=1}^{\infty} J_{2n}(\beta) \cos[2n(\frac{\pi}{2} - \omega_e t)]\} \cos(\omega_o t) \\ & - E_o \sin(\frac{\Phi_0}{2}) \{2 \sum_{n=1}^{\infty} J_{2n-1}(\beta) \sin[(2n-1)(\frac{\pi}{2} - \omega_e t)]\} \cos(\omega_o t) \end{aligned} \quad (2.28)$$

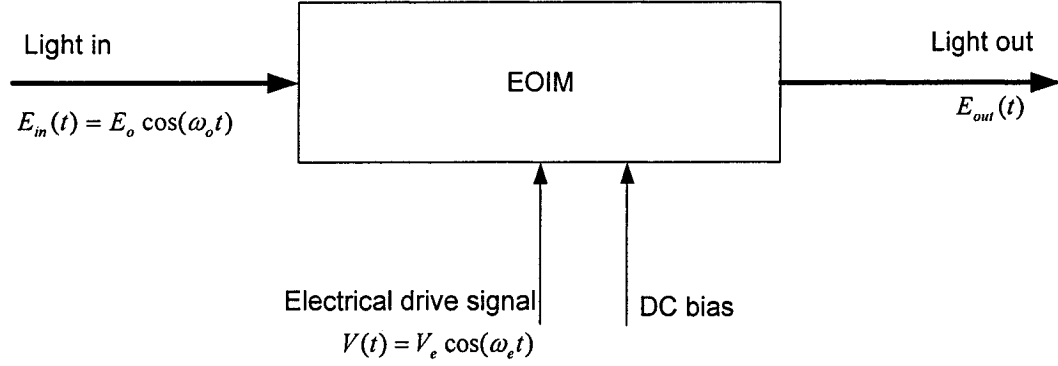


Figure 2.8 Schematic diagram of optical intensity modulation.

Same as in phase modulation, when the modulating signal is small, only the first-order sidebands need to be considered. Supposing that the modulator is operating at its linear

region, i.e. $\Phi_0 = \frac{\pi}{2}$, Eq. (2.28) can be re-written

$$\begin{aligned} E_{out}(t) \approx & \frac{\sqrt{2}}{2} E_o \{J_0(\beta) \cos(\omega_o t) \\ & - J_1(\beta) \cos[(\omega_o + \omega_e)t] - J_1(\beta) \cos[(\omega_o - \omega_e)t]\} \end{aligned} \quad (2.29)$$

Its spectrum is

$$\begin{aligned}
E_{out}(\omega) = & \frac{\sqrt{2}}{2} E_o \left\{ \frac{J_0(\beta)}{2} [\delta(\omega - \omega_o) + \delta(\omega + \omega_o)] \right. \\
& - \frac{J_1(\beta)}{2} [\delta(\omega - (\omega_o + \omega_e)) - \delta(\omega + (\omega_o + \omega_e))] \\
& \left. - \frac{J_1(\beta)}{2} [\delta(\omega - (\omega_o - \omega_e)) - \delta(\omega + (\omega_o - \omega_e))] \right\} \quad (2.30)
\end{aligned}$$

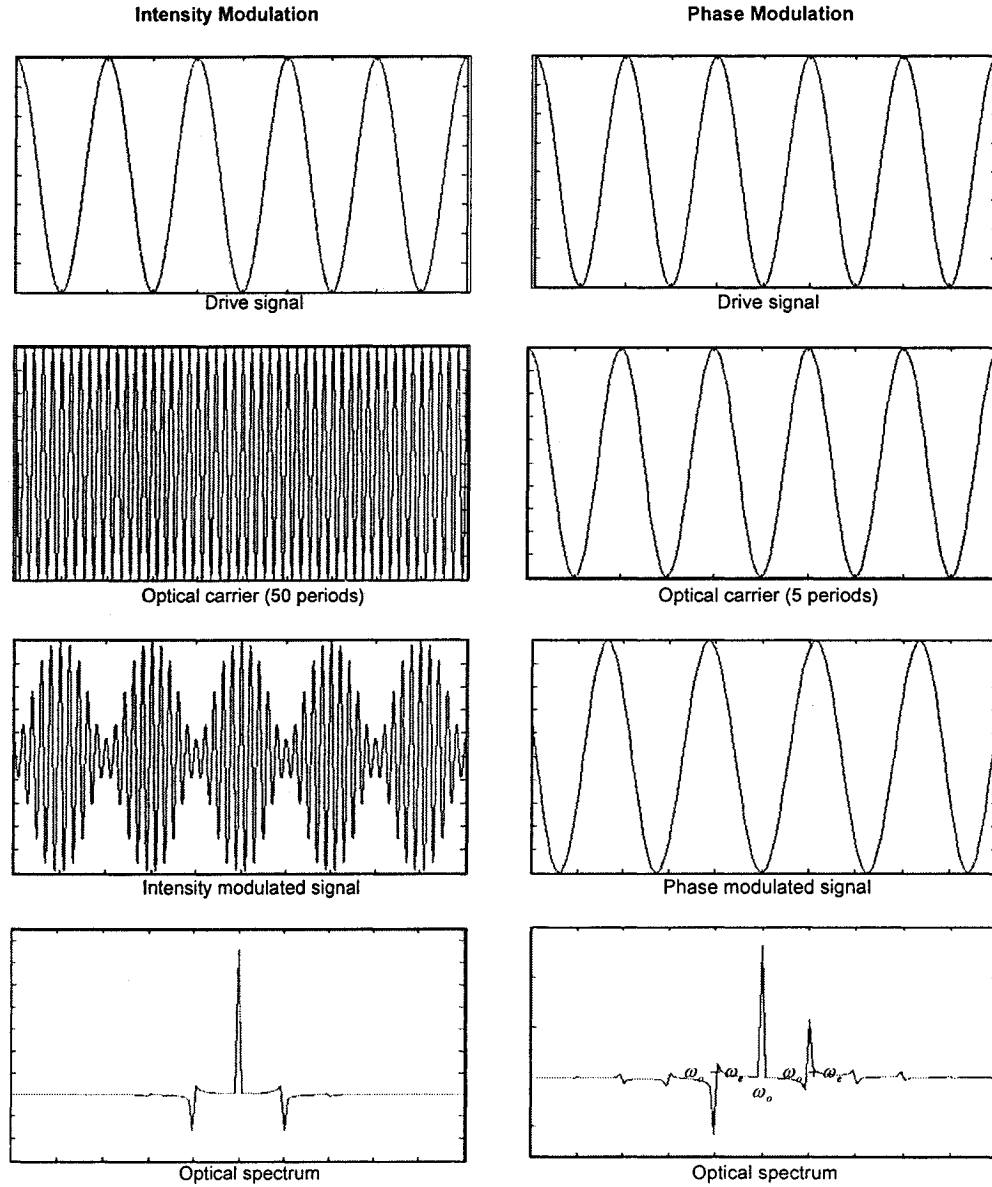


Figure 2.9 Optical spectra of an intensity modulated signal and a phase modulated signal.

Fig. 2.9 illustrates the optical signals and their spectra at the output of an EOIM and an EOPM. In the simulation, we apply the same drive signal and optical carrier for both intensity modulation and phase modulation. To make the spectra clearly, the frequency of the drive signal is one tenth of that of the optical carrier. Note that both spectra consist of a series of sidebands, each with its own amplitude and phase.

From the spectrum plots, it is interesting to note that the lower sideband and upper sideband of an intensity modulated signal are in phase, while they are out of phase for a phase modulated signal. If the intensity modulated optical signal is applied to a photodetector, photo current reflecting the modulating signal would be generated. However, if the phase modulated signal is applied to a photodetector, only a dc current would be generated. This is understandable since the beating between the optical carrier and the upper sideband will completely cancel the beating between the optical carrier and the lower sideband. The fact can also be explained using Eqs. (2.26) and (2.30).

To recover the modulating signal from a phase-modulated optical signal using a photodetector, a straightforward solution is to use PM-IM conversion. Usually, PM-IM conversion is frequency dependent. Furthermore, as we will show later, the PM-IM conversion has a frequency response equivalent to a bandpass filter. Sequentially, the combination of the transfer function of the PM-IM conversion with a transfer function of a delay-line filter can provide a frequency response with many interesting functionalities.

2.3.2 General structure and system transfer function

Fig. 2.10 shows a general architecture of an all-optical microwave filter using an EOPM. It is different from the architecture shown in Fig. 2.5, here a device to perform PM-IM conversion is employed for each tap. The transfer function of each PM-IM device is denoted as $H_k(\omega)$, where k is the tap order. Assume that the system is linear and time invariant, and if an incoherent light source is used, no optical interferences would be generated at the optical signal combining element. The transfer function of the whole system is given

$$H(\omega) \propto \sum_{k=0}^N H_k(\omega) \cdot a_k \cdot e^{-j\omega kT}. \quad (2.31)$$

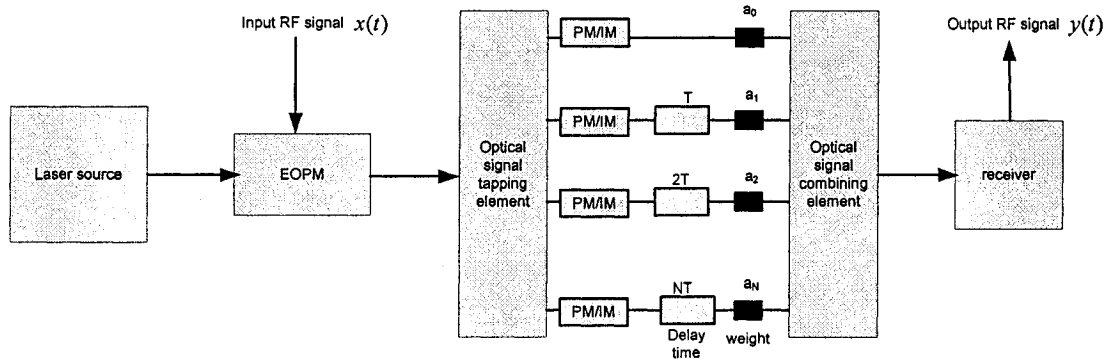


Figure 2.10 Schematic diagram of an all-optical microwave filter using an EOPM.

If $H_k(\omega)$ for all the channels are identical and represented in a unique form as

$H_{PM-IM}(\omega)$. Eq. (2.31) can thus be simplified as

$$H(\omega) \propto H_{PM-IM}(\omega) \cdot \sum_{k=0}^N a_k \cdot e^{-j\omega kT}. \quad (2.32)$$

The second term at the right-hand side is actually a transfer function of an all-optical microwave filter based on intensity modulation. If we denote the second term as $H_{IM}(\omega)$, then Eq. (2.32) becomes

$$H(\omega) \propto H_{PM-IM}(\omega) \cdot H_{IM}(\omega). \quad (2.33)$$

It can be seen that the overall transfer function of a phase-modulation-based microwave filter is equal to the transfer function of an intensity-modulation-based microwave filter multiplied by the transfer function of the PM-IM conversion.

Note that the transfer function of PM-IM conversion always has a notch at the dc frequency. Therefore, although the second term $H_{IM}(\omega)$ is a transfer function of a low-pass filter, the overall transfer function is equivalent to a bandpass filter since the resonance at the baseband is eliminated by the notch induced by the PM-IM conversion.

On the other hand, if $H_k(\omega)$ is different for each tap, for example, $H_1(\omega) = -H_2(\omega)$, all-optical microwave filters with bipolar (positive and negative) coefficients can be obtained. The fundamental principle of this can be explained by the following example: for instance, if for one tap, the upper sideband of the phase-modulated signal is completely suppressed, but for the other tap, the lower sideband of the same phase-modulated signal is completely suppressed, the recovered microwave signals from the

two taps will have inverse signs since the lower sideband and the upper sideband of the phase-modulated signal is out of phase.

In our research, to obtain two $H_k(\omega)$ that have identical amplitude responses, but opposite phase responses, two different techniques would be employed. The first technique is to pass the phase modulated multi-carrier signal through an array of CFBGs with either positive or negative dispersions; each of the CFBGs has a central reflection wavelength corresponding to the wavelength of one of the carriers. The second technique is to pass the phase modulated multi-carrier signal through an optical filter, by locating the wavelengths of the carriers at either the positive or negative slopes of the optical filter. In the following chapters, these two techniques will be investigated in details.

2.4 Summary

In this Chapter, the key components used in all-optical microwave filters, including EOPMs, EOIMs and photodetectors have been reviewed. General architectures and system transfer functions of an intensity-modulator-based microwave filter and a phase-modulator-based microwave filter have been discussed. For a phase-modulator-based microwave filter, we have the following important conclusions:

1. There is always a notch at dc for all-optical microwave filters based on phase modulation, no matter what type of PM-IM conversion is employed;

2. Bipolar coefficients can be obtained by passing a phase modulated signal through PM-IM devices with same amplitude responses but opposite phase responses;
3. The transfer function of a microwave filters based on phase modulation is a product of two transfer functions: the transfer function of the PM-IM conversion and the transfer function of a conventional delay-line filter.

Chapter 3

BANDPASS-EQUIVALENT FILTERS BASED ON PHASE MODULATION

In this Chapter, an all-optical microwave bandpass filter based on phase modulation using a single narrow linewidth laser source is proposed and demonstrated. Because the filter uses a narrow linewidth laser source, it is suitable for direct deployment in a radio-over-fiber link. In the proposed filter, a section of Hi-Bi fiber is used to obtain a time delay difference between two polarization modes along the fast and slow axes. Since the two polarization modes are orthogonal, no coherence interference would be generated at the photodetector, although a light source with high coherence is used. In the proposed filter, the PM-IM conversion is realized by using a length of single-mode fiber. Because the baseband resonance is suppressed by the notch of the PM-IM conversion, a bandpass-equivalent filter is realized. A two-tap all-optical microwave bandpass filter with a null-to-null bandwidth of 8.7 GHz and a notch rejection level greater than 30 dB is demonstrated.

3.1 Introduction

It has been pointed out in Section 2.2.1 that coherence is a major limitation in obtaining an all-optical microwave filter with stable frequency response, since the frequency response under coherent operation is extremely sensitive to environmental changes. To

obtain a stable frequency response, most of the proposed all-optical microwave filters are operating in the incoherent regime, using either a broadband light source or a laser array. In a radio-over-fiber system, however, microwave or millimeter-wave signals are usually modulated on a narrow linewidth optical carrier to avoid chromatic-dispersion-induced power penalty. For an all-optical microwave filter that can be directly incorporated into a radio-over-fiber system, the filter must be able to operate on a single high-coherent optical source. The major problem to be solved when using a coherent source, as discussed earlier, is the interferences between the time-delayed optical signals which are extremely sensitive to environmental changes.

Several techniques have been proposed to solve this problem. One approach proposed by Zhang et al. [78] is to use a length of Hi-Bi fiber. A time delay difference between the two orthogonally polarized lightwaves is obtained when the lightwaves are traveling within the Hi-Bi fiber along the two orthogonal directions with different refractive indices. No interference is observed because of the orthogonality of the two polarization modes. A stable two-tap lowpass filter was demonstrated. Based on the same idea, a bandpass filter was demonstrated by cascading two sections of Hi-Bi fibers in combination with a polarization splitter [79]. However, this scheme is under coherent operation. The bandpass filtering is achieved through optical interference. Therefore, it requires very precise polarization control and its notch rejection level is limited. More recently, Chan et al. [80] [81] demonstrated a bandpass filter based on a double-pass modulation technique, in which the light source is a telecommunication-type laser. This filter is stable with a high notch rejection level. However, a specially designed dual-

output EOIM or an extra photodetector has to be used. In addition, to achieve large FSR, the modulator must be specially designed with an FBG integrated in it.

In this Chapter, we propose a new approach to implementing all-optical microwave bandpass-equivalent filters with a narrow linewidth light source. The filter consists of an EOPM, a length of Hi-Bi fiber, a 25-km single-mode fiber and a narrow linewidth laser source. In the proposed approach, different time delays are achieved when the two orthogonal polarization modes are traveling along the Hi-Bi fiber. The baseband resonance is eliminated by use of the EOPM in combination with the length of single-mode fiber serving as a PM-IM device. The proposed filter is immune to optical interference because of the orthogonality of the two polarization modes. A two-tap all optical microwave bandpass filter with a null-to-null bandwidth of 8.7 GHz and a notch rejection level greater than 30 dB implemented in the 25-km radio-over-fiber link is demonstrated. Compared with techniques reported in [79-81], this approach does not require any specially designed components, and no complicated polarization control is needed. In addition, this approach enables stable filter operation with large FSR and high notch rejection level.

3.2 Filter architecture

3.2.1 Principle

The block diagram of the all-optical microwave bandpass filter is shown in Fig. 3.1. It consists of a laser diode, an EOPM, a length of Hi-Bi fiber, two polarization controllers (PCs), a length of standard single-mode fiber, and a photodetector.

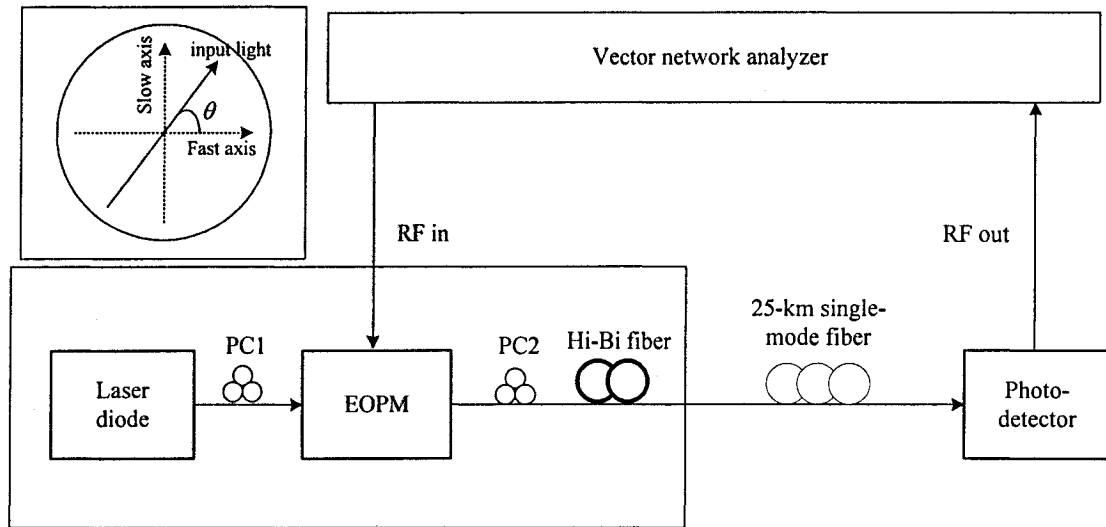


Figure 3.1 Block diagram of the all-optical microwave bandpass filter. Insert: Linearly polarized light launch into the Hi-Bi fiber.

A linearly polarized light from the laser diode (LD) is fed via the first PC to the EOPM, which is driven by a microwave signal generated by a vector network analyzer. The second PC after the EOPM is used to adjust the azimuth angle θ of the launched light with respect to the fast axis of the Hi-Bi fiber, in which two orthogonal polarization modes are excited provided that the azimuth angle is not equal to 0° or 90° , as depicted in the insert in Fig. 3.1. Because of the birefringence, the two polarization modes will experience different time delays after traveling along the Hi-Bi fiber. The time-delayed optical signals are then distributed over the standard single-mode fiber of a length of 25 km. The 25-km single-mode fiber in the proposed filter plays two roles: as a dispersive device to realize PM-IM conversion; as a transmission medium to distribute the signal. After the 25-km fiber, the optical signals are converted to electrical signals having different delays by a photodetector. A microwave filter with two taps is thus realized.

Without loss of generality, the phase-modulated signal can be given in the form of Eq. (3.1) when the modulation depth is small

$$\begin{aligned}
E_{PM_out}(t) \\
\approx E_o \{ J_0(\beta) \cos(\omega_o t) \\
+ J_1(\beta) \cos[(\omega_o + \omega_e)t + \frac{\pi}{2}] + J_1(\beta) \cos[(\omega_o - \omega_e)t + \frac{\pi}{2}] \}
\end{aligned} \tag{3.1}$$

Based on the analysis in Section 2.3.2, the transfer function of an all-optical microwave filter based on phase modulation is equal to the product of two separated components: the transfer function of the PM-IM conversion $H_{PM-IM}(\omega)$ and the transfer function of a conventional delay-line filter $H_{IM}(\omega)$. In our experiment, the PM-IM conversion is realized using 25-km single-mode fiber. Because of the chromatic dispersion of the single-mode fiber, different frequency elements in Eq. (3.1) will experience different time delays, which can be described in terms of phase delays. If the phase delays of the optical carrier, the upper sideband, and the lower sideband are ϕ_o , ϕ_u and ϕ_l respectively, the electrical field after the single-mode fiber can be given

$$\begin{aligned}
E_{SMF_out}(t) \\
\approx E_o \{ J_0(\beta) \cos(\omega_o t + \phi_o) \\
+ J_1(\beta) \cos[(\omega_o + \omega_e)t + \frac{\pi}{2} + \phi_u] + J_1(\beta) \cos[(\omega_o - \omega_e)t + \frac{\pi}{2} + \phi_l] \}
\end{aligned} \tag{3.2}$$

Accordingly, the microwave signal recovered at the output of the photodetector can be described by

$$I_{RF} = \Re \cdot 2E_o^2 J_0(\beta) J_1(\beta) \sin\left(\frac{2\phi_o - \phi_u - \phi_l}{2}\right) \cos\left(\omega_e t + \frac{\phi_u - \phi_l}{2}\right), \quad (3.3)$$

where

$$\begin{aligned} \phi_o &= \beta_o \cdot L_{SMF} \\ \phi_u &= \beta_o \cdot L_{SMF} + \beta_o' \cdot L_{SMF} \cdot \omega_e + \frac{1}{2} \beta_o'' \cdot L_{SMF} \cdot \omega_e^2, \\ \phi_l &= \beta_o \cdot L_{SMF} - \beta_o' \cdot L_{SMF} \cdot \omega_e + \frac{1}{2} \beta_o'' \cdot L_{SMF} \cdot \omega_e^2 \end{aligned} \quad (3.4)$$

In Eq. (3.4), L_{SMF} is the length of the standard single-mode fiber; $\beta_o, \beta_o', \beta_o''$ are the propagation constants of the optical carrier and its first- and second-order derivatives,

$$\begin{aligned} \beta_o &= \frac{2\pi n_{eff}}{\lambda_o} \\ \beta_o' &= \left. \frac{d\beta}{d\omega} \right|_{\omega=\omega_o}, \\ \beta_o'' &= \left. \frac{d^2\beta}{d\omega^2} \right|_{\omega=\omega_o} = D \cdot \frac{\lambda_o^2}{2\pi c} \end{aligned} \quad (3.5)$$

where n_{eff} is the effective refractive index of the standard single-mode fiber, λ_o is the wavelength of the optical carrier, and D is the chromatic dispersion. Substituting Eq. (3.5) to Eq. (3.4) and normalizing the recovered RF signal, we have

$$I_{RF} = M \cdot E_o^2 \cdot \sin \frac{D \cdot \lambda_o^2 \cdot L_{SMF} \cdot \omega_e^2}{4\pi c} \cdot \cos(\omega_e t + \beta_o' \cdot L_{SMF} \cdot \omega_e). \quad (3.6)$$

For convenience, M is used here to represent the constant item $-2\Re J_0(\beta)J_1(\beta)$ in Eq. (3.6). Therefore, for $H_{PM-IM}(\omega)$, its normalized amplitude response is denoted as

$$H_{PM-IM}(\omega) = \sin \frac{D \cdot \lambda_o^2 \cdot L_{SMF} \cdot \omega^2}{4\pi c}, \quad (3.7)$$

In practice, we are more interested in the magnitude response $|H_{PM-IM}(\omega)|$, which can be directly measured using a vector network analyzer. In the remainder of the thesis, only magnitude response of a transfer function will be considered.

Fig. 3.2 shows the simulated transfer function $H_{PM-IM}(\omega)$ of the PM-IM conversion by a single-mode fiber, where $D = 17$ ps/nm/km, $\lambda_o = 1550$ nm, $L_{SMF} = 25$ km, $c = 3 \times 10^8$ m/s. It can be seen that the first notch is at dc and the second notch is at a frequency of about 17.3 GHz. Thanks to the phase rotation induced by the standard single-mode fiber, the beating between the carrier and the upper sideband is totally in phase with the beating between the carrier and the lower sideband when the RF signal is at 12.1 GHz and 21 GHz, at which the transfer function reaches the first and the second maxima in Fig. 3.2.

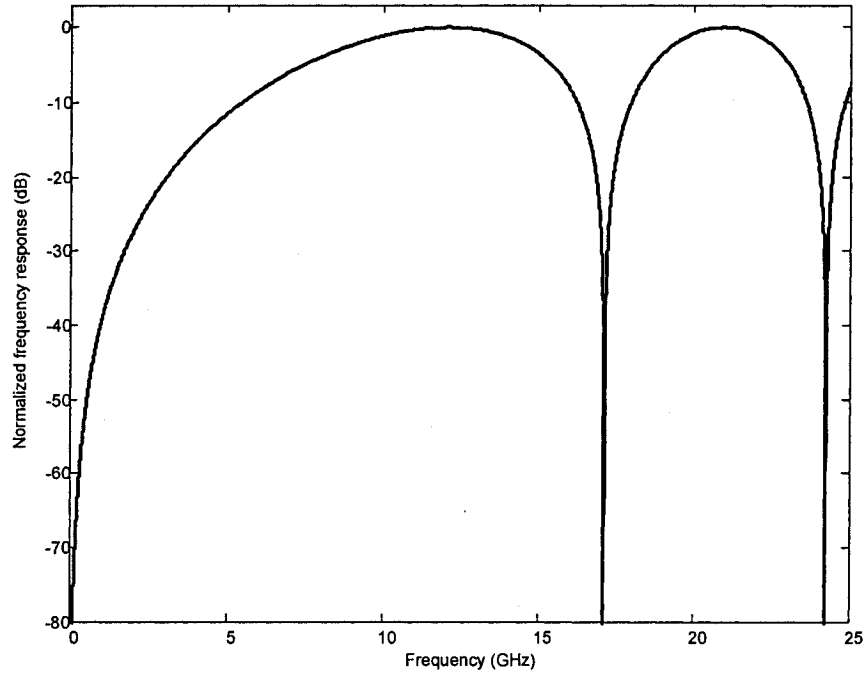


Figure 3.2 Frequency response $H_{PM-IM}(\omega)$ of the PM-IM conversion using 25-km single-mode fiber.

On the other hand, based on Eq. (2.22), the normalized frequency response of $H_{IM}(\omega)$ is

$$H_{IM}(\omega) = \cos^2 \theta + \sin^2 \theta \cdot e^{-j\omega T}, \quad (3.8)$$

where θ is the azimuth between the polarization of the light and the fast axis of the Hi-Bi fiber; $T = \frac{\Delta n \cdot L_{PMF}}{c}$ is the time delay difference between the two polarization modes, where Δn and L_{PMF} are respectively the birefringence and the length of the Hi-Bi fiber, and c is the light velocity in free space.

Using Eq. (3.7) and Eq. (3.8), the frequency response of the overall system is

$$H(\omega) = \underbrace{\sin \frac{D\lambda_o^2 \omega^2 L_{SMF}}{4\pi c}}_{H_{PM-IM}(\omega)} \cdot \underbrace{[\cos^2 \theta + \sin^2 \theta \cdot e^{-j\omega T}]}_{H_{IM}(\omega)}. \quad (3.9)$$

It is worth pointing out that the polarization mode dispersion (PMD) of the single-mode fiber and the chromatic dispersion of the Hi-Bi fiber are very small, and are not taken into account in the above analysis.

3.2.2 Experimental results

Table 3.1 is a list of the components used in the experiment.

Table 3.1 List of components used in the experimental setup in Fig. 3.1.

Laser diode	Center wavelength: 1550 nm Linewidth: 150 kHz
EOPM	Working frequency: 0-10 GHz
EOIM	Working frequency: 0-20 GHz
Hi-Bi fiber	Length: 42 m Beating length: 3.75 mm
Standard single-mode fiber	Length: 25 km Dispersion: 17 ps/nm/km at 1550 nm

Photodetector	Working frequency: 0-20 GHz
Vector network analyzer	Sweeping frequency range: 45 MHz-50 GHz Power: 3 dBm

The measured transfer function of the PM-IM conversion ($H_{PM-IM}(\omega)$) using 25-km standard single-mode fiber in combination with an EOPM is shown in Fig. 3.3. As can be seen, the first and second notch are respectively located at dc and 17.3 GHz, which agree well with the theoretical results in Fig. 3.2.

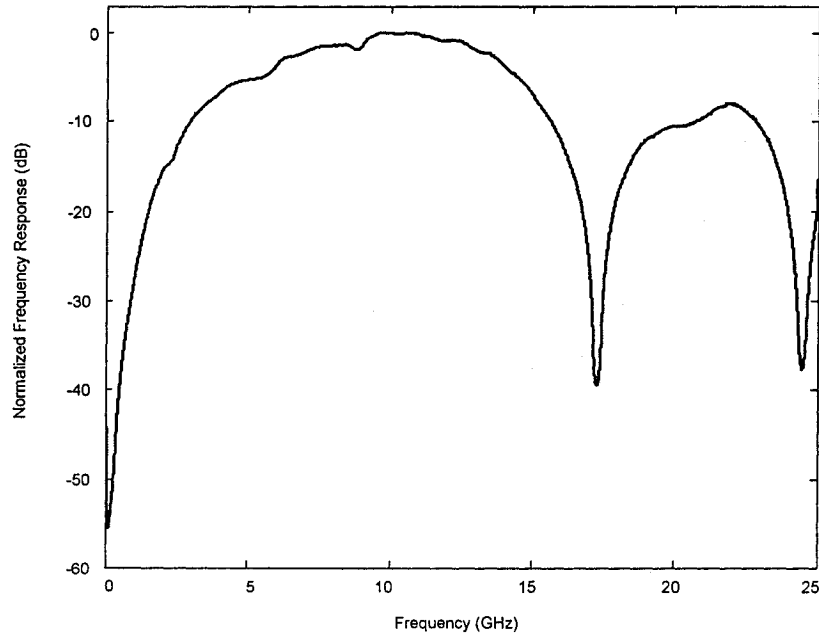


Figure 3.3 Measured frequency response $H_{PM-IM}(\omega)$ of the PM-IM conversion by 25-km SMF.

To obtain an optimized bandpass-equivalent filter, the filter is carefully designed by choosing the lengths of the Hi-Bi fiber and the single-mode fiber to let the FSR of $H_{IM}(\omega)$ equal to the value of the second notch of $H_{PM-IM}(\omega)$, i.e. 17.3 GHz. By adjusting the azimuth angle θ of the polarization of the input optical light with respect to the fast axis of the Hi-Bi fiber, the weight of each polarization can be tuned. From Eq. (3.8) we can see that the maximum notch depth appears only when the two polarization modes are equally excited, i.e. $\theta = 45^\circ$. This is verified by the experimental results shown in Fig. 3.4, where an EOIM is employed in combination with a length of Hi-Bi fiber. As can be seen from Fig. 3.4, the notch depth is dependent on the azimuth angle, and the maximal notch depth is obtained when the azimuth angle is 45° .

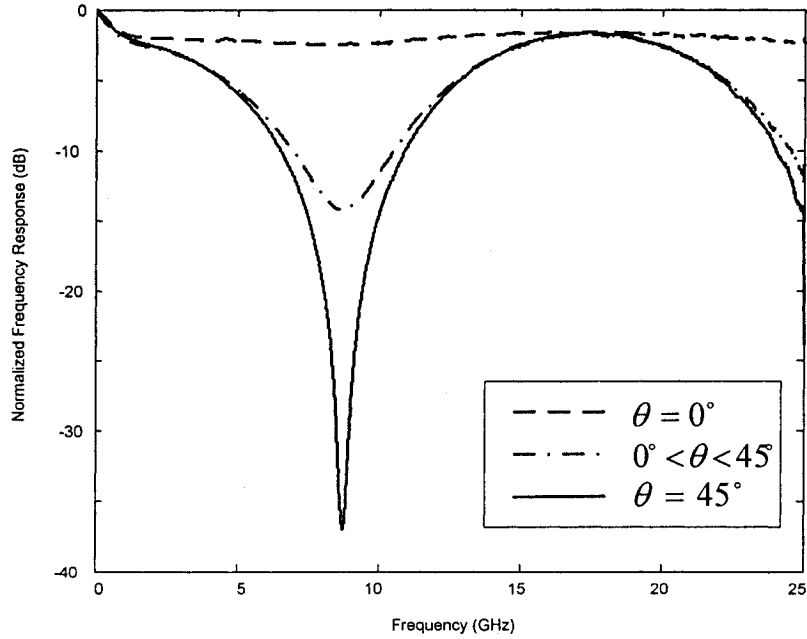


Figure 3.4 Measured frequency responses with different azimuth angles.

The overall frequency response of the proposed filter is shown in Fig. 3.5. Although the lowest measurement frequency of the vector network analyzer is limited to 45 MHz, it can be extrapolated from Eq. (3.9) that there is a notch at dc. As expected, a bandpass-equivalent filter with a null-to-null bandwidth of 8.7 GHz and a notch depth over 30 dB is obtained. Meanwhile, no obvious coherent noise generated by the optical interference between the time delayed optical signals is observed, although the linewidth of the light source is far more less than the FSR of the filter, which clearly indicates that the filter is free of limitations imposed by optical coherence. The degradation of the response in higher frequencies is caused by the unflat responses of the EOPM and the photodetector.

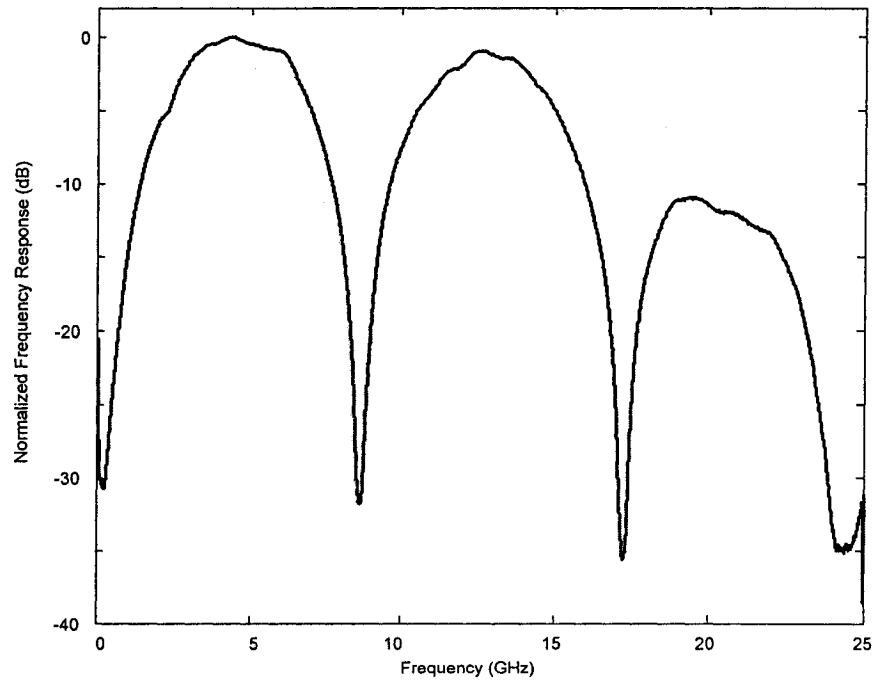


Figure 3.5 Frequency response of the proposed filter.

3.3 Further discussions

As discussed above, the demonstrated filter is obtained by matching the FSR of $H_{IM}(\omega)$ to the second notch of $H_{PM-IM}(\omega)$; therefore, in order to achieve the tunability, both $H_{PM-IM}(\omega)$ and $H_{IM}(\omega)$ are required to be tunable. The tunability of $H_{IM}(\omega)$ can be obtained by use of a differential group-delay module, which has six phase delay sections and each section consists of a birefringent crystal and a magneto-optic polarization switch [83]. To tune $H_{PM-IM}(\omega)$, an equivalent nonlinearly CFBG with linear dispersion, as presented in [84], could be cascaded with the single-mode fiber. Because the dispersion of the CFBG varies in the order of several hundreds of ps/nm , an acceptable tuning range of $H_{PM-IM}(\omega)$ can be achieved by using a laser source with a small wavelength tuning range. Based on this tuning scheme, it is possible to obtain a digitally tunable filter since both the laser source and the differential group-delay module can be controlled by a micro-processor.

It should be pointed out that the proposed filter does not have a periodic frequency response in the whole frequency range since the dispersion-dependant frequency response $H_{PM-IM}(\omega)$ is non-periodic. From Eq. (3.9), it is easy to find that with an increase of the modulating frequency, the null-to-null spacing of $H_{PM-IM}(\omega)$ will decrease. For practical application, however, the filter response is not required to be periodical. Therefore, it is possible to choose a suitable length of Hi-Bi fiber and the accumulated dispersion of the PM-IM conversion device, to make the overall frequency response meet the design requirement.

In Fig. 3.6, the single section of Hi-Bi fiber in Fig. 3.1 is replaced by two sections of Hi-Bi fibers, and a four-tap bandpass filter is obtained. Here, the lengths of the first and the second section Hi-Bi fibers are 62.9 m and 125.8 m, respectively. If the linearly polarized light enters the first section of Hi-Bi fiber at an azimuth of 45° , and the spliced angle between the two Hi-Bi sections is also 45° , four taps with identical weights are achieved. The time delay between the adjacent taps is determined by the shorter Hi-Bi fiber; it is about 86 ps, corresponding to an FSR of 11.7 GHz, which matches the first peak of $H_{PM-IM}(\omega)$. The frequency response of the four-tap bandpass filter is shown in Fig. 3.7. Because on each polarization plane of the second Hi-Bi fiber there are two degenerated light components, interferences will happen on each polarization plane. However, since all the optical taps travel within the same optical link, environmental perturbations imposed on the different taps are identical; therefore, a stable operation is possible.

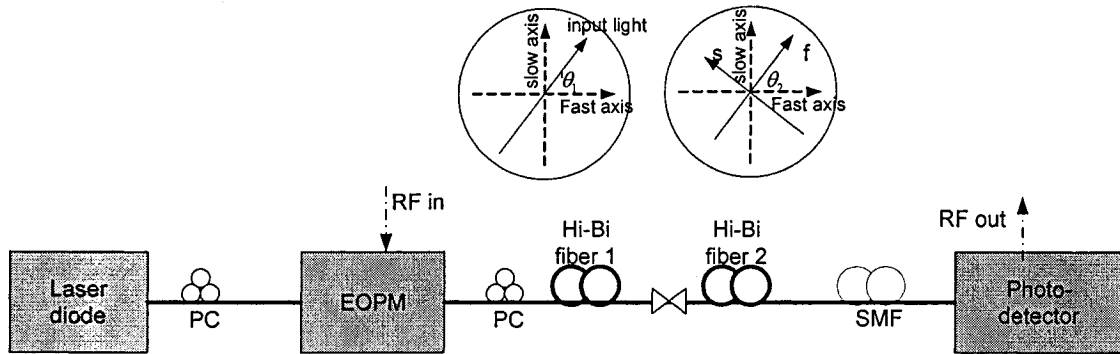


Figure 3.6 The setup of a four-tap all-optical microwave bandpass filter.

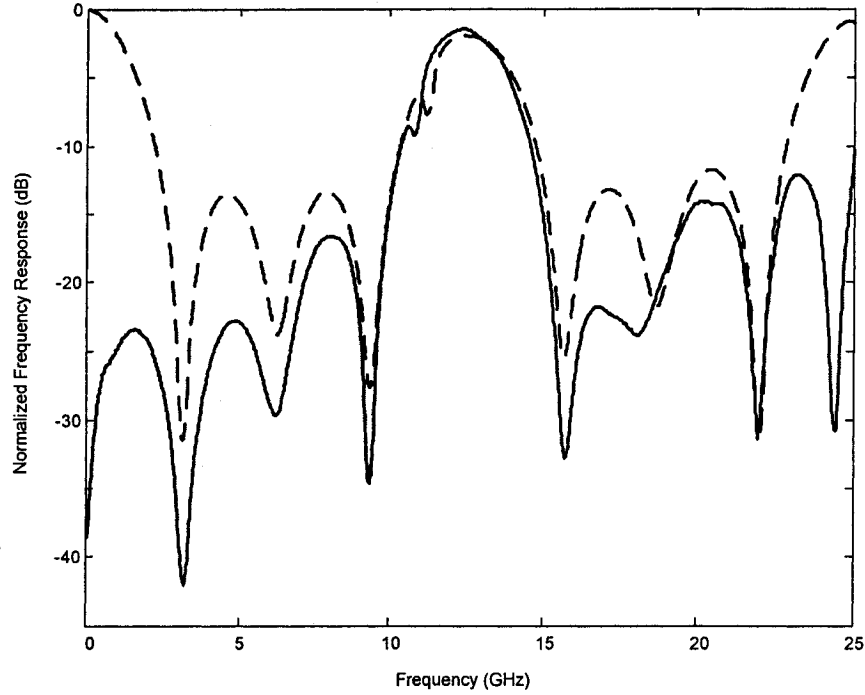


Figure 3.7 Frequency response $H(\omega)$ of the four-tap bandpass filter (solid line), frequency response $H_{LP}(\omega)$ of the corresponding low pass filter (dashed line).

3.4 Summary

In this Chapter, we have demonstrated an all-optical microwave bandpass-equivalent filter based on phase modulation. Since a narrow linewidth light source is used, the proposed filter can be deployed in a radio-over-fiber link. A stable bandpass transfer function free of optical interference was obtained by using a particularly simple structure, in which only a length of Hi-Bi fiber and a dispersive device (25-km single-mode fiber) were required after the EOPM. Bandpass filters using a section and two sections of Hi-Bi fiber corresponding to a two-tap and four-tap all-optical microwave filters were demonstrated. There are two key advantages of the proposed filter. First, the

use of a single telecommunication-type laser ensures that the proposed filter can be directly applied in a radio-over-fiber link for all-optical bandpass microwave filtering. Second, the bandpass filtering function was realized in a 25-km fiber link, which enables that the microwave or millimeter-wave signals are not only processed but also distributed by the proposed filter. In addition, the feasibility of the proposed fiber to be tunable was discussed. The architecture to implement a tunable all-optical microwave bandpass-equivalent filter was suggested as well.

Chapter 4

BIPOLAR FILTERS BASED ON PHASE MODULATION INCORPORATING CFBGs

The conventional all-optical microwave filter under incoherent operation can only have positive coefficients, as discussed in Chapter 2. This is a serious limitation, since an all-optical microwave filter with all-positive coefficients can only function as a lowpass filter [2] [8]. But, for many applications, all-optical microwave filters with passband functionality are required. In addition, with all positive coefficients it is not possible to implement filters with flat bandpass and sharp transitions.

In Chapter 3, although an EOPM combined with a dispersive device is employed to eliminate the baseband resonance, and thus to obtain a bandpass-equivalent filter, there are no negative coefficients that are actually generated in such a configuration, and bandpass filters with improved performance such as flat top and high MSR are still impossible to implement.

To have a higher flexibility, it is necessary to implement incoherent filters with negative coefficients. In this Chapter, we will first review some techniques reported on the implementation of bipolar all-optical microwave filters. Then a brief review of the FBG, a key component in all-optical microwave filters, is given. Finally, an all-optical

microwave bandpass filter with negative coefficients using an EOPM in combination with linearly chirped FBGs (LCFBGs) is proposed and experimentally demonstrated.

4.1 Bipolar all-optical microwave filters

Compared with unipolar filters, bipolar filters (filters with both positive and negative coefficients) are more flexible to achieve more functionalities. Based on signal processing theory [8], delay-line filters with all positive coefficients can only provide lowpass filtering functionality. Fig. 4.1 shows a frequency response of a seven-tap filter with coefficients $[1 \ 1 \ 1 \ 1 \ 1 \ 1 \ 1]$, in which a baseband resonance is observed. On the contrary, filters with both positive and negative coefficients can provide more versatile transfer functions. Fig 4.2 shows a frequency response of a seven-tap bipolar filter with coefficients $[-1 \ 1 \ -1 \ 1 \ -1 \ 1 \ -1]$, where the baseband resonance is eliminated and a filter with bandpass filtering functionality is obtained. Fig. 4.3 shows a frequency response of an eleven-tap filter with both positive and negative coefficients $[-0.1 \ 0 \ 0.2 \ 0 \ -0.6 \ 1 \ -0.6 \ 1 \ -0.6 \ 0 \ 0.2 \ 0 \ -0.1]$, a flat-top bandpass transfer function with sharp transitions is observed.

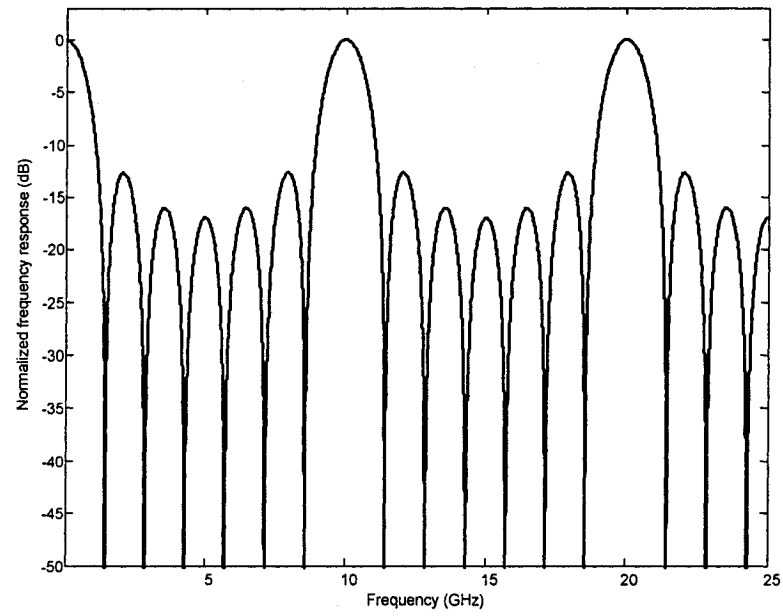


Figure 4.1 Frequency response of a filter with all positive coefficients [1 1 1 1 1 1 1].

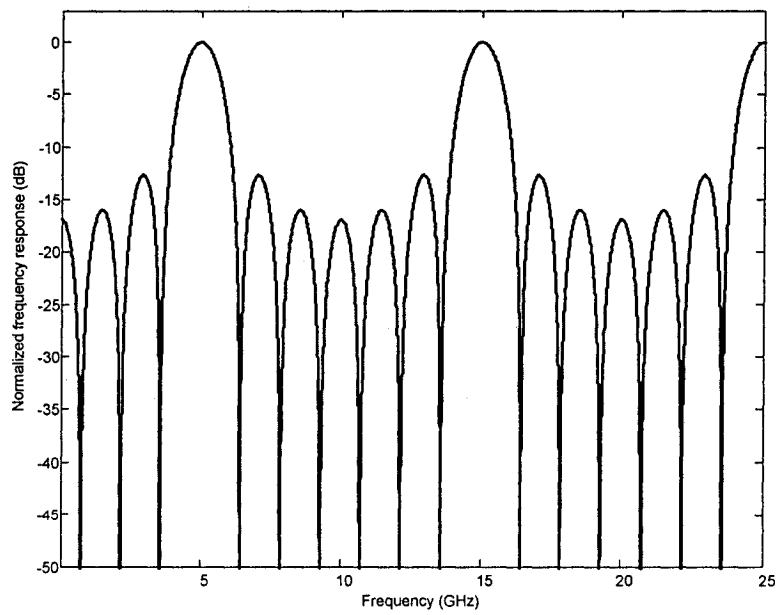


Figure 4.2 Frequency response of a filter with positive and negative coefficients [-1 1 -1 1 -1 1 -1].

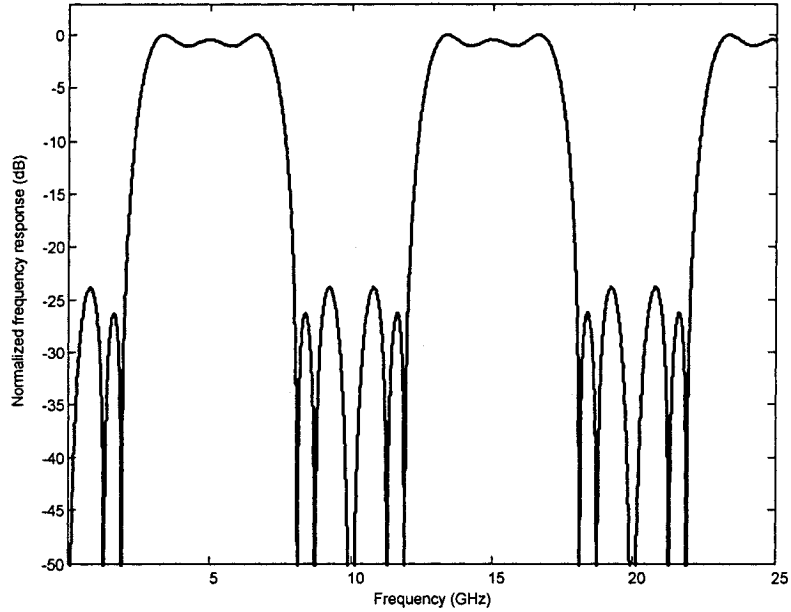


Figure 4.3 Frequency response of a filter with positive and negative coefficients $[-0.1 \ 0 \ 0.2 \ 0 \ -0.6 \ 1 \ -0.6 \ 0 \ 0.2 \ 0 \ -0.1]$.

Bipolar all-optical microwave filters can find many applications, such as in radio-over fiber systems. Since 1980s, a lot of efforts have been directed to implement incoherent all-optical microwave filters with negative coefficients. The first technique, known as differential detection, was proposed in 1984 [13]. Fig. 4.4 shows an all-optical microwave filter with bipolar coefficients implemented by using differential detection technique.

In the configuration, the tapped delay-line element is decomposed into a positive-coefficient section and a negative-coefficient section. Both sections consist of all positive taps, but the output of each section is fed to a pair of photodetectors placed in a differential configuration. Thus, the recovered RF signals from different sections have opposite signs and signal subtraction is achieved in the combiner. The applicability of

this approach has been experimentally demonstrated in [44] [85]-[87]. The differential detection technique allows the implementation of any kind of negative coefficient filters. However, it requires more components and also a careful path balance must be achieved in the microwave section before signal combination.

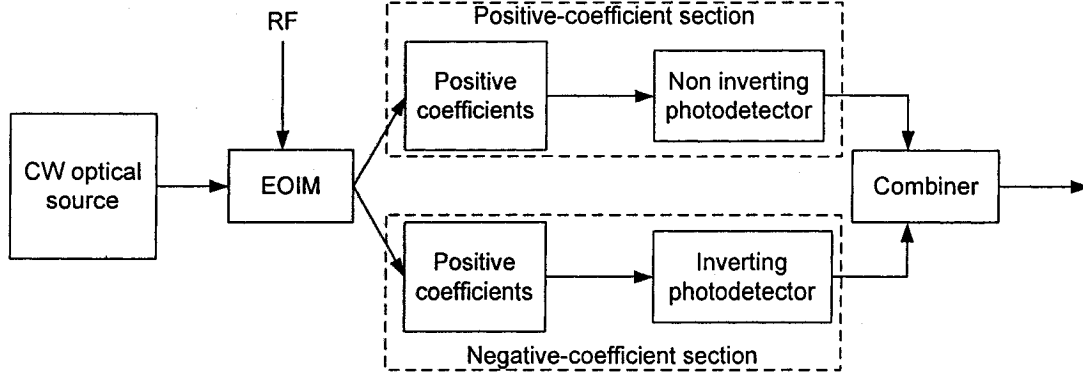


Figure 4.4 An all-optical microwave filter with bipolar coefficients implemented using differential detection technique.

Ideally, it would be better to achieve negative coefficients directly in the optical domain. In [88], negative coefficients are obtained by use of wavelength conversion based on cross gain modulation (XGM) in a semiconductor optical amplifier (SOA). Fig. 4.5 shows the proposed two-tap notch filter with a negative coefficient. As can be seen, the RF signal carried by λ_1 along the lower arm is converted to λ_2 with π phase shift due to the XGM. Similar techniques using carrier depletion effect in a Fabry-Perot laser diode [89] or in a distributed-feedback laser diode [90] have also been proposed. These techniques have the main advantage that negative coefficients are obtained directly in the optical domain. However, the filter bandwidth is limited by the conversion speed of

the SOA or the laser diode. In addition, filters with multiple taps are not easy to implement based on these techniques [88-90].

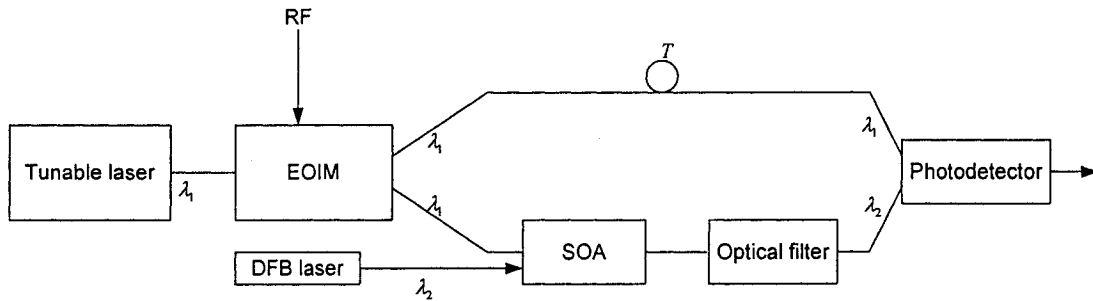


Figure 4.5 Negative coefficient generation using SOA-based wavelength conversion.

More recently, two more flexible techniques for the implementation of negative coefficients directly in the optical domain have been reported. The first one is based on the slicing of a broadband amplified spontaneous emission (ASE) source by uniform FBG filters [91]. Fig. 4.6 shows the setup of a two-tap filter using this method. Here the positive coefficient is implemented by a tunable laser; the negative coefficient is obtained by carving notches in the EDFA ASE spectrum via an FBG. With this technique, phase inversion is directly achieved in the optical domain without bandwidth limitation, and it is easy to implement multitap filters with arbitrary coefficients provided that more laser sources and more FBGs are used. A main drawback of this approach is that this configuration requires complicated optical sources.

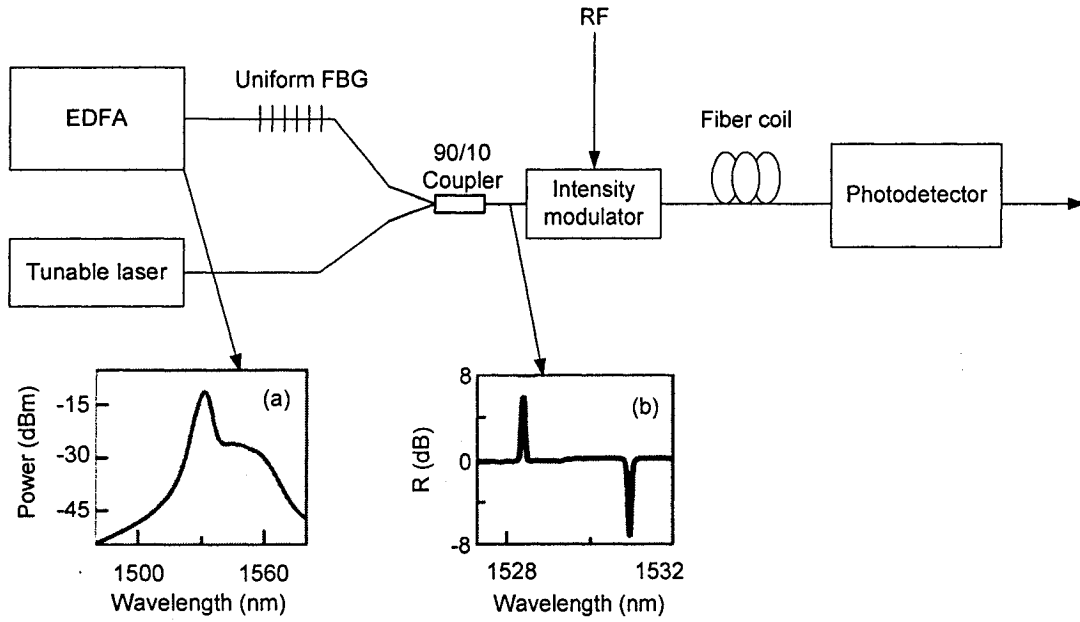


Figure 4.6 A bipolar microwave filter with negative coefficients based on the slicing of a broadband ASE source by use of a uniform FBG. Insert (a): the ASE spectrum of the EDFA. Insert (b): the modulated optical waveform.

The second technique recently reported [92] relies on a counter phase modulation in a MZI-based EOIM by biasing the modulator at the linear regions of the positive and negative slopes of the transfer function. The concept is illustrated in Fig. 4.7. Fig. 4.7(a) shows a typical transmittance function of an EOIM in terms of the applied bias voltage V_{BIAS} . Two linear modulation regions with opposite slopes centered at V_{BIAS}^- and V_{BIAS}^+ can be observed. As shown in Fig. 4.7(b), the same RF modulation signal applied to the modulator at different bias points will make the modulated optical signals with the same average power but inverted envelopes. Consequently, the recovered RF signals via a photodetector will be out of phase. This principle can be realized by using two EOIMs

biasing at the counter slopes [92]; it also can be realized by employing one EOIM if the modulator is specially designed with two input ports [93] or two output ports [94].

In this chapter and the following chapter, we will propose two approaches to implementing all-optical microwave filters with negative coefficients based on optical phase modulation. Compared with the techniques used in [13] [44] [85-94], the filters proposed in our research have a simpler structure with higher scalability.

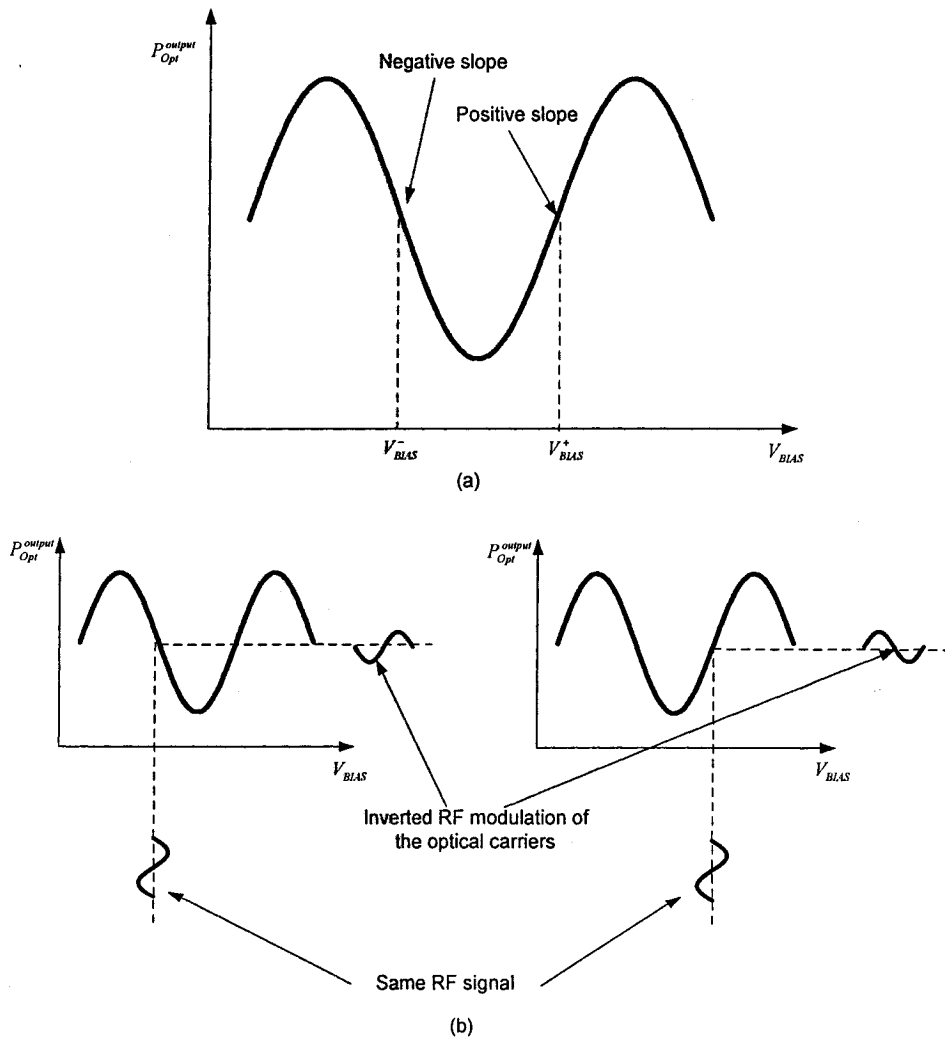


Figure 4.7 RF signal inversion in an MZI-based EOIM.

4.2 Fiber Bragg gratings

Since FBGs are key components in all-optical microwave filters we proposed. In this Section, we will give a brief review of FBGs. Both uniform and chirped FBGs will be discussed.

An FBG is a passive optical device that has periodic perturbation of the refractive index along the fiber length. An FBG is fabricated by exposing a UV (ultra violet) interference pattern to the fiber core from the transverse direction, to change the refractive index of the fiber core periodically.

Mathematically, the refractive index variation profile of a uniform FBG $\delta n_{eff}(z)$ can be written as [95]

$$\delta n_{eff}(z) = \overline{\delta n_{eff}}(z) \left\{ 1 + \nu \cos\left[\frac{2\pi}{\Lambda} z + \phi(z)\right] \right\}, \quad (4.1)$$

where $\overline{\delta n_{eff}}$ is the dc index change spatially averaged over a grating period, ν is the fringe visibility of the index change, Λ is the period of the grating and $\phi(z)$ describes the grating chirp. For a uniform FBG, $\phi(z)$ is a constant.

Fiber gratings can be broadly classified into two types: Bragg gratings and transmission gratings (long period gratings). In Bragg gratings, the coupling occurs from a forward propagating guided mode to a backward propagating mode. In transmission gratings, the

coupling occurs between modes traveling in the same direction, from the core mode to the cladding mode. In this thesis, only Bragg gratings are employed.

Basically, the coherent interference of partial reflectance within an FBG creates a bandpass reflection response and stop-band transmission response. The center wavelength of the reflection is called Bragg wavelength, which is related to the grating period by

$$\Lambda = \lambda_B / 2n_{eff}, \quad (4.2)$$

where λ_B is the Bragg wavelength and n_{eff} is the effective refractive index.

Coupled-mode theory is a powerful tool to analyze the properties of FBGs. The coupled-mode equations [95] used to describe the dominant interaction occurring between a mode of amplitude $A(z)$ and an identical counter-propagating mode of amplitude $B(z)$ within an FBG can be written as

$$\begin{aligned} \frac{dR}{dz} &= j\hat{\sigma}R(z) + j\kappa S(z) \\ \frac{dS}{dz} &= -j\hat{\sigma}S(z) - j\kappa^* R(z) \end{aligned}, \quad (4.3)$$

where $R(z) \equiv A(z) \exp[j\delta z - \phi(z)/2]$ and $S(z) \equiv B(z) \exp[-j\delta z + \phi(z)/2]$, κ is the ‘ac’ coupling coefficient and $\hat{\sigma}$ is the general ‘dc’ self-coupling coefficient given by

$$\hat{\sigma} \equiv \delta + \sigma - \frac{1}{2} \frac{d\phi(z)}{dz}. \quad (4.4)$$

The detuning δ , which is independent of z , is expressed as

$$\delta \equiv \beta - \frac{\pi}{\Lambda} = 2\pi n_{eff} \left(\frac{1}{\lambda} - \frac{1}{\lambda_B} \right), \quad (4.5)$$

where β is the propagation constant.

In addition, for an FBG, the following simple relations exist:

$$\begin{aligned} \sigma &= \frac{2\pi}{\lambda} \overline{\delta n_{eff}} \\ \kappa &= \kappa^* = \frac{\pi}{\lambda} v \cdot \overline{\delta n_{eff}}, \end{aligned} \quad (4.6)$$

where κ^* is the conjugate of “ac” coupling coefficient κ .

If the FBG is a uniform grating, $\phi(z)$ and $\overline{\delta n_{eff}}$ are constant, and the coupled-mode equations have an analytic solution. The amplitude reflectance ρ for a uniform FBG of length L is given by [95]

$$\rho = \left. \frac{S(0)}{R(0)} \right|_{S(L)=0} = \frac{-j\kappa \sinh(\gamma L)}{\gamma \cosh(\gamma L) + j\hat{\sigma} \sinh(\gamma L)}, \quad (4.7)$$

where $\gamma \equiv \sqrt{\kappa^2 - \hat{\sigma}^2}$. And power reflection coefficient r is

$$r = |\rho|^2 = \frac{\sinh^2(\gamma L)}{\cosh^2(\gamma L) - \frac{\hat{\sigma}^2}{\kappa^2}}. \quad (4.8)$$

The group delay and dispersion of the reflected light can be derived from the phase of the amplitude reflection coefficient ρ in Eq. (4.7). If we denote $\theta_\rho \equiv \text{phase}(\rho)$, the first derivative $d\theta_\rho / d\omega$ can be identified as time delay τ_ρ , given in the form

$$\tau_\rho = \frac{d\theta_\rho}{d\omega} = -\frac{\lambda^2}{2\pi c} \frac{d\theta_\rho}{d\lambda}. \quad (4.9)$$

Likewise, the dispersion d_ρ (in ps/nm) is the change rate of time delay with respect to wavelength, which is given

$$d_\rho = \frac{d\tau_\rho}{d\lambda} = -\frac{2\pi c}{\lambda^2} \frac{d^2\theta_\rho}{d\omega^2}. \quad (4.10)$$

Fig. 4.8 and Fig. 4.9 are the calculated reflectivities and time delays of two uniform FBGs listed in Table 4.1.

Table 4.1 Parameters of two uniform FBGs.

	FBG #1	FBG #2
Grating length (cm)	1	1
Design wavelength (nm)	1550	1550
Fringe visibility v	1	1
$\overline{\delta n_{eff}}$	1e-4	4e-4

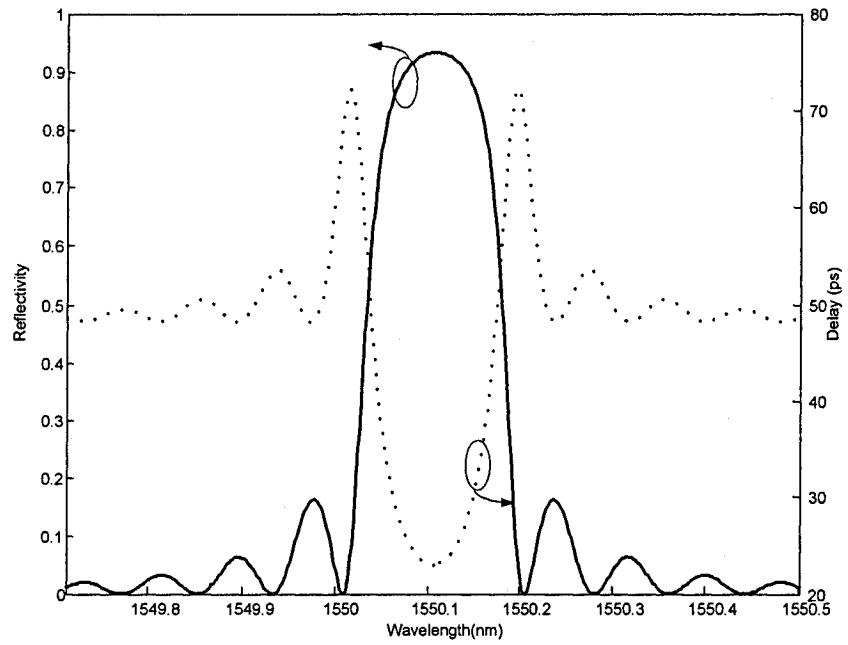


Figure 4.8 Calculated reflection spectrum and group delay for a uniform FBG with $\kappa L = 2$.

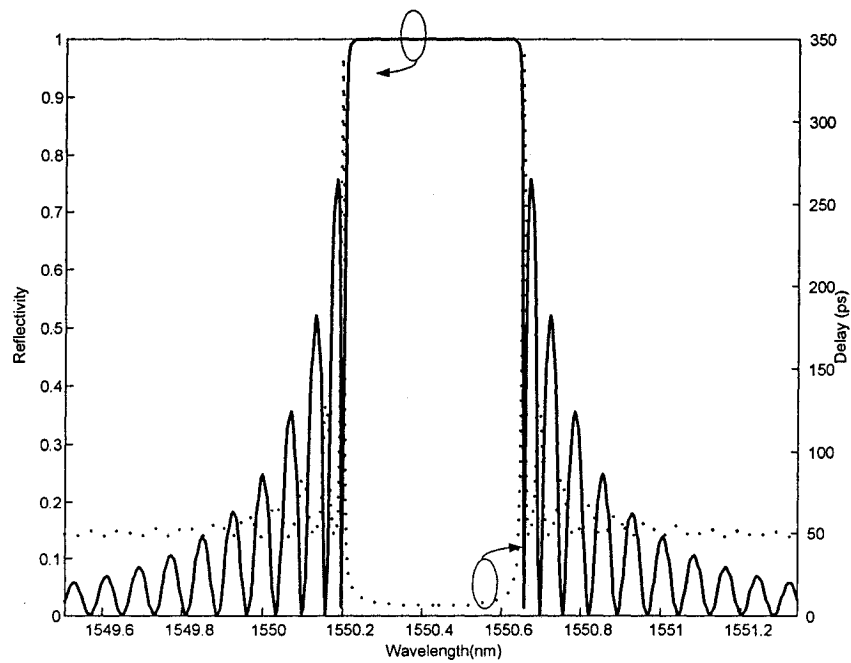


Figure 4.9 Calculated reflection spectrum and group delay for a uniform FBG with $\kappa L = 8$.

In all-optical microwave filters, uniform FBGs have been widely employed as tapping and weighting elements. CFBGs are another type of FBGs extensively used in all-optical microwave filters, which are often used as dispersive devices. In general, in a CFBG, the optical period Λ is not a constant, but varies linearly with respect to the grating length. Thus, the Bragg wavelength $\lambda_B = 2n_{eff}\Lambda$ also varies along the grating length. As a result, different frequency components of an incident optical signal are reflected at different points of the CFBG, depending on where the Bragg condition is satisfied. Briefly, for CFBGs, the Bragg grating wavelength, λ_B , can be described as a function of the axial position z of the CFBG

$$\lambda_B(z) = 2n_{eff}(z)\Lambda(z), \quad (4.11)$$

where $n_{eff}(z)$ is the effective refractive index averaged over the grating period at position z , and $\Lambda(z)$ is the grating period at position z . In practice, chirp can be easily realized by axially varying Λ . Mathematically, chirp can be simply incorporated into the coupled-mode equation as a z -dependent term in the self-coupling coefficient $\hat{\sigma}$, described by the phase term in Eq. (4.4) using

$$\frac{1}{2} \frac{d\phi(z)}{dz} = -\frac{4\pi n_{eff} z}{\lambda_D^2} \frac{d\lambda_D}{dz}, \quad (4.12)$$

where the chirp $d\lambda_D/dz$ is a measure of the change rate of the designed wavelength with respect to the position in the grating.

In engineering, the piecewise-uniform approach is often preferred for modeling CFBGs. By this means, a non-uniform grating can be taken as multiple uniform sections, and each section is identified by a 2×2 matrix. By multiplying all of these matrices, a single matrix that can describe the properties of the whole grating is obtained. Using this approach, the calculated reflectivity spectrum and group delay of a CFBG with 5 cm length, $\overline{\nu \delta n_{eff}} = 6 \times 10^{-4}$ and a linear chirp 0.4 nm/cm is shown in Fig. 4.10.

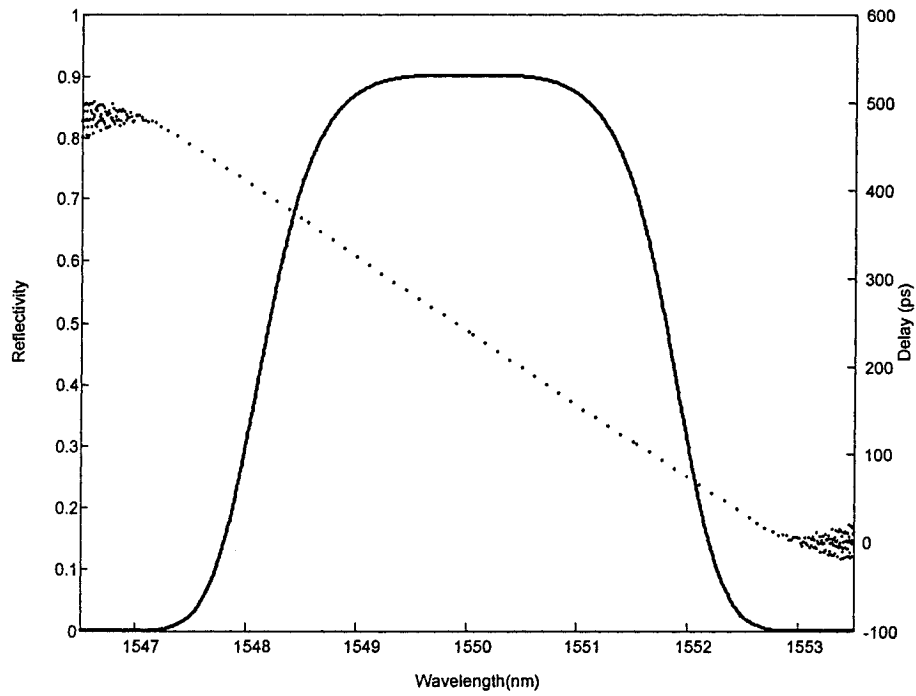


Figure 4.10 Calculated reflection spectrum and group delay for a CFBG.

4.3 All-optical microwave filters with negative coefficients based on PM-IM conversion using LCFBGs

In this section, a novel all-optical microwave bandpass filter with bipolar coefficients is presented. Positive and negative coefficients are obtained through PM-IM conversion by reflecting the phase modulated optical carriers from linearly CFBGs (LCFBGs) with positive and negative dispersions. A two-tap transversal microwave filter with one negative coefficient is experimentally implemented.

4.3.1 Principle

The principle of the proposed filter with negative coefficients is shown in Fig. 4.11. The phase modulated optical spectrum is illustrated on the left side of Fig. 4.11, which consists of an optical carrier (ω_o) and two first-order sidebands ($\omega_o - \omega_e, \omega_o + \omega_e$, where ω_e represents the modulating microwave frequency). Based on the discussion in Chapter 2, the modulation process of an EOPM generates a series of sidebands with amplitude determined by the corresponding Bessel function coefficients. However, when the modulation depth is small, the higher-order sidebands can be neglected and only the first-order upper and lower sidebands need to be considered. At the output of the EOPM, the two sidebands are π out of phase. It is different from an intensity modulation where the two sidebands at the output of an EOIM are in phase. As pointed out in Chapter 3, if the phase modulated signal is directly detected by a photodetector, the modulating signal cannot be recovered and only a dc is observed because beating between the carrier and the upper sideband exactly cancels the beating between the

carrier and the lower sideband. This behavior is expected since the phase modulation does not alter the amplitude of the input optical carrier and the square-law photodetector works like an envelope detector. However, as shown in Fig. 4.11, if the modulated optical signal passes through a dispersive device, the phase relationship between any two optical frequency components will change due to the chromatic dispersion. When this dispersed optical signal is fed to a photodetector, the RF signal can be recovered, which implies that the phase modulation is converted to intensity modulation by the dispersive device.

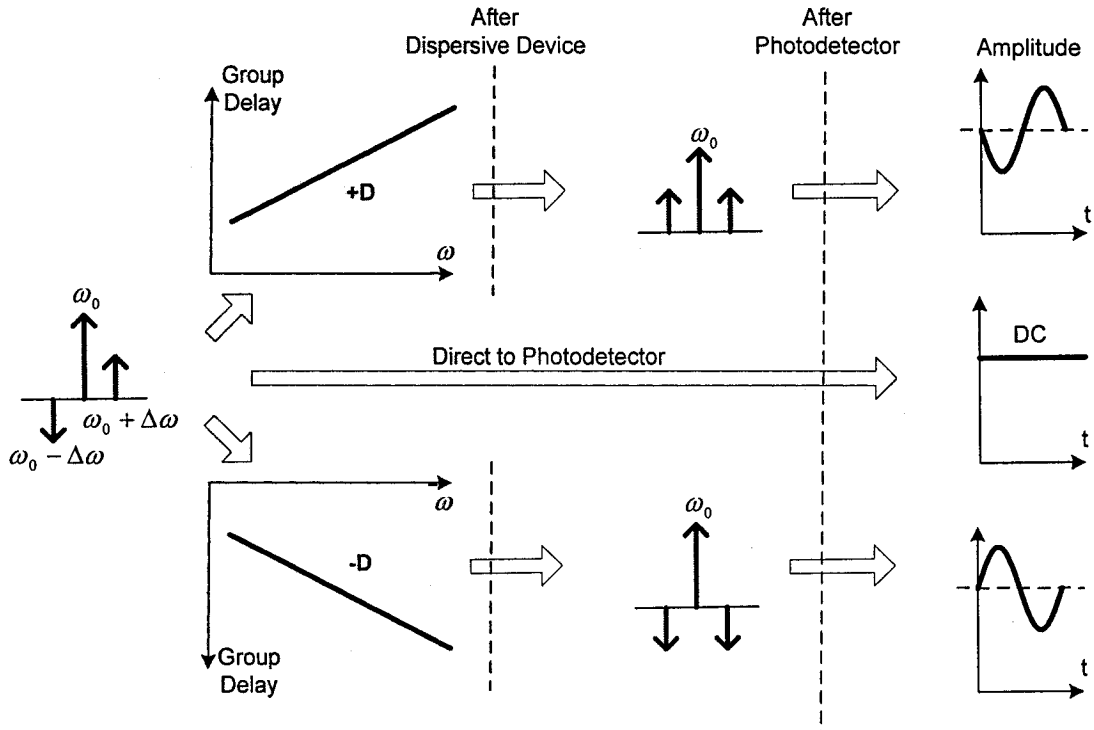


Figure 4.11 Illustration of the recovered RF modulating signals that sustain a positive, zero or negative chromatic dispersion

More interestingly, when the dispersion $D = \partial\tau / \partial\omega > 0$ (the upper case in Fig. 4.11), the higher optical frequency component experiences more phase shift than that of the

lower frequency component; and eventually the PM-IM conversion is fully achieved when all these three frequency components are exactly in phase. On the contrary, when $D = \partial\tau / \partial\omega < 0$ (the lower case in Fig. 4.11), the lower frequency component will experience more phase shift than the higher one, and the PM-IM conversion is fully obtained when the two sidebands have same phases but are π out of phase with the carrier. Consequently, the recovered RF signals from the different dispersive devices will have a π phase inversion, which can be directly applied to implement negative coefficients in an all-optical microwave filter.

Mathematically, the recovered microwave signal from such a PM-IM conversion followed by a direct detection has the same form as Eq. (3.6):

$$I_{RF} = M \cdot P_o \cdot \sin\left(\frac{1}{2}D\omega_e^2\right) \cdot \cos(\omega_e t + \phi), \quad (4.13)$$

where $M = -2\Re J_0(\beta)J_1(\beta)$ is a constant, P_o is the power of the optical carrier, D is the chromatic dispersion of the dispersive device, and ϕ is the phase delay of the recovered microwave signal, which is also determined by D and ω_e .

Based on Eq. (4.13), two important conclusions can be drawn to help us build a multi-tap microwave bandpass filter with negative coefficients. First, both positive and negative coefficients can be obtained by letting the phase modulated optical carriers experience chromatic dispersion with different signs, since $\sin(\frac{1}{2}D\omega_e^2)$ is obviously an odd function. LCFBGs are a good candidate to be used as the dispersive devices since

LCFBGs can provide very linear group delay profiles. The group delay slope of an LCFBG can be easily reversed by connecting the optical input to the opposite port of the grating. Second, the PM-IM conversion efficiency reaches the maxima when

$$\sin\left(\frac{1}{2}D\omega_e^2\right) = \pm 1, \quad (4.14)$$

which implies that the FSR of the proposed filter should be carefully designed to match the PM-IM conversion maxima; then an optimized filtering output can be obtained.

Based on the theoretical analysis, a fundamental architecture for the proposed filter is presented in Fig. 4.12. Optical carriers from an array of N LDs emitting at $\lambda_1, \lambda_2, \dots, \lambda_n, \dots, \lambda_N$ are combined via an optical combiner and applied to an EOPM. Through an optical circulator, the modulated optical signals are de-multiplexed by an AWG and fed to N LCFBGs via either the short wavelength or the long wavelength port, depending on whether the LCFBGs are employed to implement positive or negative taps. The reflected and dispersed optical signals are then multiplexed by the same AWG and sent to a photodetector to recover the modulating RF signal.

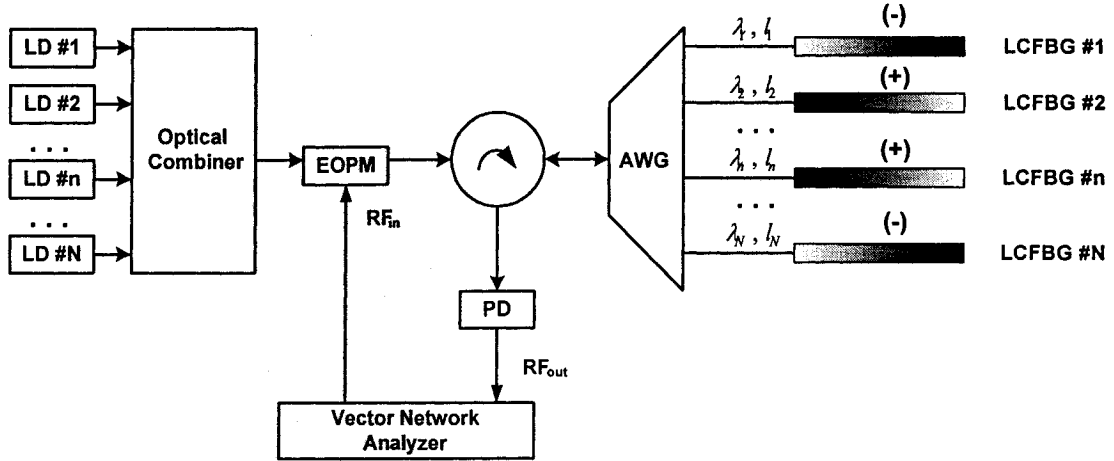


Figure 4.12 System configuration of the proposed all-optical microwave bandpass filter with negative coefficients

The recovered RF signal can be expressed as a vector summation of the resulting electrical signals from the N carriers and the frequency response of the proposed all-optical microwave filter is then written as

$$H(\omega) = M \cdot \sum_{k=1}^N P_k \cdot \sin\left(\frac{1}{2} D_k \omega_e^2\right) \cdot e^{j\omega_e(k-1)T}, \quad (4.15)$$

where P_k and D_k represent the optical power of the k -th LD and the dispersion of the k -th LCFBG, respectively. Basically, P_k determines the weight of the k -th tap and the sign of D_k determines whether this tap is positive or negative. The length difference between any two adjacent optical paths ($l_{n+1} - l_n = \Delta l, n = 1, 2, \dots, N-1$) determines the central frequency of the passband, i.e., $FSR = 1/T = c/2n_{eff} \cdot \Delta l$, where c is the optical wave propagation velocity in free space and n_{eff} is the effective refractive index.

Although a multi-channel optical coupler can be used to replace the AWG, the use of AWG can reduce the system insertion loss and at the same time eliminate the inter-tap interference. The LCFBGs are required to have different central wavelengths corresponding to those of the LD array. The lengths and chirp rates of the LCFBGs should be identical to ensure that the dispersions of the LCFBGs are identical in magnitude. In addition, the small implementation error of the delay line length of the fiber link between the AWG and each LCFBG can be accurately compensated by slightly tuning the corresponding LD wavelength to have it reflected at different positions in the LCFBG.

4.3.2 Experimental results

To prove the fundamental concept of this approach, a two-tap microwave filter with one negative coefficient is experimentally implemented. As shown in Fig. 4.13, optical carriers from two laser sources at wavelength λ_1 and λ_2 are combined via a 3-dB coupler and applied to an EOPM. Through an optical circulator and a second 3-dB coupler, the modulated optical signals are fed to two LCFBGs with opposite dispersion. Thanks to the reflectivity spectrum difference between the two gratings, in this configuration the gratings also serve as wavelength selective components and one grating only reflects one of the two input wavelengths. The reflected and dispersed optical signals are then multiplexed by the second coupler and sent to a photodetector to recover the modulating RF signal.

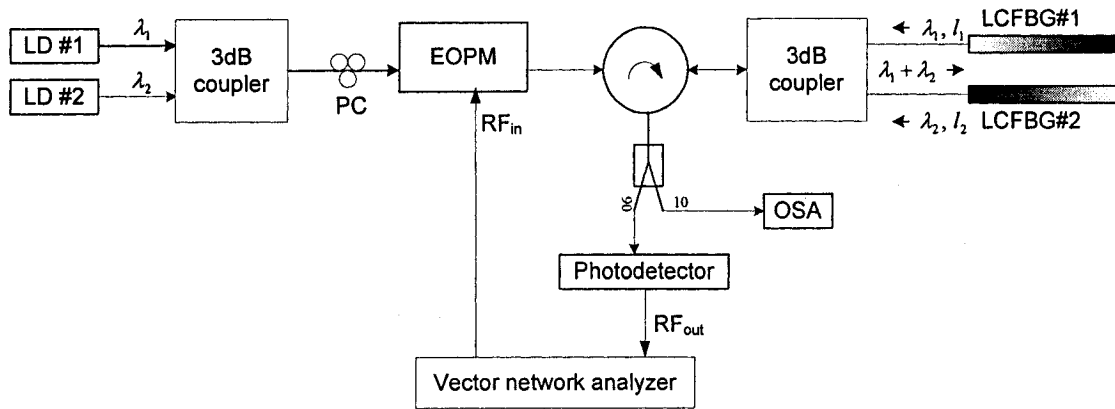


Figure 4.13 Experimental setup of a two-tap microwave filter with one negative coefficient. OSA: optical spectrum analyzer.

The detailed parameters of the components used in the experiments are listed in Table 4.2.

Table 4.2 List of components used in the experimental setup in Fig. 4.13.

LD #1	Wavelength tunable range: 1525-1625 nm Linewidth: 150 KHz
LD #2	Wavelength tunable range: 1525-1625 nm Linewidth: 150 KHz
LCFBG #1	Length: 8 cm Dispersion: 1350 ps/nm
LCFBG #2	Length: 8 cm Dispersion: -1327 ps/nm

EOPM	Working frequency: 0-10 GHz
Photodetector	Working frequency: 0-20 GHz
Vector network analyzer	Sweeping frequency range: 45 MHz-50 GHz Power: 3 dBm

Two LCFBGs used in this experiment are fabricated through one linearly chirped phase mask. By applying different tension to the fiber during the ultra-violet exposing process, a central wavelength shift of 0.7 nm is achieved. A Gaussian apodization profile is applied to flatten and smooth the reflectivity response and the group delay ripples. Both gratings have a length of 8 cm. The measured group delay and reflectivity responses for both gratings are shown in Fig. 4.14 and Fig. 4.15, respectively. LCFBG #1 is measured at the short wavelength port and LCFBG #2 is measured at the long wavelength port. The average dispersion of LCFBG #1 and LCFBG #2 are calculated to be 1350 ps/nm and -1327 ps/nm, respectively.

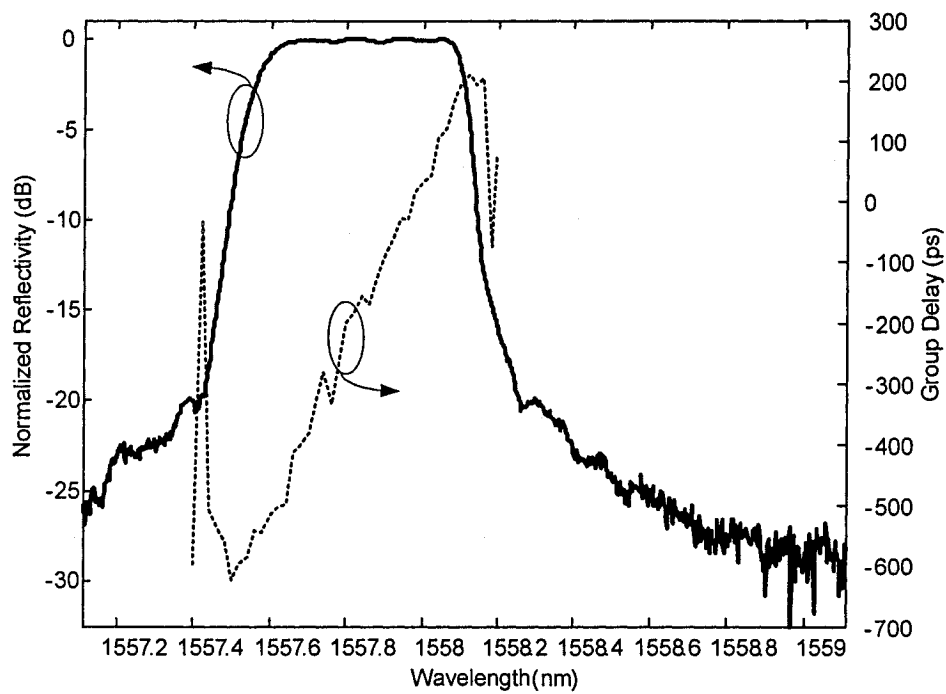


Figure 4.14 Measured reflectivity and group delay of LCFBG #1

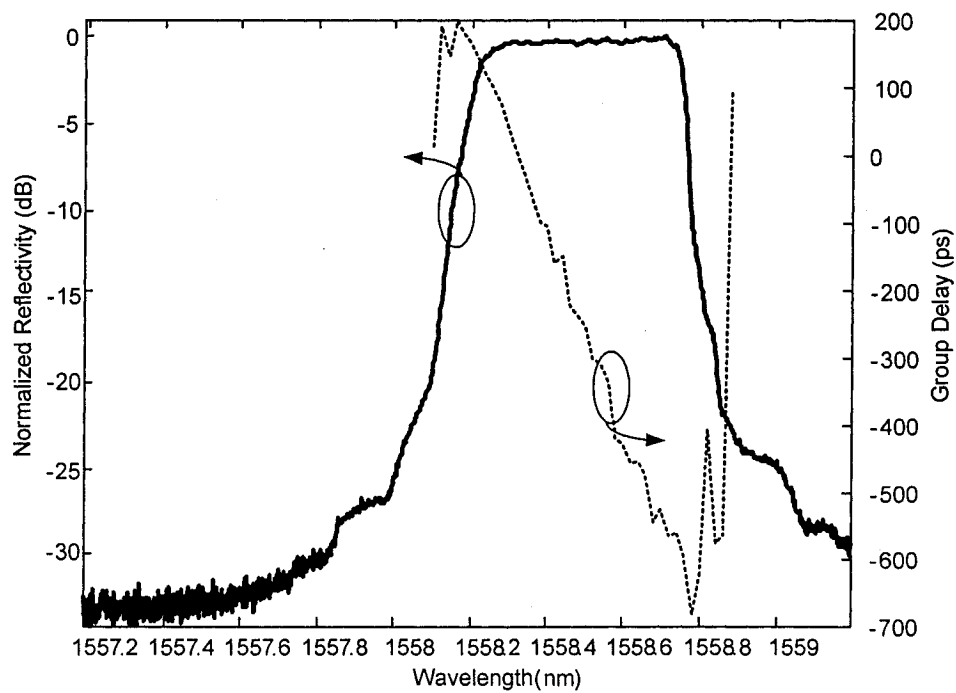


Figure 4.15 Measured reflectivity and group delay of LCFBG #2

To verify the principle of the proposed negative coefficient generation, two experiments are implemented for comparison. First, the wavelengths of the two tunable laser sources are tuned to be reflected by LCFBG #1 via the same port, as shown in Fig. 4.16(a), in which these two phase-modulated optical signals are reflected from different positions of LCFBG #1, but the experienced dispersions are identical thanks to the linearity of the group delay profile. The frequency response of the implemented filter observed from the vector network analyzer is shown in Fig. 4.16(b). The measured FSR is about 2.4 GHz, corresponding to a time interval of 417 ps. Comparing the measured frequency response with the simulated lowpass response, it is clearly seen that the baseband resonance of the lowpass filter is eliminated due to the PM-IM conversion, which is an bandpass-equivalent filter with positive-only coefficients [66].

By keeping λ_1 fixed while λ_2 is tuned to be reflected by LCFBG #2, as shown in Fig. 4.17(a), which has a reversed group delay slope with respect to that of LCFBG #1. The measured frequency response of the proposed filter is shown in Fig. 4.17(b). In this case, the FSR is 2.25 GHz, corresponding to a time interval of 444 ps. It can be observed from Fig. 4.17(b) that it is a transfer function of a bandpass filter and a negative coefficient is indeed obtained.

Comparing the frequency responses in Fig. 4.16(b) and in Fig. 4.17(b), we can see that the lowpass resonance of the bandpass-equivalent filter is only partially suppressed by the dc notch generated from the PM-IM conversion, a relatively high sidelobe at the low frequency is observed. For the frequency response in Fig. 4.17(b), since it is a true

bandpass filter with a negative tap, no lowpass resonance exists in the frequency response; a frequency response with higher MSR is obtained.

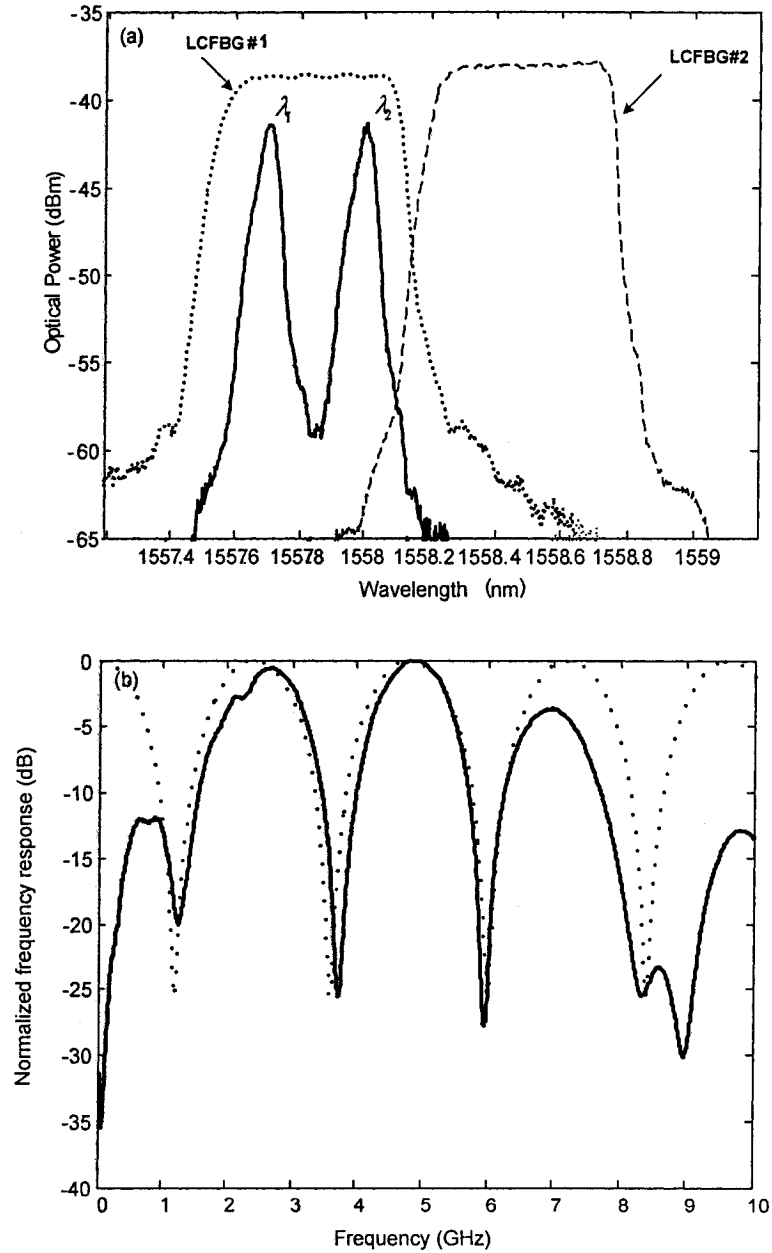


Figure 4.16 Experimental results of the implemented filter with two positive taps. (a) Measured optical spectrum (solid line) before the photodetector when both laser sources are reflected from the same port of LCFBG #1; (b) frequency responses: measured (solid line) and simulated (dotted line) which shows a lowpass filtering.

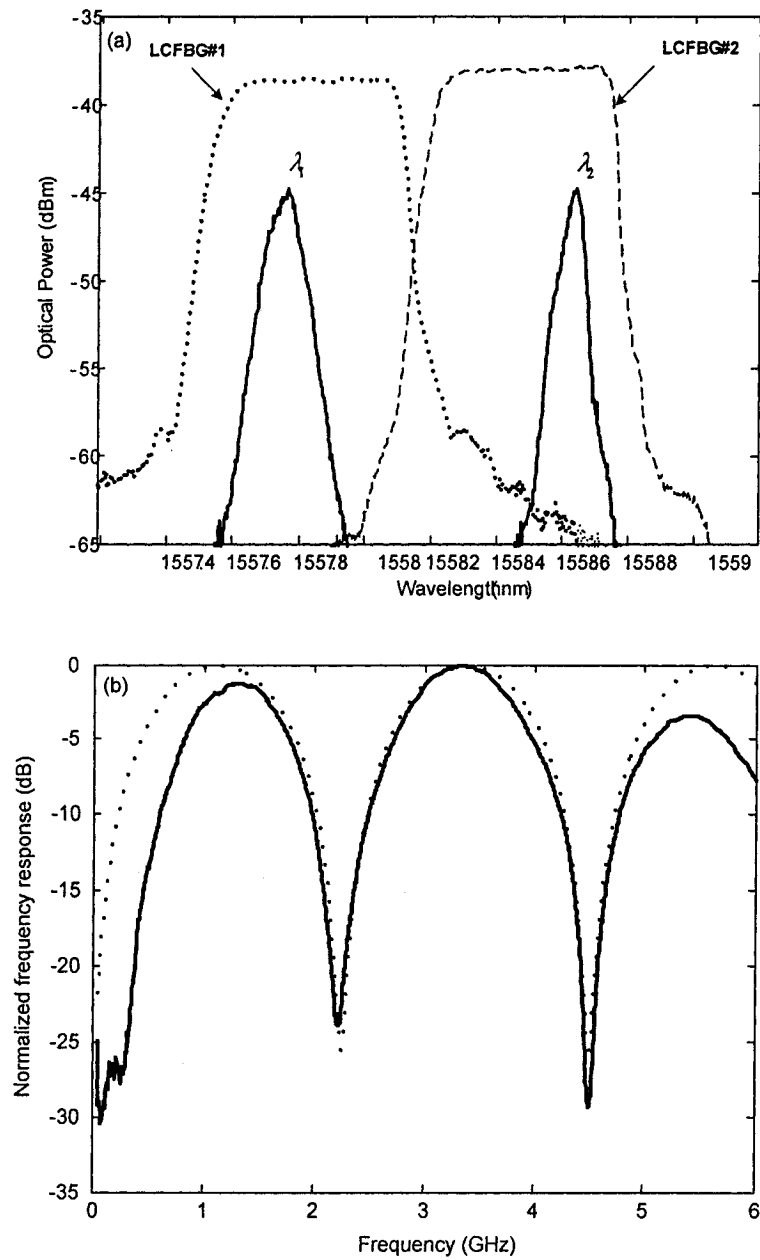


Figure 4.17 Experimental results of the two-tap filter with one negative coefficient. (a) Measured optical spectrum (solid line); (b) frequency responses: measured (solid line) and simulated (dotted line)

It should be pointed out that the LCFBGs play a key role in this system. The ripples in both the reflectivity and group delay response should be suppressed to obtain desired

frequency response and tunability of the all-optical microwave filter. Therefore, an apodization profile should be applied during the FBG fabrication.

4.4 Summary

In this Chapter, we first reviewed the major techniques reported in literature to obtain all-optical microwave filters with negative coefficients. Since FBGs are very important in all-optical microwave filters as tapping or dispersion devices, a brief review of FBGs (including uniform FBGs and CFBGs) was then given.

To achieve an all-optical microwave filter with negative coefficients, we proposed a novel and simple method to generate negative coefficients using phase modulation combined with LCFBGs. A two-tap bandpass microwave filter based on the proposed negative coefficient generation approach was demonstrated. The proposed filter has a very simple structure and high MSR. More taps with either positive or negative coefficients can be easily realized by simply adding more optical sources and CFBGs, which provides the possibility to implement microwave bandpass filters with flat-top response and high MSR. For practical applications, the proposed filter can be miniaturized by using DWDM light sources. The size can be further reduced with better performance if the LCFBGs can be integrated with the AWG on a single substrate.

Chapter 5

BIPOLAR FILTERS BASED ON PHASE MODULATION INCORPORATING AN OPTICAL FILTER

In Chapter 4, the importance of negative coefficients for all-optical microwave filters was discussed and a novel and simple approach based on an EOPM and LCFBGs to obtain bipolar microwave filters was presented. In this Chapter, a different approach to implementing all-optical microwave filters with negative coefficients will be proposed and demonstrated. In the proposed approach, the positive and negative coefficients are obtained by locating the optical carriers at the opposite slopes of the transfer function of an optical filter to convert the phase-modulated signals to intensity-modulated signals, with phase inversion of the recovered RF modulating signal. A tunable two-tap all-optical microwave filter with one negative coefficient is demonstrated.

5.1 Principle

The principle of the proposed filter with negative coefficients is shown in Fig. 5.1. Fig. 5.1(a) shows the typical intensity transmission function of an optical filter using an unbalanced MZI or a Lyot-Sagnac loop. Two optical carriers, namely carrier 1 and carrier 2, are tuned at the opposite slopes of the transfer function of the optical filter. As shown in Fig. 5.1(b), the same RF modulating signal is modulated onto both carriers via

an EOPM, which will introduce an instant frequency shift to the carriers. The value of the frequency shift is proportional to the first-order derivative of the RF modulating signal. Under this situation, the optical filter is equivalent to a frequency discriminator, by which the frequency shift is converted to the variation of optical intensity. It can be seen from Fig. 5.1, if carrier 1 and carrier 2 are phase modulated, the output optical signals from the optical filter will be with the same average power but with counter envelopes, negative coefficients are thus obtained by means of direct detection.

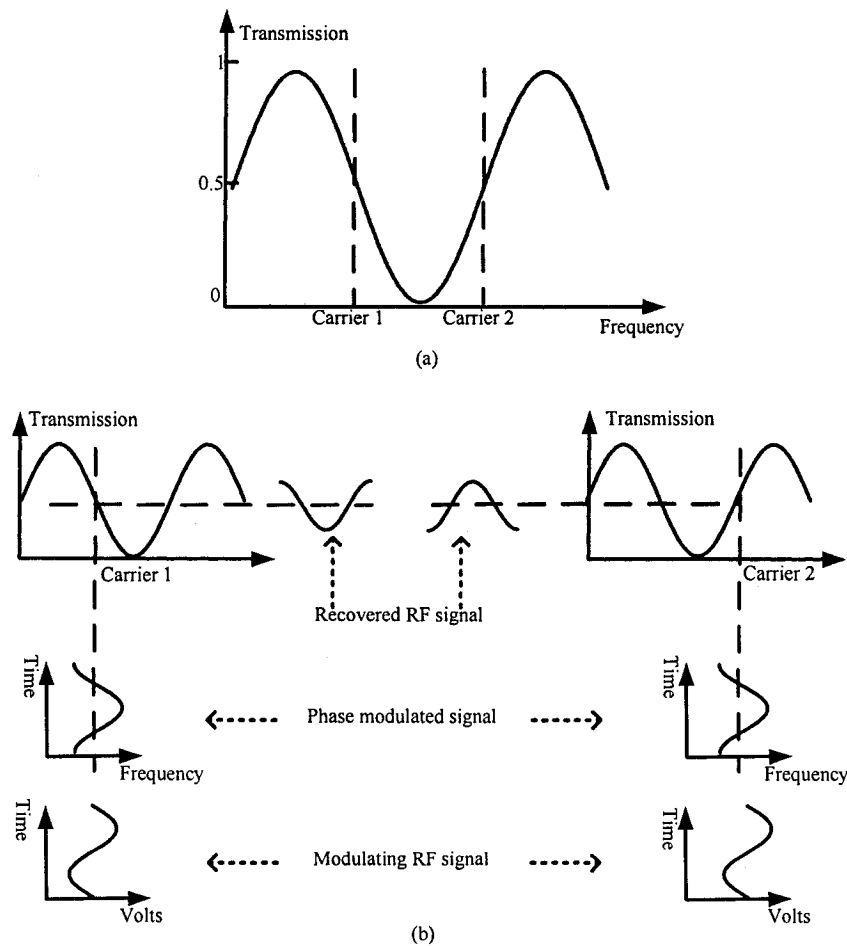


Figure 5.1 Principle of the all-optical microwave filter with negative coefficients. (a) Intensity transfer function of an optical filter. (b) Illustration of the generation of RF modulating signals in counter phase.

In theory, the phase-modulated signal can be given in the form of Eq. (5.1) when the modulation depth is small:

$$E_{PM}(t) \approx E_o \{ J_0(\beta) \cos \omega_o t + J_1(\beta) \cos[(\omega_o + \omega_e)t + \frac{\pi}{2}] - J_1(\beta) \cos[(\omega_o - \omega_e)t - \frac{\pi}{2}] \} \quad (5.1)$$

On the other hand, the transfer function of the optical filter can be described by Eq. (5.2). For simplicity, only filters based on an unbalanced MZI and a Sagnac loop with one section of Hi-Bi fiber will be considered here.

$$H(\omega) = \frac{1}{2}(1 + e^{-j\omega\tau}) = \cos \frac{\omega\tau}{2} \cdot e^{-j\frac{\omega\tau}{2}}, \quad (5.2)$$

where ω is the angular frequency of the input optical signal; τ is the time delay difference between two optical paths. Consequently, the signal after the optical filter is

$$\begin{aligned} E_{IM}(t) \approx E_o \{ & J_0(\beta) \cos \frac{\omega_o\tau}{2} \cdot \cos(\omega_o t - \frac{\omega_o\tau}{2}) \\ & + J_1(\beta) \cos \frac{(\omega_o + \omega_e)\tau}{2} \cos[(\omega_o + \omega_e)t + \frac{\pi}{2} - \frac{(\omega_o + \omega_e)\tau}{2}] \\ & - J_1(\beta) \cos \frac{(\omega_o - \omega_e)\tau}{2} \cos[(\omega_o - \omega_e)t - \frac{\pi}{2} - \frac{(\omega_o - \omega_e)\tau}{2}] \} \end{aligned} \quad (5.3)$$

If the photodetector has a responsivity $\mathcal{R}$, the recovered RF signal at the output of the photodetector is

$$\begin{aligned}
I_{RF} &= \\
&\Re \cdot E_o^2 [J_0(\beta) \cdot J_1(\beta) \cdot \cos \frac{\omega_o \tau}{2} \cdot \cos \frac{(\omega_o + \omega_e) \tau}{2} \cdot \cos(\omega_e t + \frac{\pi}{2} - \frac{\omega_e \tau}{2}) \\
&- J_0(\beta) \cdot J_1(\beta) \cdot \cos \frac{\omega_o \tau}{2} \cdot \cos \frac{(\omega_o - \omega_e) \tau}{2} \cdot \cos(\omega_e t + \frac{\pi}{2} - \frac{\omega_e \tau}{2})] \quad (5.4) \\
&= M_c \sin(\omega_o \tau) \cdot \sin \frac{\omega_e \tau}{2} \cdot \cos(\omega_e t + \frac{\pi}{2} - \frac{\omega_e \tau}{2})
\end{aligned}$$

For clarity, M_c is used to stand for the constant $-\Re \cdot E_o^2 \cdot J_0(\beta) \cdot J_1(\beta)$. From Eq. (5.4), it is easily seen that for a specific value of ω_e , the corresponding I_{RF} can have different signs by adjusting the carrier frequency ω_o to let the value of $\sin(\omega_o \tau)$ have different signs.

In our proposed approach, the ideal arrangement of a two-tap filter with one negative coefficient is based on the placement of the two optical carriers at the quadrature (3 dB) points of the opposite slopes of the transfer function. For example, let the center frequencies of carrier 1 and carrier 2, namely ω_1 and ω_2 , satisfy the equation

$$\cos^2 \frac{\omega \tau}{2} = \frac{1}{2}, \text{ which also means } \sin(\omega \tau) = \pm 1. \text{ Without loss of generality, we suppose}$$

$$\text{that } \omega_1 \tau = \frac{\pi}{2} + 2n\pi \text{ and } \omega_2 \tau = -\frac{\pi}{2} + 2n\pi, \text{ where } n = 0, \pm 1, \pm 2 \dots \text{ By using a}$$

dispersive device to induce a time delay between the two optical carriers, the overall recovered RF signal can be regarded as the summation of different taps since this is an incoherent filter, which is given by

$$I_{RF} = M \cdot \sin \frac{\omega_e \tau}{2} [-P_1 \cdot \cos(\omega_e t + \frac{\pi}{2} - \frac{\omega_e \tau}{2}) + P_2 \cdot \cos(\omega_e(t - T) + \frac{\pi}{2} - \frac{\omega_e \tau}{2})], \quad (5.5)$$

where P_1 and P_2 are the optical powers of optical carrier 1 and optical carrier 2, respectively, T is the dispersive device induced time delay between the two taps, which can be expressed as the product of the accumulated dispersion χ and the wavelength spacing $\Delta\lambda$ between the two carriers. To obtain the highest notch rejection level, P_1 and P_2 should be identical. In this situation, the normalized transfer function of the optical microwave filter is expressed by

$$|H(\omega)| = \left| \sin \frac{\omega\tau}{2} \cdot (1 - e^{-j\omega T}) \right|. \quad (5.6)$$

The term $\sin \frac{\omega\tau}{2}$ in Eq. (5.6) is determined by the frequency response of the PM-IM conversion; the term $(1 - e^{-j\omega T})$ is induced by the summation of the recovered RF signals from the two taps. On the contrary, if the two optical carriers are located at the quadrature frequencies with equal slopes, the normalized transfer function of the optical microwave filter is given

$$|H(\omega)| = \left| \sin \frac{\omega\tau}{2} \cdot (1 + e^{-j\omega T}) \right|. \quad (5.7)$$

It is worth pointing out that although in Eq. (5.7) the filter has only positive coefficients, the baseband resonance is still partially suppressed by the dc notch caused by the PM-IM conversion.

5.2 Experimental results

Based on the proposed approach, a two-tap all-optical microwave filter with one negative coefficient, shown in Fig. 5.2, is set up. Two tunable laser sources are used to generate the two optical carriers. The two carriers are phase modulated by an RF signal generated by a vector network analyzer and then fed into a Sagnac loop serving as an optical filter. The intensity transfer function of the Sagnac loop is shown in Fig. 5.3. It can be seen that the optical filter has a free spectral range FSR_λ of around 0.19 nm. A time delay difference is obtained by passing the filtered optical carriers through a 25-km single-mode fiber, which has an accumulated dispersion of 425 ps/nm at 1550 nm. Considering the loss induced by the first coupler (about 3dB), the EOPM (about 8dB), the optical filter (about 3dB), the 25 km fiber (about 5dB) and other components, such as PC and fiber optic connectors, an EDFA is applied in this configuration to compensate the attenuation.

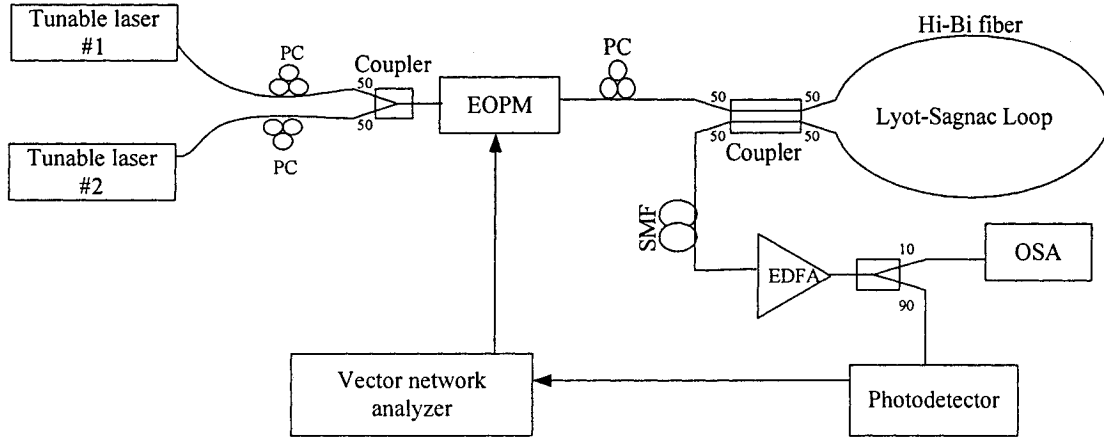


Figure 5.2 Experimental setup.

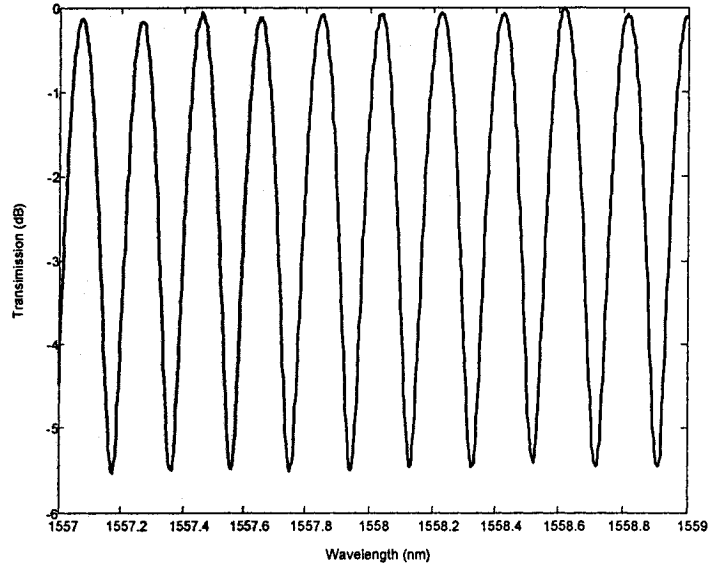


Figure 5.3 Intensity transfer function of the Sagnac-loop optical filter.

The detailed parameters of the main components used in the experiments are listed in Table 5.1.

Based on the theoretical analysis, it is expected that if the wavelength of the first tunable laser source, namely λ_1 , is fixed at the quadrature point of one positive slope of the transfer function of the optical filter, different microwave transfer functions will be achieved when the wavelength of the second tunable laser source, namely λ_2 , is tuned to satisfy the conditions $|\lambda_1 - \lambda_2| = (2n + 1) \frac{FSR_\lambda}{2}$ or $|\lambda_1 - \lambda_2| = 2n \cdot \frac{FSR_\lambda}{2}$, where $n = 0, 1, 2, 3 \dots$. In the first case, a true bandpass frequency response is expected since λ_1 and λ_2 are located at the opposite slopes, and bipolar coefficients should be obtained; in the second case, a bandpass equivalent filter will be obtained because λ_1 and λ_2 are at the

equal slopes, and the coefficients will have the same polarity. In the following experiments, the two different cases are experimented.

Table 5.1 List of components used in the experimental setup in Fig. 5.2.

Tunable laser #1	Wavelength tunable range: 1525-1625 nm Linewidth: 150 KHz
Tunable laser #2	Wavelength tunable range: 1525-1625 nm Linewidth: 150 KHz
EOPM	Working frequency: 0-10 GHz
Photodetector	Working frequency: 0-20 GHz
Vector network analyzer	Sweeping frequency range: 45 MHz-50 GHz Power: 3 dBm

First, λ_1 is fixed at 1557.282 nm, and λ_2 is tuned to 1558.246 nm, as shown in Fig. 5.4(a). The spacing between λ_1 and λ_2 is 0.964 nm, which is 5 times of FSR_λ . Since the wavelengths of the two carriers are located at the points with equal slopes, no negative coefficients are obtained. But, as discussed earlier, the baseband resonance is suppressed by the PM-IM conversion, a bandpass-equivalent filter is obtained. The filter frequency response is shown in Fig. 5.4(b). As can be seen, a high sidelobe at the baseband is observed. The sidelobe is caused by the baseband resonance, which is only

partially suppressed by the PM-IM conversion. The FSR of the microwave filter is 2.4 GHz, corresponding to a time delay of 410 ps.

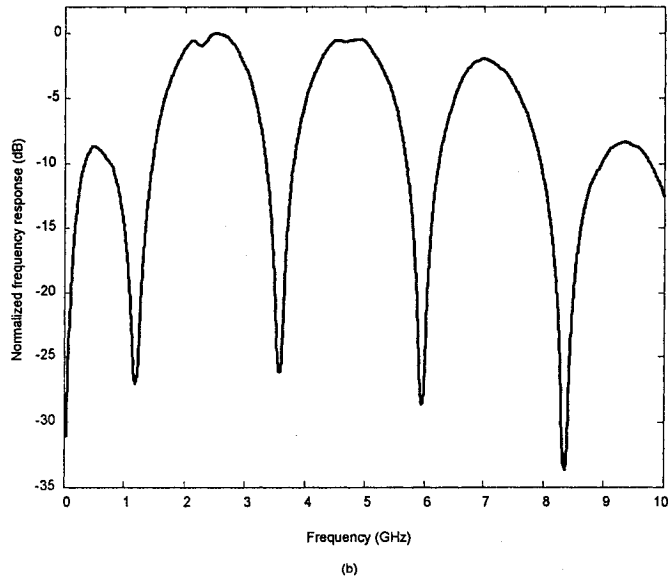
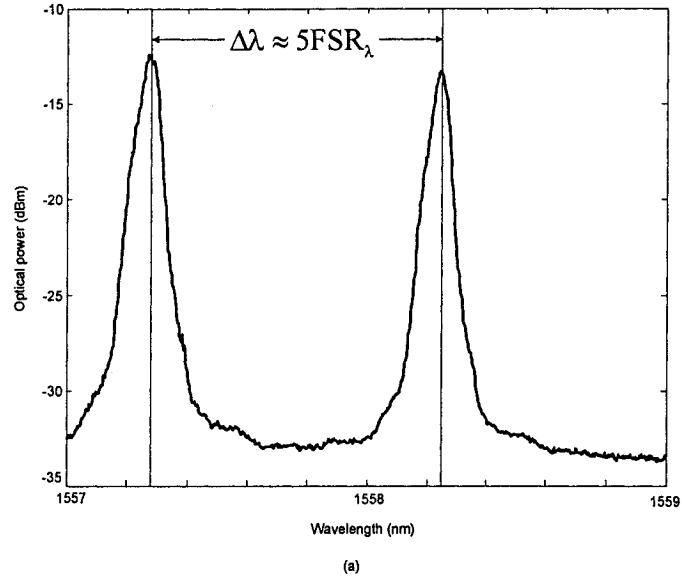
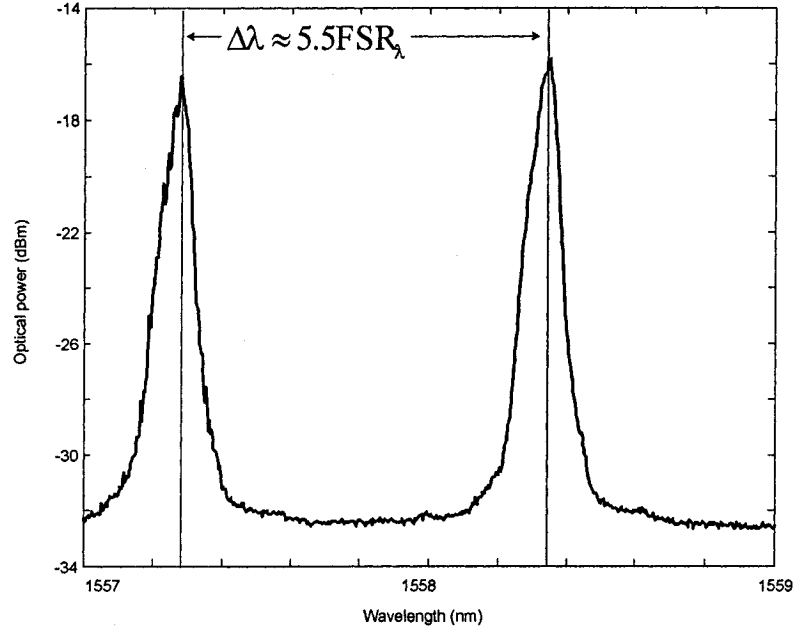


Figure 5.4 Bandpass-equivalent filter with only positive coefficients. (a) Optical spectrum of the two carrier optical source generated by two tunable laser sources; (b) Frequency response of the bandpass-equivalent filter.

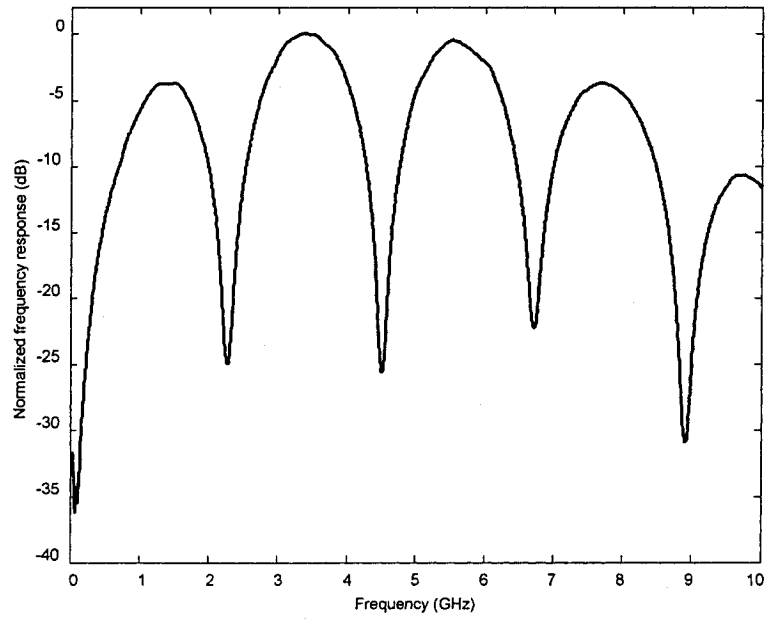
Then, λ_2 is tuned to a wavelength of 1558.336 nm, as shown in Fig. 5.5(a). The spacing between λ_1 and λ_2 is now 1.054 nm, 5.5 times of FSR_λ . Since the wavelengths of the two carriers are now located at the points with opposite slopes, a negative coefficient is obtained. The frequency response of the microwave filter is shown in Fig. 5.5(b). It is a true bandpass filter with a negative coefficient. No sidelobe is observed at the baseband. The FSR of the filter is 2.2 GHz, which corresponds to a time delay of 448 ps.

The tunability of the proposed microwave filter is also investigated. When λ_2 is tuned to 1557.964 nm, the spacing between λ_1 and λ_2 is 0.682 nm, 3.5 times of FSR_λ , as shown in Fig. 5.6(a). In this case, a negative coefficient is still obtained, but with a smaller time delay difference, the FSR is thus increased, as shown in Fig. 5.6(b). In the experiment, the FSR of the microwave filter is 3.3 GHz, which corresponds to a time delay of 290 ps.

In Figs. 5.4 to 5.6, the degradation of the magnitude response shown in higher frequencies is mainly due to the power penalty induced by the chromatic dispersion of the 25-km single-mode fiber.

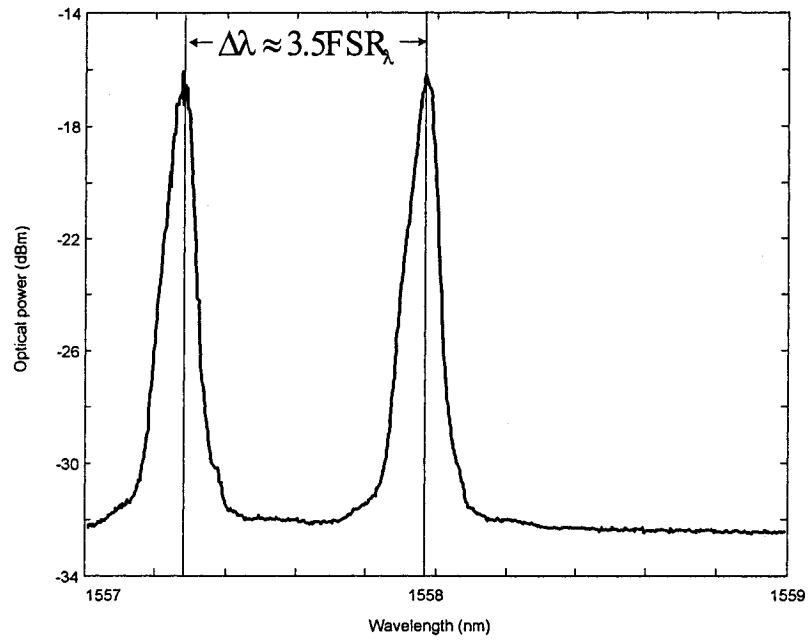


(a)

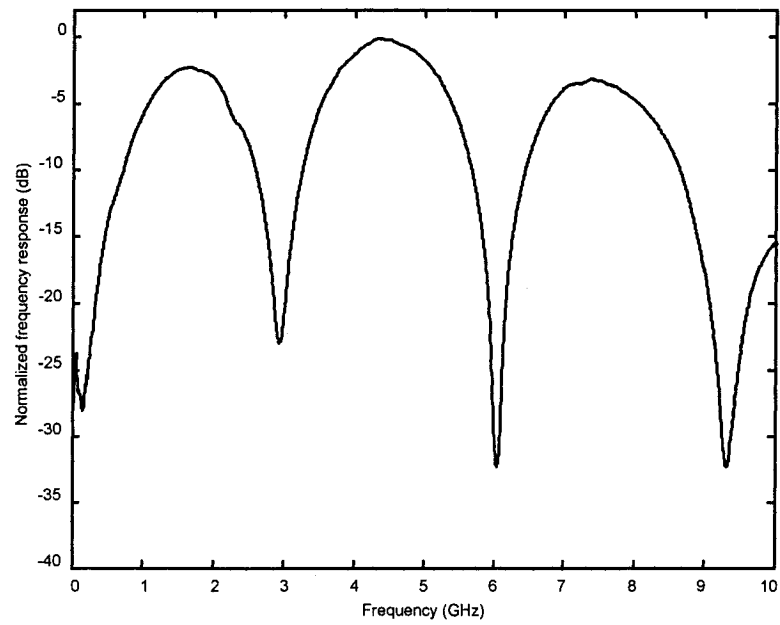


(b)

Figure 5.5 True bandpass filter with a negative coefficient. (a) Optical spectrum of the two carrier source generated from the two tunable laser sources. (b) Frequency response of the true bandpass filter.



(a)



(b)

Figure 5.6 Tunability of the proposed bandpass filter. (a) Optical spectrum of the two carrier source generated from the two tunable lasers. (b) Frequency response of the all-optical bandpass microwave filter.

5.3 Further discussions

We should note that the performance of the proposed all-optical microwave filter, especially the notch rejection level, depends highly on the stability of the optical sources and the optical filter. The use of the state-of-the-art optical sources will solve the laser stability problem. For example, the wavelength drift of JDS-Uniphase laser diodes with case temperature is much better than $1 \text{ pm}/^{\circ}\text{C}$. In our experiment, the performance of the proposed microwave filter is mainly affected by the instability of the optical filter. We believe that this problem can be solved by using an optical filter with proper packaging and temperature control or by using a temperature-insensitive Sagnac loop.

In the proposed configuration, a two-tap all-optical microwave bandpass filter is implemented. For microwave filter with multiple taps, modified configurations may be proposed to realize evenly time distributed taps with arbitrary positive or negative coefficients. A configuration that meets this requirement is shown in Fig. 5.7. In the configuration, optical delay lines are employed to introduce time delays. Two AWGs are used to de-multiplex and multiplex the multiple carriers. Positive or negative coefficients can be arbitrarily determined by selecting the wavelengths of the optical sources.

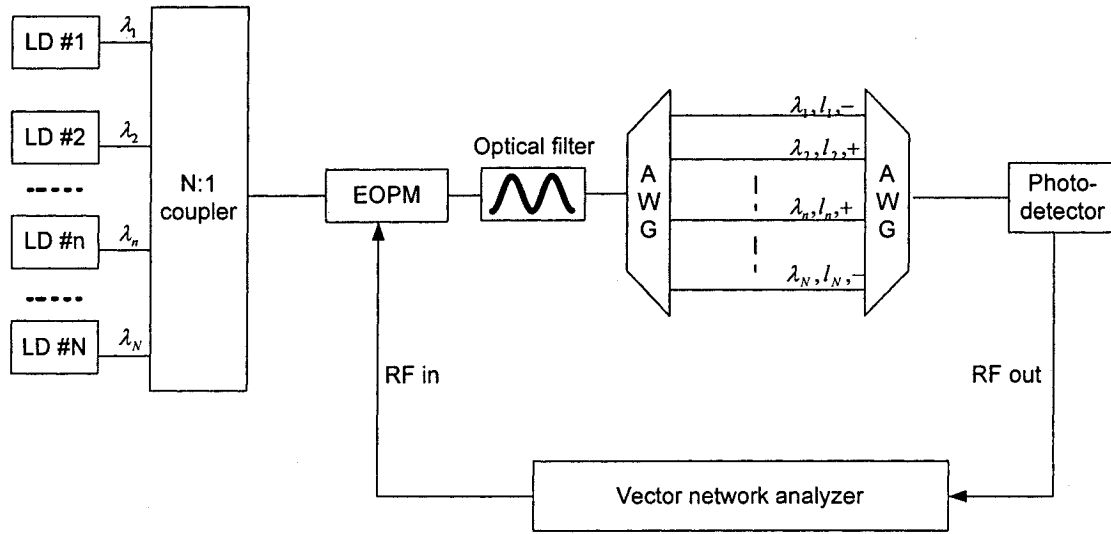


Figure 5.7 A configuration of multi-tap bandpass filters.

Another important issue that should be carefully addressed is that in our experiment, the RF signal is phase modulated on the optical carrier; however, the PM-IM conversion is completed through frequency discrimination. It means that the recovered RF signal is the first-order derivative of the modulating RF signal. For a sinusoidal signal, it is not a serious problem, since the first-order derivative of a sinusoidal signal is still sinusoidal, with $\pi/2$ phase difference. However, if the modulating signal is not a sinusoidal signal, for example, if it is a binary phase shift keying (BPSK) signal, the recovered digital signal may be distorted. In the following section, we will investigate and prove that the digital signal can be recovered at the output of the all-optical microwave filter without distortion.

In general, a BPSK signal can be described by

$$s(t) = -V_e m(t) \sin \omega_e t, \quad (5.8)$$

where $m(t)$ is a polar baseband data signal, and $V_e \cos(\omega_e t)$ is the subcarrier with amplitude V_e and angular frequency ω_e . Assuming the optical carrier is at the frequency ω_o with amplitude E_o , the phase modulated signal can be written as

$$E_{PM} = E_o \cos[\omega_o t + D_p s(t)], \quad (5.9)$$

where D_p is the phase modulation index. Therefore, the instant frequency is

$$\omega_i = \frac{d[\omega_o t + D_p s(t)]}{dt} = \omega_o + D_p \frac{ds(t)}{dt}. \quad (5.10)$$

Substituting Eq. (5.8) into Eq. (5.10), we have

$$\omega_i = \omega_o - D_p \cdot V_e [m(t) \cdot \omega_c \cdot \cos \omega_c t + \frac{dm(t)}{dt} \sin \omega_c t]. \quad (5.11)$$

If we assume that the frequency discriminator has a response linearly proportional to the instant frequency shift, the envelope of the optical carrier after PM-IM conversion can be expressed by

$$E_e \propto m(t) \cdot \omega_c \cdot \cos \omega_c t + \frac{dm(t)}{dt} \sin \omega_c t. \quad (5.12)$$

Since the photodetector is operating as an envelope detector, the AC components after the photodetector have the same elements as Eq. (5.12). As a result, it is possible to extract the baseband data signal if a coherent detection is employed, i.e., mixing the signal in Eq. (5.12) with a local reference signal $\cos \omega_c t$, the resulting waveform is

$$\begin{aligned}
s_r(t) &\propto m(t) \cdot \omega_c \cdot \cos^2 \omega_c t + \frac{dm(t)}{dt} \sin \omega_c t \cdot \cos \omega_c t \\
&= \frac{1}{2} m(t) \cdot \omega_c + \frac{1}{2} m(t) \cdot \omega_c \cdot \cos 2\omega_c t + \frac{1}{2} \frac{dm(t)}{dt} \cdot \sin 2\omega_c t
\end{aligned} \tag{5.13}$$

As can be seen, using a lowpass filter, the baseband data signal $m(t)$ can be recovered.

Fig. 5.8 shows the waveforms of the baseband, subcarrier and BPSK signals, where the baseband signal $m(t)$ is at 10 Mb/s, the subcarrier frequency is 1 GHz. When the BPSK signal is phase-modulated onto an optical carrier, the normalized instant phase shift and the corresponding normalized signal amplitude after the optical filter are shown in Fig. 5.9. Note that the optical filter is taken as a linear frequency discriminator and only the envelope of the optical carrier is taken into account. For clarity, Fig. 5.9 shows the situations at instances 0.3 ms and 0.4 ms, from which we can see that the instant frequency shift is the first-order derivative of the BPSK signal. Using coherent detection, we can get the signal described by Eq. (5.13). From the upper part of Fig. 5.10, it can be observed that the envelope is exactly the same with the baseband signal. Using a lowpass filter to remove the high frequency elements, the baseband signal can be extracted, as shown in the lower part of Fig. 5.10.

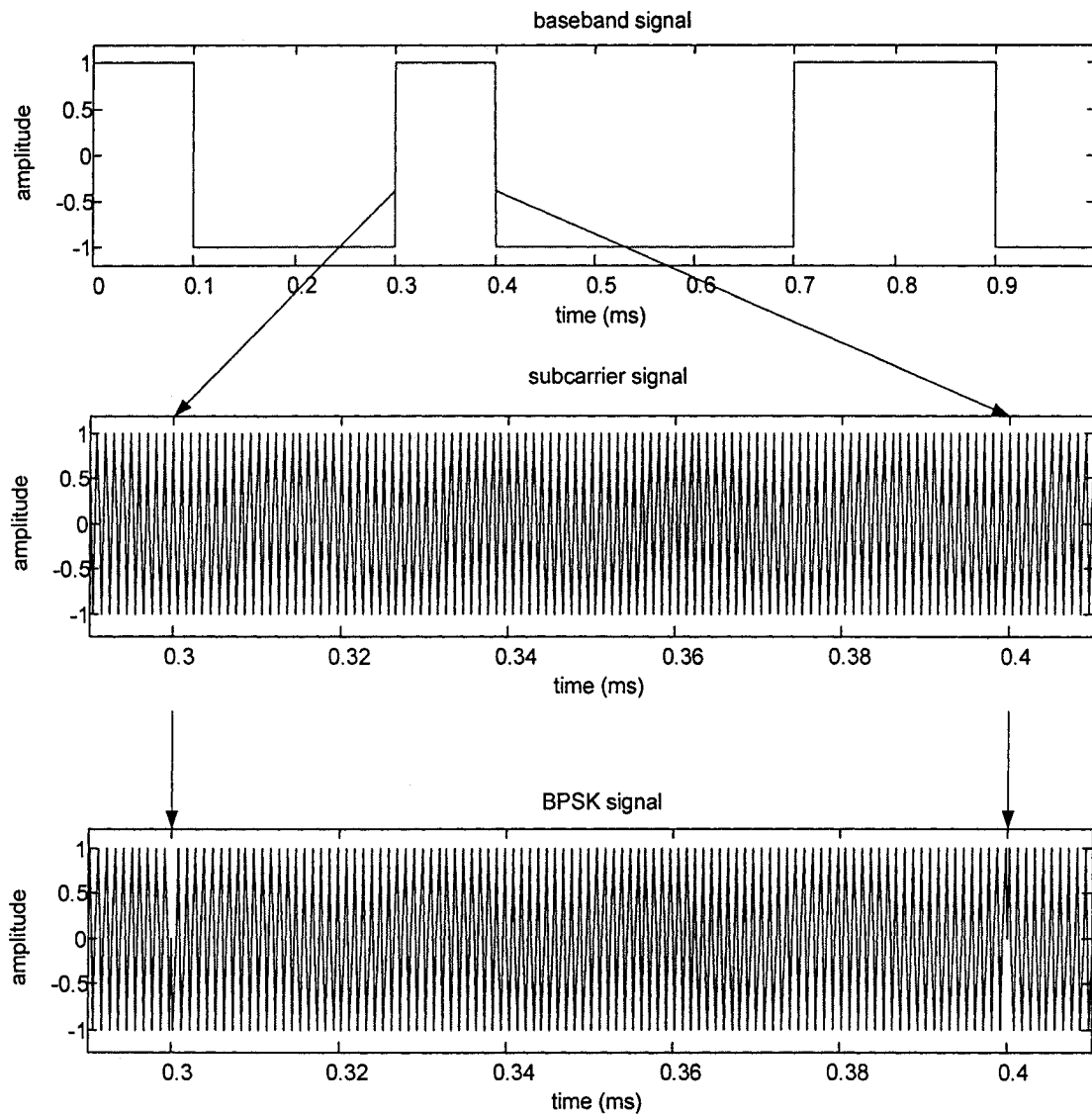


Figure 5.8 The baseband, subcarrier and BPSK signals.

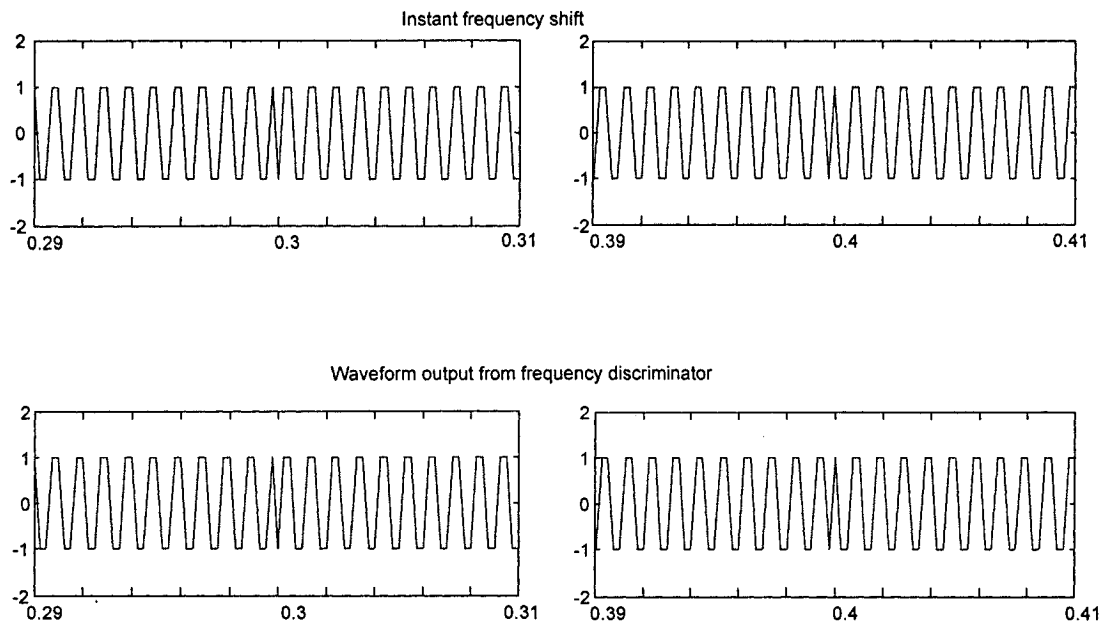


Figure 5.9 The normalized instant frequency shift of the optical carrier and the envelope of the optical carrier after the optical filter.

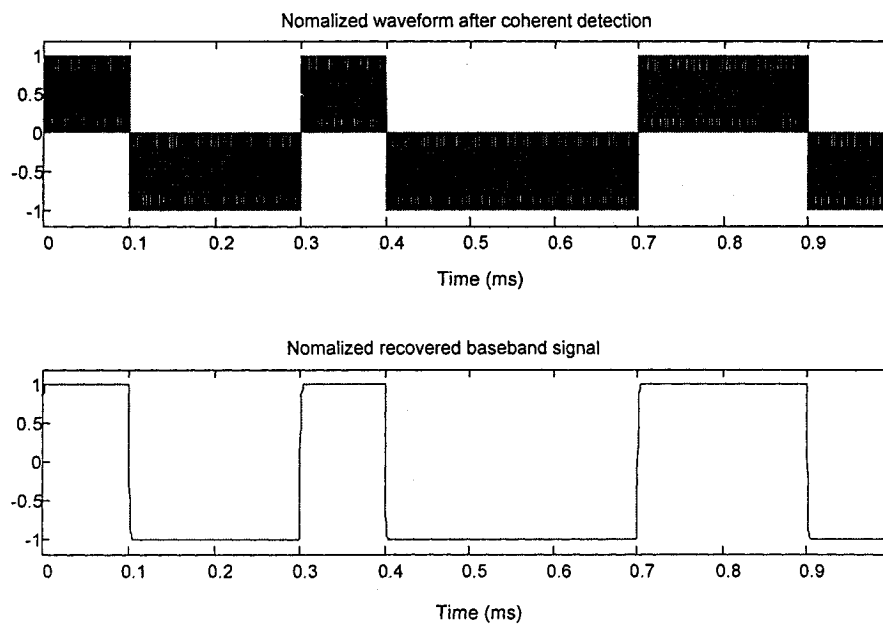


Figure 5.10 Normalized coherent-detected signal and recovered baseband signal.

5.4 Summary

In this Chapter, a novel method for obtaining negative coefficients through PM-IM conversion by using an EOPM and an optical filter was proposed. A two-tap bandpass filter with one negative coefficient based on the proposed approach was demonstrated. The tunability of the proposed filter was also investigated. The proposed approach has the potential to implement all-optical multitap microwave filters with arbitrary positive and negative coefficients. In addition, the capability of the proposed filter to recover the original baseband signal after sustaining phase modulation and frequency discrimination has also been investigated. It has been proved in theory that the original data information could be recovered at the output of the filter. This conclusion was also verified by simulations.

Chapter 6

SUMMARY AND FUTURE WORK

6.1 Summary

In this thesis, three new approaches to implementing all-optical microwave bandpass filters were proposed. All three approaches were based on phase modulation. In the first approach, to achieve different time delays, a Hi-Bi fiber was used. Since the two polarization modes were orthogonal, the filter was immune to coherence interference. The PM-IM conversion was achieved using a 25-km single-mode fiber. In the second approach, the PM-IM conversion was achieved by using LCFBGs. Different time delays were obtained by employing fiber delay lines with different lengths and reflecting the optical carriers at different locations of the LCFBGs. In the third approach, the PM-IM conversion was realized by locating the optical carriers at the positive or negative slopes of the optical filter. Different time delays were obtained by passing the optical carriers through 25-km single-mode fiber, serving as a dispersive device.

The objectives of the work have been achieved: (1) to implement an all-optical microwave bandpass filter using a laser source with strong coherence, but immune to coherent interference. Such a filter could be directly incorporated into a radio-over-fiber system to process the microwave signal without extra O/E and E/O conversions; (2) to

explore techniques for negative coefficient generation, which is the key to obtain all-optical microwave filters with improved filtering functionalities.

In Chapter 2, the key components including EOIMs, EOPMs and photodetectors were reviewed. Then, a comparison between phase modulation and intensity modulation was made. A study on the general structure and transfer function for all-optical microwave filters based on either intensity modulation or phase modulation was carried out.

In Chapter 3, an all-optical bandpass microwave filter using a laser source with narrow linewidth was demonstrated. The coherence limitation problem was solved by use of a length of Hi-Bi fiber in which the two orthogonal polarization modes would not interfere. By use of the EOPM in combination with the dispersive device to eliminate the baseband resonance, a bandpass-equivalent filter suitable for deployment in a radio-over-fiber link was obtained. Both the theoretical analysis and the experimental results were presented in this chapter.

In Chapter 4, after a review of the existing techniques to realize bipolar microwave filters, a novel and simple approach to implementing an all-optical microwave filter with negative coefficients were proposed. In the proposed approach, an EOPM in combination with LCFBGs was used to realize the PM-IM conversion. Bipolar coefficients were obtained when the optical carriers were reflected by the LCFBGs with positive and negative dispersions. A two-tap all-optical microwave bandpass filter with one negative coefficient was experimentally demonstrated.

In Chapter 5, we proposed an approach to implementing bipolar microwave filters using an EOPM in combination with an optical filter. The PM-IM conversion was realized by passing the optical carriers through the optical filter. Bipolar coefficients were obtained by locating the optical carriers at the positive or negative slopes of the optical filter. A two-tap all-optical microwave bandpass filter with one negative coefficient was experimentally demonstrated. In addition, the capability to recover the original baseband signal after sustaining phase modulation and frequency discrimination has also been investigated. It has been proved in theory that the original data information could be recovered at the output of the filter, which was also verified by simulations.

6.2 Future work

Because all the work presented in this thesis is at the proof-of-concept stage, further investigation to implement the proposed filters to practical applications would be required.

In this thesis, a two-tap bandpass-equivalent filter was proposed with a target to be directly deployed into a radio-over-fiber link. This filter employed a section of Hi-Bi fiber as the optical tapping and time delay device. The different time delays were obtained when the orthogonal polarization modes traveling in the Hi-Bi fiber along the fast and slow axes with different refractive indices. However, when the filter is incorporated into a radio-over-fiber link, the PMD of the fiber link will introduce extra time delays, which may deteriorate the filter response. In addition, mode coupling caused by micro-bending and other factors, and, phase noise induced by the system may

also change the frequency response of the all-optical microwave filter. These impacts were not investigated in this thesis. Furthermore, for practical applications, a filter with multiple taps is necessary. In our experiment, optical interference was observed in the four-tap filter. The impact of the optical interference on filters with multiple taps should be further investigated both theoretically and experimentally.

Two techniques to implement bipolar microwave filters were proposed and demonstrated in this thesis. However, the experimental verifications were performed based on two-tap filters. Further work is needed to implement filters with multiple taps to get high Q and flat -top frequency responses.

In addition, for the three proposed filter architectures, the transfer functions are all equal to the product of the transfer function of PM-IM conversion and the frequency response of a conventional lowpass or bandpass filter, which implies that the operating bandwidth of the filter may be limited by the PM-IM conversion. To overcome this limitation, we may explore the possibility of using single sideband phase modulation, by which a bipolar system is easy to obtain by suppressing either the +1 order or -1 order sideband. One possible method worth attempting is to use an FBG to select the carrier and one sideband to realize single sideband phase modulation.

Finally, for a radio-over-fiber system, further research would be carried out to evaluate the system performance when using one of the proposed all-optical microwave filters to reject electrical noise and microwave interference in a practical radio-over-fiber system.

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