

Photonic Dispersive Delay Line for Broadband Microwave Signal Processing

Jiejun Zhang

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School of Electrical Engineering and Computer Science

Faculty of Engineering

University of Ottawa



uOttawa

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ABSTRACT

The development of communications technologies has led to an ever-increasing requirement for a wider bandwidth of microwave signal processing systems. To overcome the inherent electronic speed limitations, photonic techniques have been developed for the processing of ultra-broadband microwave signals. A dispersive delay line (DDL) is able to introduce different time delays to different spectral components, which are used to implement signal processing functions, such as time reversal, time delay, dispersion compensation, Fourier transformation and pulse compression. An electrical DDL is usually implemented based on a surface acoustic wave (SAW) device or a synthesized C-sections microwave transmission line, with a bandwidth limited to a few GHz. However, an optical DDL can have a much wider bandwidth up to several THz. Hence, an optical DDL can be used for the processing of an ultra-broadband microwave signal. In this thesis, we will focus on using a DDL based on a linearly chirped fiber Bragg grating (LCFBG) for the processing of broadband microwave signals. Several signal processing functions are investigated in this thesis. 1) A broadband and precise microwave time reversal system using an LCFBG-based DDL is investigated. By working in conjunction with a polarization beam splitter, a wideband microwave waveform modulated on an optical pulse can be temporally reversed after the optical pulse is reflected by the LCFBG for three times thanks to the opposite dispersion coefficient of the LCFBG when the optical pulse is reflected from the opposite ends. A theoretical bandwidth as large as 273 GHz can be achieved for the time reversal. 2) Based on the microwave time reversal using an LCFBG-based DDL, a microwave photonic matched filter is implemented for simultaneously

generating and compressing an arbitrary microwave waveform. A temporal convolution system for the calculation of real time convolution of two wideband microwave signals is demonstrated for the first time. 3) The dispersion of an LCFBG is determined by its physical length. To have a large dispersion coefficient while maintaining a short physical length, we can use an optical recirculating loop incorporating an LCFBG. By allowing a microwave waveform to travel in the recirculating loop multiple times, the microwave waveform will be dispersed by the LCFBG multiple times, and the equivalent dispersion will be multiple times as large as that of a single LCFBG. Based on this concept, a time-stretch microwave sampling system with a record stretching factor of 32 is developed. Thanks to the ultra-large dispersion, the system can be used for single-shot sampling of a signal with a bandwidth up to a THz. The study in using the recirculating loop for the stretching of a microwave waveform with a large stretching factor is also performed. 4) Based on the dispersive loop with an extremely large dispersion, a photonic microwave arbitrary waveform generation system is demonstrated with an increased the time-bandwidth product (TBWP). The dispersive loop is also used to achieve tunable time delays by controlling the number of round trips for the implementation of a photonic true time delay beamforming system.

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LIST OF ACRONYMS

ASE	Amplified spontaneous emission
ADC	Analog-to-digital conversion
AWG	Arbitrary waveform generation
CMOS	Complementary metal oxide silicon
CW	Continuous wavelength
DAC	Digital-to-analog conversion
DCF	Dispersion compensating fiber
DD	Direct detection
DDL	Dispersive delay line
DSP	Digital signal processing
EA	Electrical amplifier
EBG	Electromagnetic bandgap
EDFA	Erbium doped fiber amplifier
FSR	Free spectral range
FBG	Fiber Bragg grating
GDD	Group delay dispersion
FP	Fabry-Perot
FPI	Fabry-Perot interferometer
IC	Integrated circuit
IDT	Interdigital transducer
IM	Intensity modulator

LCFBG	Linearly chirped fiber Bragg grating
LCMW	Linearly chirped microwave waveform
LD	Laser diode
LTI	Linear time-invariant
MLL	Mode-locked laser
MPF	Microwave photonic filter
MZM	Mach-Zehnder modulator
MZI	Mach-Zehnder interferometer
OBPF	Optical bandpass filter
OC	Optical circulator
OEO	Opto-electronic oscillator
OF	Optical filter
OVA	Optical vector analyzer
PAA	Phased array antenna
PBC	Polarization beam combiner
PBS	Polarization beam splitter
PC	Polarization controller
PCMW	Phase-coded microwave waveform
PD	Photodetector
POF	Programmable optical filter
SAW	Surface acoustic wave
SBS	Stimulated Brillouin scattering
SG	Signal generator

SMF	Single mode fiber
SNR	Signal-to-noise ratio
SS	Spectral-shaping
SSB	Single-sideband
TBWP	Time-bandwidth product
TFBG	Tilted fiber Bragg grating
TLS	Tunable laser source
TPS	Temporal pulse shaping
UV	Ultraviolet
WDM	Wavelength division multiplexer
WS	Waveshaper
WTT	Wavelength-to-time

CHAPTER 1 INTRODUCTION

1.1 Background Review

A delay line is a device that introduce a time delay to a signal that travels through it. A delay line is one of the fundamental building blocks for signal processing and has found broad applications. For example, in [1], a network consisting of an array of delay lines with tunable progressive time delays was used for beam steering of a phased array antenna (PAA); in [2, 3], a dispersive delay line that introduces different time delays to different frequency components of an input signal was implemented for dispersion compensation in a communications system; in [4], a delay line was used as a memory unit for a computation system. In general, a delay line can be implemented in the electrical domain or in the optical domain. In this thesis, we focus on the investigation of delay lines implemented in the optical domain with large time delay tunability and wide bandwidth for broadband microwave signal processing.

A delay line can be implemented in the electrical domain and optical domain. So far, digital signal processors are most widely used to generate a time delay to a microwave signal and to perform a variety of signal processing functionalities. In a digital signal processor, a microwave is sampled and stored digitally. After a certain time delay, a digital-to-analog conversion module will be used to reconstruct the original signal. Such a digital system can generate an arbitrarily long time delay to a microwave signal. However, its operation bandwidth is strictly limited to less than a few tens of GHz due to the limited speed of existing electronic systems. In addition, such system is complex and has a high cost.

An analog electrical delay line can generate a time delay from a few nanoseconds to microseconds at a much lower cost. Different approaches have been proposed to realize an analog electrical delay line, such as a long electrical cable, a surface acoustic wave (SAW) device [5-10], an electromagnetic bandgap (EBG) element [11-13], and an integrated circuit (IC) [14-16]. For example, a SAW delay line, implemented using two interdigital transducers (IDTs) on a piezoelectric substrate with a certain spacing, can generate a time delay up to hundreds of nanoseconds. An IDT is a device that consists of two interlocking comb-shaped arrays which are metallic electrodes [6]. An electrical signal is converted to SAW at the transmitting IDT, propagates along the surface of the piezoelectric substrate and is converted back to the electrical domain by the receiving IDT. Thanks to the low group velocity of the SAW compared to that of a microwave signal in an electrical wire, an SAW device can achieve a large time delay with a relatively small foot print. In [7], an SAW delay line with a time delay of 750 ns has been achieved at a central frequency of 280 MHz and a bandwidth of 190 MHz. To achieve a microwave frequency-dependent time delay, an SAW device can be implemented using a chirped reflector or complementarily chirped IDTs [8]. A linear group delay response of 0.4 $\mu\text{s}/\text{MHz}$ was achieved. For wideband microwave communication and sensing applications, SAW delay lines are required to operate at GHz range. To achieve this, SAW elements are integrated based on the complementary metal oxide silicon (CMOS) platform due to the high photographic resolution. In [9], IDTs embedded in silicon oxide layer that is coated with a piezoelectric film is used to achieve a SAW delay with an operation bandwidth of 4 GHz. In [10], two IDTs fabricated on a piezoelectric layer sandwiched between two silicon oxide layers have achieved a SAW delay line with a bandwidth of 23.5 GHz. An EBG element is another device that can be used to effectively achieve a wideband electrical delay line, but with a smaller time delay. An

EBG element has a periodic structure created by periodically modulating the transmission line impedance, such as a one-dimensional (1-D) transmission line. When the wavelength of the input signal satisfies the Bragg condition of the periodic structure, the signal will be reflected, resulting in a time delay determined by the location of the reflection [11]. Different types of EBG can be fabricated to achieve a desired spectral response. A uniform EBG, which has a uniform impedance modulation strength and period, is used as a reflector for a narrowband signal. In an apodized EBG element, the impedance modulation period is constant, but the modulation strength is changing according to a certain profile to optimize the frequency response of the device, for example, to achieve a flattop response or to suppress the sidelobes. In a chirped EBG, the period of the impedance modulation is usually linearly changing, so that different spectral components of the input signal will be reflected at different locations in an EBG element, resulting a chirped time delay and a large reflection bandwidth [12, 13]. In [13], an EBG strip waveguide with a length 6.8 cm was demonstrated with a dispersion coefficient of 0.15 ns/GHz over a bandwidth of 5 GHz. It can be seen that the time delay achieved by EBG element is usually much smaller than that of a SAW device, although its operation bandwidth can easily reach tens of GHz.

An electrical delay line based on an SAW or EBG device suffers from either a small operation bandwidth or a small time delay. In wideband microwave communication and radar applications, delay lines with a large bandwidth and large tunable time delay are required [17, 18]. Recently, there has been an increasing interest in using photonics to generate and process microwave signals, thanks to the broad bandwidth offered by modern photonics [19, 20]. Fig. 1.1 illustrates a microwave photonic system for the generation of a time delay for a microwave signal. As can be seen, a microwave signal is modulated on the light from a tunable laser source

(TLS) and sent to a photonic delay line. The delay line provides a time delay to the signal that travels through it. The microwave is then recovered by a photodetector (PD) at the output of the optical delay line. The system shown in Fig. 1.1 can achieve a large bandwidth up to hundreds of GHz that is only limited by the bandwidths of the modulator and the PD. In the system, the optical delay line can be realized by a variety of optical devices, such as a single mode fiber (SMF), a dispersion compensating fiber (DCF), a fiber Bragg grating (FBG), an integrated waveguide [21-24] and a photonic crystal waveguide [25, 26]. In this thesis, SMFs are used to provide large fixed time delays, while FBG-based delay lines are used to achieve tunable time delays.

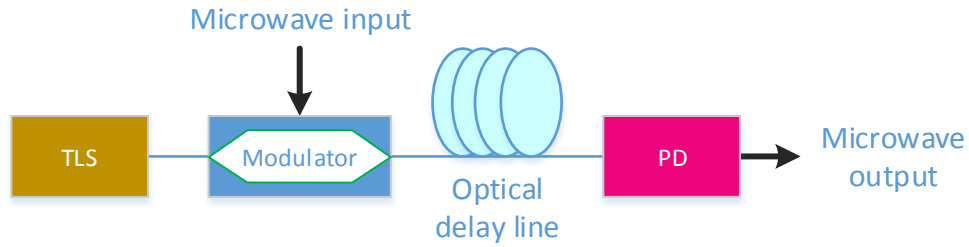


Fig. 1.1 Block diagram of a microwave photonic system to generate a time delay to a microwave signal using an optical delay line. TLS: tunable laser source; PD: photodetector.

The time delay introduced by an optical fiber with a length of L can be expressed as:

$$\tau = \frac{n_{eff}(\lambda) \cdot L}{c} \quad (1-1)$$

where $n_{eff}(\lambda)$ is the effective refractive index of the optical fiber at the wavelength of λ ; c is the light speed in vacuum. Due to the low loss of an optical fiber, a time delay as large as several milliseconds is possible for an optical signal by using a long SMF. However, a delay line with a large time delay tuning range is difficult to implement as it is difficult to change the length of an

optical fiber. On the contrary, a tunable delay line can be usually realized based on free-space optics, in which a light is coupled out of an optical fiber into the free space with an optical lens and then re-focused into another fiber with a second lens. When the physical distance between the two lenses is changed, the time delay will be changed accordingly. Since mechanical elements are used, such a tunable delay line is usually bulky and lossy.

To avoid the use of mechanical elements, a delay line with a tunable time delay can be implemented by exploiting the chromatic dispersion effect in an optical fiber with the assistance of a TLS. Due to the chromatic dispersion, the effective refractive index of an optical fiber is dependent on the optical wavelength. According to (1-1), when the optical wavelength is tuned from λ_0 to λ , the resultant time delay change can be expressed as

$$\Delta\tau = \frac{L}{c} [n_{eff}(\lambda) - n_{eff}(\lambda_0)] \quad (1-2)$$

Within a small wavelength tuning range, the effective refractive index can be considered as a linear function to the optical wavelength, i.e., high order dispersion is ignored. The time delay change in (1-2) can be rewritten as:

$$\Delta\tau = DL\Delta\lambda \quad (1-3)$$

where $\Delta\lambda = \lambda - \lambda_0$ and

$$D = \frac{1}{c} \cdot \frac{n_{eff}(\lambda) - n_{eff}(\lambda_0)}{\lambda - \lambda_0} \quad (1-4)$$

is the dispersion coefficient and can be considered as a constant within a small wavelength tuning range in which high order dispersion is negligible.

Since the dispersion coefficient of an SMF is small, the tunable time delay range is small. To have a large time delay tunable range, we may replace the SMF by a DCF, which has a larger dispersion coefficient and the time delay can be tuned in a much larger range. A standard SMF has a dispersion coefficient of 17 ps/km/nm at around 1550 nm. A DCF can be designed to have a significantly larger dispersion coefficient. A linearly chirped fiber Bragg grating (LCFBG) designed with a linearly increasing or decreasing grating period can also be used as the dispersion element to achieve tunable time delay [27].

For the microwave delay system shown in Fig. 1.1, changing the amount of time delay for the microwave signal can be realized by tuning the wavelength of the TLS. To achieve a larger tunable time delay, it is preferable that the dispersion coefficient of a delay line can be as large as possible, which would make the system bulky due to the required length of the fiber. An FBG is another widely used optical delay line that can provide a large dispersion coefficient with a much greater compactness.

Similar to an EBG delay line, an FBG is a device with a bandgap structure formed by periodically modulating the refractive index of an optical fiber. When the wavelength of the optical signal launched into the FBG satisfies the Bragg condition [27], the optical signal will be reflected. A time delay determined by the location of the reflection will be introduced. To achieve wavelength-dependent group delay response, an FBG can be fabricated with a bandgap structure with a varying period, such as an LCFBG of which the period is linearly increasing or decreasing. When a broadband optical signal is launched into the fiber, different wavelength

components will be reflected at different locations of the LCFBG, resulting in a wavelength-dependent time delay. An LCFBG with a reflection bandwidth of tens of nanometers and a physical length of over one meter is already commercially available, which indicates that it can be used to delay a signal with several THz bandwidth and with a time delay tuning range of 10 ns, making it very promising for wideband microwave communication and radar applications.

In this thesis, we focus on the use of an LCFBG-based optical dispersive delay line (DDL) to function as a wideband electrical DDL to realize ultra-wideband microwave processing, including time reversal, pulse compression, temporal convolution, time-stretched sampling, increasing the TBWP of a microwave signal generator and true-time delay beamforming.

1.2 Major Contributions of This Thesis

Several microwave photonic systems based on optical DDL are proposed and experimentally demonstrated in this thesis for the processing of broadband microwave signals.

First, we demonstrate a broadband and precise microwave time reversal system using an LCFBG-based DDL. By working in conjunction with a polarization beam splitter, a wideband microwave waveform modulated on an optical pulse can be temporally reversed after the optical pulse is reflected by the LCFBG for three times thanks to the opposite dispersion coefficient of the LCFBG when the optical pulse is reflected from the opposite end. An operation bandwidth of over 4 GHz is experimentally demonstrated, which is larger than any other time reversal module ever reported. In addition, the time reversal has a theoretical bandwidth of 273 GHz when optoelectronic devices with sufficiently large bandwidths are used to perform the conversion

between electrical and optical signals. Such a bandwidth is at least an order of magnitude larger than existing digital signal processing systems.

Second, based on the time reversal module, more complex microwave signal processing functions are realized, including temporal convolution of two microwave signals and microwave pulse compression using a matched filter. In the matched filter, the time reversal is used to generate a microwave signal that is the complex conjugate of the impulse response of a microwave photonic filter (MPF), which then can act as the matched filter for the generated signal. Both systems have achieved significantly larger operation bandwidth compared to their electronic counterpart. In the convolution system, two optical DDL are used. One is used to perform time reversal on one of the microwave signals to be convolved, while the other is used to assist the integration operation that is required for the convolution.

For many application, a DDL with a large dispersion is needed. However, the dispersion of an LCFBG-based optical DDL is limited by its physical length at a given operation bandwidth. To overcome this limitation, a fiber optic recirculating loop incorporating an LCFBG is proposed. When an optical signal recirculates in the loop, it will be reflected and dispersed by the LCFBG multiple times, resulting in a significantly increased equivalent dispersion coefficient and a maximum tunable time delay exceeding the length limit of the DDL. The recirculating loop is used to implement a photonic time-stretched sampling system for microwave signal with an equivalent sampling rate of 2.88 TSa/s. This sampling rate is the highest ever reported for a single-shot sampling system. The recirculating loop is also used to increase the TBWP a signal generated by a photonic microwave arbitrary waveform generator (AWG), and to achieve

tunable true-time delay beamforming for a PAA. Again, large operation bandwidths are achieved for both systems, making them highly desirable for modern radar systems.

1.3 Organization of This Thesis

This thesis consists of nine chapters.

In Chapter 1, a brief introduction to electrical delay lines, optical delay lines and dispersive delay lines is presented. The applications of DDLs for the processing of broadband microwave signal are discussed.

In Chapter 2, an introduction to an FBG and an LCFBG is given. An LCFBG and a dispersive loop to be used as DDLs for broadband signal processing are theoretically investigated. Several signal processing functions that can be realized by an LCFBG-based DDL or a dispersive loop are discussed, including time reversal, pulse compression, temporal convolution, time-stretched sampling, large TBWP waveform generation and true-time delay beamforming.

In Chapter 3, the implementation of wideband and precise microwave time-reversal is demonstrated based on the triple use of an optical DDL, which is an LCFBG.

In Chapter 4, a microwave photonic system, which consists of an MPF and a time reversal module, is demonstrated for the simultaneous generation and compression of a microwave pulse with a large TBWP.

In Chapter 5, temporal convolution of two broadband microwave signals is demonstrated based on a microwave photonic system, in which only a low speed PD is needed.

In Chapter 6, a time-stretching sampling system with an extremely high sampling rate is demonstrated with a fiber-optic recirculating loop, in which an LCFBG is incorporated as the optical DDL to provide dispersion to a signal recirculating in the loop for multiple times.

In Chapter 7, a microwave photonic signal generator based on the spectral shaping and wavelength-to-time (SS-WTT) mapping technique is demonstrated, in which a fiber optic recirculating loop incorporating an LCFBG is used to perform WTT mapping and to achieve a long temporal duration of generated signal.

In Chapter 8, a photonic true-time delay beam forming system is implemented, which is realized by controlling the number of round trips of a microwave signal recirculating in a recirculating loop using an optical switch.

In Chapter 9, a conclusion is drawn. Future work is also discussed.

CHAPTER 2 SIGNAL PROCESSING BASED ON A DISPERSIVE DELAY LINE

A DDL introduces different time delays to different spectral components of an input signal. When the input signal is a short pulse, the spectral information of the pulse will be mapped and can be processed in the time domain using a DDL. When the input signal is a chirped pulse, either a temporally compressed or stretched pulse can be achieved at the output of the DDL. When the input signal is a continuous wavelength (CW) signal, a wavelength-dependent time-delay signal can be achieved at the output of the DDL. Based on these concepts, various signal processing functions have been demonstrated based on a DDL, such as microwave filtering [28-33], Fourier transformation [34], frequency up-conversion [35-37], time reversal [38, 39], pulse compression [40], temporal stretching [41-49] and true-time delay beamforming [50, 51]. In this thesis, we focus on using an FBG-based optical DDL in a microwave photonic system to function as a wideband electrical DDL for a microwave signal, which is then used to realize several microwave processing functionalities, including time reversal, pulse compression, temporal convolution, time-stretched sampling, increasing the TBWP of a microwave signal generator and true-time delay beamforming.

2.1 Fiber Bragg Gratings Based Delay Lines

In this thesis, FBGs are designed to have certain group delay responses for different applications. An introduction to FBG fabrication and the operation principle of an FBG-based delay line is given in this Section.

2.1.1 FBG basics

An FBG is a device with a bandgap structure formed by periodically modulating the refractive index of an optical fiber [27]. Fig. 2.1 illustrates the fabrication of an FBG based on the phase mask technique [52]. The ultra-violet (UV) sensitivity of the fiber is usually created by doping germanium in the fiber core, which can be further enhanced by hydrogen-loading. First, a photo-sensitive fiber is placed closely to a phase mask, which is illuminated by a UV laser beam that scans along the fiber. An interference pattern between the -1st and +1st order diffracted light waves is generated behind the phase mask and projected to the photosensitive fiber. The interference pattern has a period half that of the phase mask. A periodic refractive index change is created in the fiber core as the UV exposed area will have a slightly increased refractive index, which is generally in the order of 10^{-4} depending on the time of exposure and the intensity of the UV laser beam.

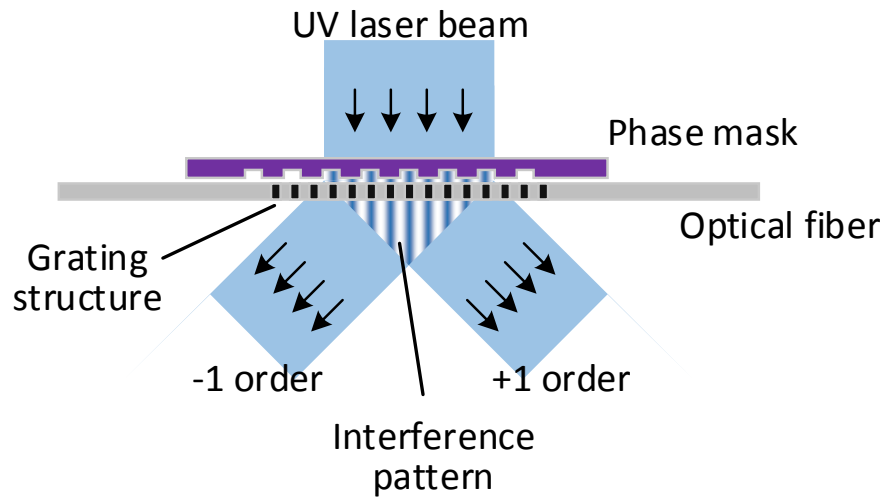


Fig. 2.1 FBG fabrication based on the phase mask technique.

Fig. 2.2 shows the operation of a uniform FBG. The fundamental principle of an FBG is the Fresnel reflection that is induced when light travels in a medium with a varying refractive

index. Due to the periodicity of the refractive index, the weak Fresnel reflection of light of a certain wavelength can be strongly enhanced if the Bragg condition is satisfied, which leads to strong reflection, while the wavelengths that do not satisfy the Bragg condition will be transmitted unaffectedly. For a uniform FBG, the reflected wavelength λ_B , also known as the Bragg wavelength, is given by the Bragg condition

$$\lambda_B = 2n_{eff}\Lambda \quad (2-1)$$

where n_{eff} is the effective refractive index of the fiber taking into consideration of the refractive index modulation in the grating region and Λ is the period of the refractive index modulation. The bandwidth and the reflectivity of a uniform FBG, on the other hand, is determined by its physical length and the refractive index modulation depth.

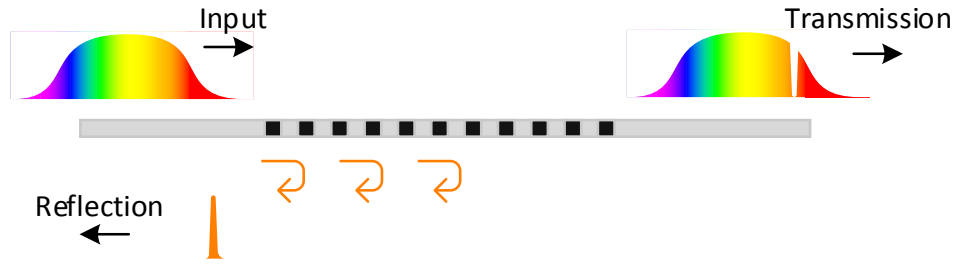


Fig. 2.2 The illustration for the operation of a uniform FBG.

The spectral response of an FBG can be fully characterized with the transfer matrix method [27]. For example, Fig. 2.3 shows a simulated reflection spectral response of a uniform FBG with a length of 5 mm, a grating period of 537.8 nm and a refractive index modulation depth of 2.5×10^{-4} . A reflection band at 1550.1 nm can be observed, with a bandwidth of approximately 0.3 nm. It should be noted that, for a uniform FBG, the group delay variation within the reflection band of the uniform FBG is small, which is a few ps in our case.

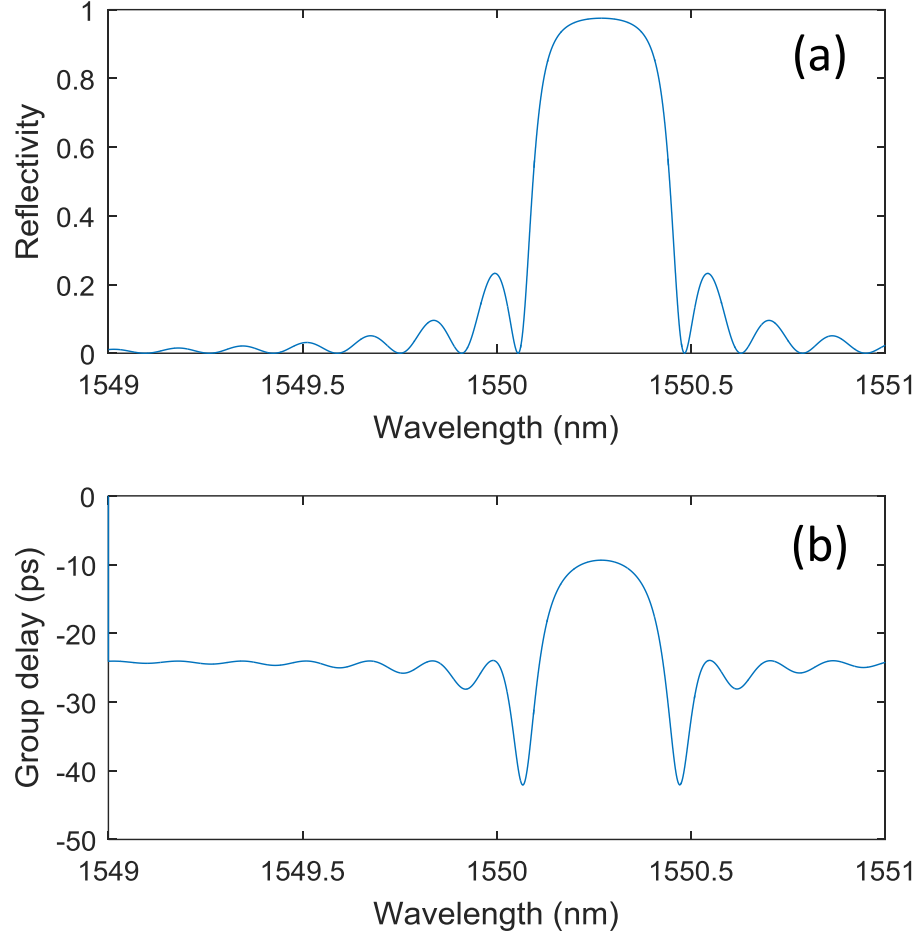


Fig. 2.3 The simulated spectral response of a uniform FBG. (a) Amplitude response; (b) group delay response.

2.1.2 LCFBG and dispersive loop

In an optical delay line, an FBG is usually used in the reflection mode, as a wavelength-dependent time delay can be conveniently achieved by letting light with different wavelengths reflected at different locations within an FBG. Such wavelength-dependent time delay is significantly greater than that of an FBG working in the transmission mode where all transmitted light travels through the whole grating region. The time delay that a uniform FBG can introduce to a reflected signal is determined by its location and thus is inconvenient to tune. To achieve a

tunable time delay, an LCFBG may be used. An LCFBG is a special FBG of which the period of refractive index modulation is linearly increasing or decreasing, as shown in Fig. 2.4. The LCFBG can be seen as some cascaded uniform FBGs with an increasing or decreasing Bragg wavelength. According to (2-1), different wavelength components of an input optical signal will be reflected at different locations in the LCFBG. A wavelength-dependent time delay can be achieved from a reflected signal. The maximum time delay difference between two reflected wavelengths are determined by the physical length of the LCFBG, as indicated in Fig. 2.4.

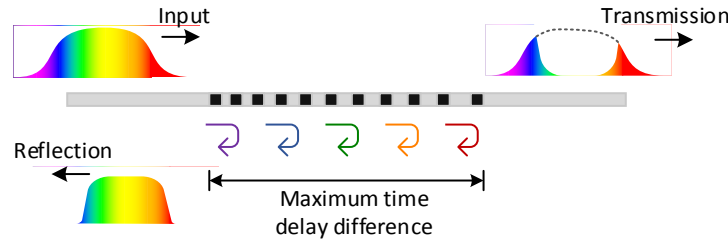


Fig. 2.4 The illustration for the operation of an LCFBG.

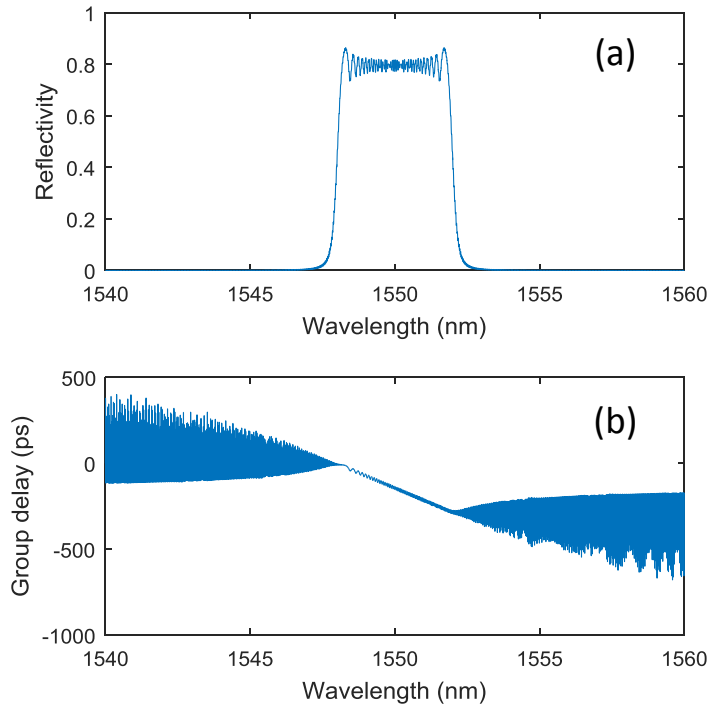


Fig. 2.5 The simulated spectral response of an LCFBG. (a) Amplitude response; (b) group delay response.

Fig. 2.5 shows the simulated spectral response of an LCFBG working in the reflection mode. In the simulation, the refractive index modulation period of the LCFBG linearly increases from 537.13 to 538.51 nm within its total length of 3 cm. The refractive index modulation depth is set to be 2.5×10^{-4} . It can be seen that an LCFBG has a wider reflection band of 4 nm at 1550 nm, due to its changing refractive index modulation period. More importantly, a wavelength-dependent linearly changing group delay response can be observed in the reflection band of the LCFB. The maximum time delay difference is calculated to be 300 ps, which is approximately equal to the time required for an optical signal to propagate for a round trip in the 3-cm long LCFBG.

An LCFBG-based tunable electrical delay line can be implemented by using the LCFBG as a reflector and a TLS as the optical carrier for the microwave signal (See Fig. 1.1). However, the time delay tuning range of an LCFBG-based delay line is fundamentally limited by its physical length. For a pulsed electrical signal, it is possible to use a dispersive fiber optical recirculating loop to significantly increase the tuning range of an LCFBG-based tunable delay line. Fig. 2.6(a) shows the schematic diagram of such a loop, which consists of an LCFBG via an optical circulator and a 2×2 switch. An input signal is launched into the dispersive loop by setting the switching at the cross state. When the pulse enters the loop, the switch is changed to bar state which forms a closed loop that a signal can recirculate for N round trips. After a certain round trips, the switch returns to the cross state and the signal can be directed out of the loop. Fig. 2.6(b) shows the expected group delay response of the dispersive loop when a signal recirculate in the loop for 1 to 7 round trips. Compared to Fig. 2.5, the time delay tuning range of the dispersive loop is N times as large as that of a single LCFBG.

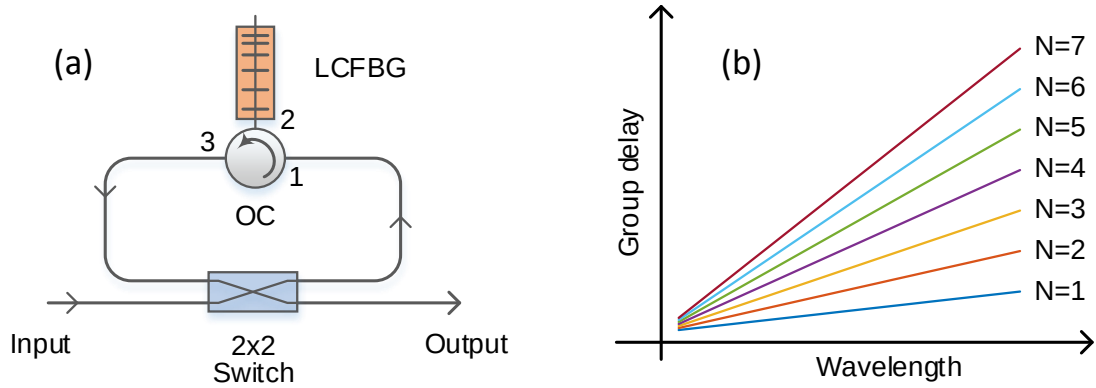


Fig. 2.6 (a) A dispersive fiber recirculating loop incorporating an LCFBG to achieve a large time delay tuning range; (b) the group delay response of the loop when a pulse recirculates in the loop for different number of round trips controlled by the 2x2 switch.

2.2 Signal Processing Based on a Single LCFBG

In this section, three signal processing functions that can be realized based on a single LCFBG to provide opposite dispersion coefficients when reflection light from different ends are discussed, including microwave time reversal, pulse compression based on matched filtering and microwave temporal convolution.

2.2.1 Time reversal

Time reversal, also known as phase conjugation in optics, is a technique widely used to increase the resolution of a detection system. Using time reversal, the energy of a signal can be focused in a detection system with a resolution that is much higher than the value of the signal wavelength [53-55]. In an acoustic time reversal system [56], for example, a short acoustic pulse is sent from a source that propagates through a complex medium and is captured by a transducer array. The recorded signal is digitized, time reversed digitally, and then transmitted. Recently, an optical time reversal system was implemented to focus light through scattering media [38]. In

2004, time reversal of an microwave signal was proposed to overcome the multipath problem for microwave communications [39]. It is shown that, time reversal is not only capable of solving the multipath problem, it can also control the microwave power distribution by focusing more power to the detector, which has been theoretically and experimentally verified in [57] and [58]. Since then, microwave time reversal has attracted significant research interests due to its promising applications in microwave imaging and microwave communication. A microwave imaging system with a significantly improved resolution by time reversal was proposed for breast cancer detection [59, 60]. In [61-63], microwave time reversal was used for hyperthermia treatment of cancer thanks to its capability to focus electromagnetic power. A microwave super-resolution system was demonstrated in [64], in which time reversal was used to focus a microwave signal with a resolution of one thirtieth of the microwave wavelength, a value that is beyond the diffraction limit. In [65], it was demonstrated that using time reversal, the phase distortion of a UWB signal in a communications system can be effectively compensated.

It is similar to acoustic time reversal, to implement microwave time reversal, digital solutions are usually employed, which involve analog-to-digital conversion (ADC), digital signal processing (DSP), and digital-to-analog conversion (DAC). In a lab environment, these functions were implemented using a real-time oscilloscope to perform sampling, a computer to perform DSP, and an arbitrary waveform generator to perform DAC [65]. The key limitations of a digital microwave time reversal system are the relatively slow speed and small bandwidth, and are only suitable for signal processing with a frequency and bandwidth of a few GHz. For example, the bandwidths of the digital microwave time reversal systems are only 2 MHz [39], 20 MHz [60], and 150 MHz [57]. In [65], a pulse with an effective bandwidth of 9.6 GHz was generated, but at the cost of a very expensive electronic AWG. For many applications, time reversal of a high

frequency and wideband microwave signal is highly demanding. It has been theoretically proved that time reversal of a microwave signal with a wider bandwidth can significantly improve the focusing efficiency of a microwave imaging system [66]. Photonic solutions have been proposed to implement high-frequency and wideband microwave time reversal. In [67], microwave time reversal was optically realized by using the three photon echo effect in an erbium-doped YSO crystal. An unprecedented time duration of 6 microseconds was demonstrated. The application of the time reversal in a temporal imaging system was discussed in [68]. Despite the extremely long time duration, the bandwidth of the time reversal was limited only to 10 MHz, which is small and could be easily achieved by a digital time reversal system. In [69], a microwave photonic system to achieve broadband microwave time reversal using a temporal pulse shaping system was proposed. Theoretically, the bandwidth can be as large as 18 GHz. However, since two independent dispersive elements were used in the system, a relatively large dispersion mismatch between the two dispersive elements was resulted, which led to large waveform distortions with a reduced system performance (defocusing).

2.2.2 Pulse compression

Pulse compression has been widely used in modern microwave sensing and communication systems to increase the range resolution [40]. Pulse compression is implemented by radiating a spread-spectrum microwave waveform, such as a linearly chirped microwave waveform (LCMW) or a phase-coded microwave waveform (PCMW), to the free space. When the radiated waveform is reflected by a target and received at a receiver, the waveform is largely compressed by passing it through a matched filter or by correlating with a reference waveform,

resulting in a significantly increased range resolution. In this thesis, the compression of a microwave pulse is investigated based on optical DDL.

Assume that a transmitted electrical pulse has a single-tone carrier frequency within a time window with a width of T , the bandwidth of the signal is then $B=1/T$, which is inversely proportional to the bandwidth. The range resolution is determined by the duration of the pulse and is expressed as

$$\Delta r = \frac{cT}{2} = \frac{c}{2B} \quad (2-2)$$

The most straightforward way to improve the range resolution of a microwave sensing system is to use a shorter electrical pulse, as can be seen from (2-2). However, this is not always practical in real applications, as a short pulse requires an extremely large bandwidth and a high peak power that cannot be handled by most electrical components. Specifically, the high peak power set a rigid requirement for the microwave wave amplifiers, it is also a challenge for the antennas due to the arcing effect that takes place at over one megawatt peak power. In order to solve this problem, microwave waveforms with large bandwidth and long temporal duration are more often transmitted, which can achieve comparable range resolution and requires a much lower peak power. These waveforms, also known as waveforms with large TBWP, include LCMW, binary phase coded waveform and some waveforms with other phase code such as linear recursive sequences, quadriphase codes, polyphaser codes and Costas codes. When the waveforms with a large TBWP are used as the transmitted signal and detected by the microwave receiver, matched filters or signal cross-correlators are usually used to compress the received

waveform to achieve a range resolution much higher than that determined by the temporal duration of the transmitted waveforms.

Here, two pulse compression examples based on an LCMW and a binary phase coded waveform using matched filtering and signal cross-correlation are investigated. First, the compression of an LCMW is simulated. An LCMW with a temporal duration of $0.4\ \mu\text{s}$ and a bandwidth of 125 MHz is generated. Fig. 2.7 shows the waveform and the corresponding spectrogram. A linearly increasing instantaneous frequency can be observed, with a chirp rate of $31.3\ \text{MHz}/\mu\text{s}$. The TBWP of the signal is calculated to be 50. Then we design a matched filter to have a group delay dispersion opposite to the chirp rate of the LCMW to perform pulse compression. The magnitude response and group delay response of the matched filter are shown in Fig. 2.8, which has a flat-top passband from DC to 125 MHz, and a dispersion of $-32\ \text{ns}/\text{MHz}$. The generated LCMW then propagates through the matched filter. A compressed waveform is derived by multiplying the spectrum of the LCMW and the frequency response of the matched filter. Fig. 2.9 shows the compressed waveform. A narrow peak with a temporal width of 4.8 ns is achieved, which indicates a compression ratio of 83.3. It can be seen that, even though the transmitted signal has a temporal duration of 400 ns, a time resolution of 4.8 ns can be achieved by using a matched filter.

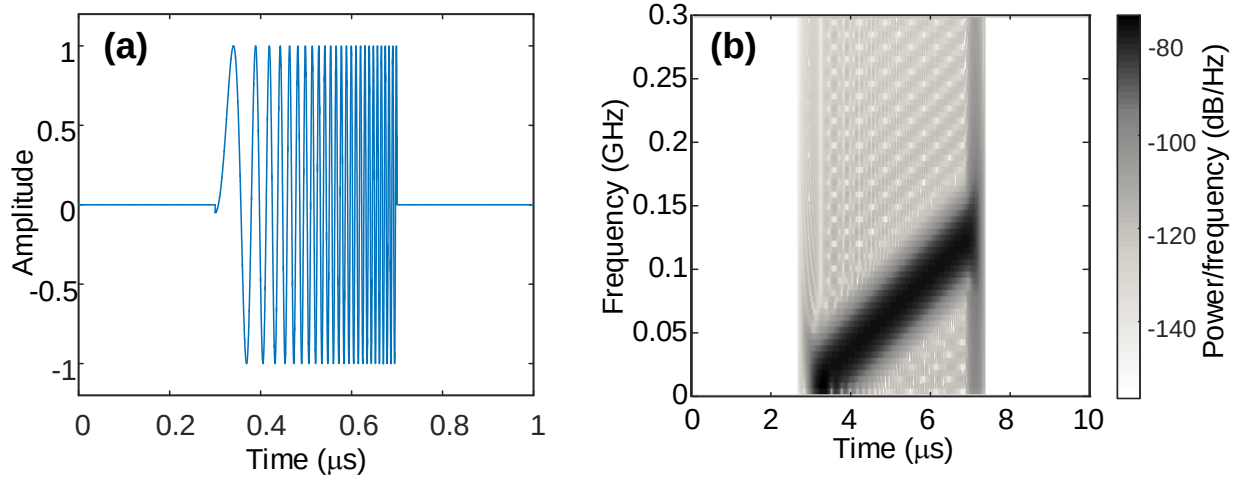


Fig. 2.7 (a) Waveform and (b) spectrogram of the LCMW used in the simulation.

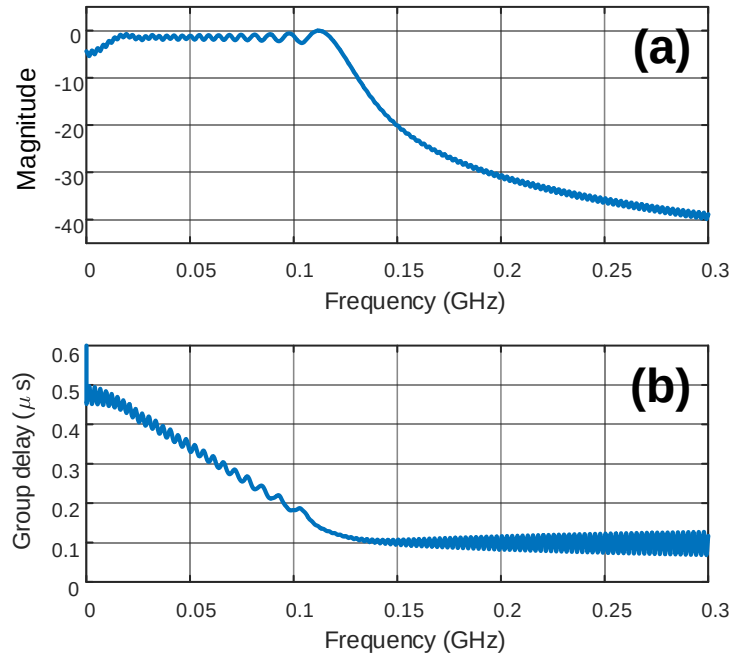


Fig. 2.8 The frequency response of the designed matched filter: (a) magnitude; (b) group delay.

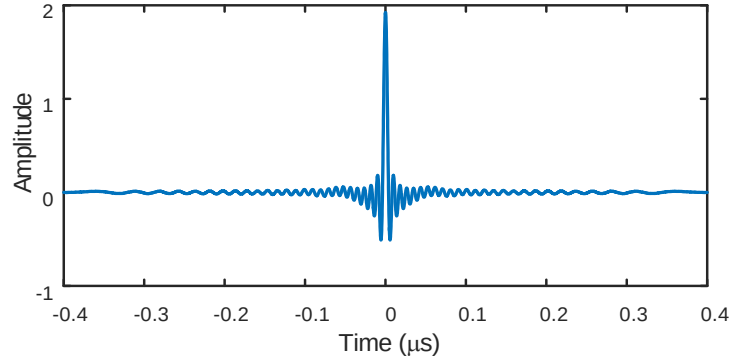


Fig. 2.9 The compressed pulse with a pulse width of 4.8 ns.

Then we show the compression of a binary phase-coded waveform using signal cross-correlation technique, where the transmitted pulse is used as a reference to cross-correlate with the received signal. A peak will be observed when the received signal matches the reference. Again, a microwave waveform with 16-bit pseudorandom binary phase code is generated at a bit rate of 200 Mbit/s and a carrier frequency of 1 GHz, as shown in Fig. 2.10. The compressed pulse is calculated by auto-correlating the waveform. The correlation result is shown in Fig. 2.11. A peak with a width of 4.3 ns can be seen, corresponding to a compression ratio of 18.6, which is approximately equal to the number of phase code bits used in the simulation. Further simulation shows that the randomly generated phase codes only influences the sidelobe suppression ratio (relating to the signal-to-noise ratio of the microwave receiver) but not the pulse compression ratio.

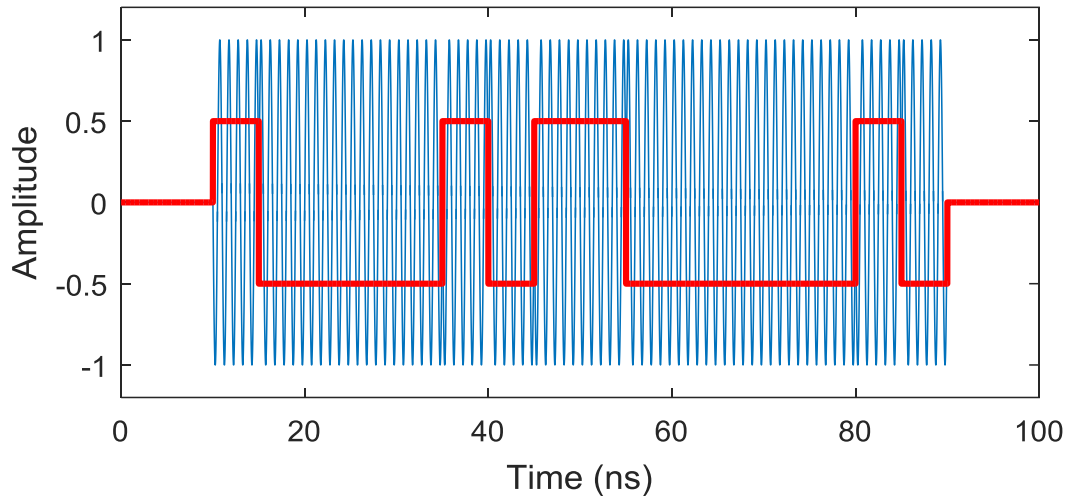


Fig. 2.10 The 16-bit pseudorandom binary phase coded signal (blue) and the phase code (red $\times \pi$).

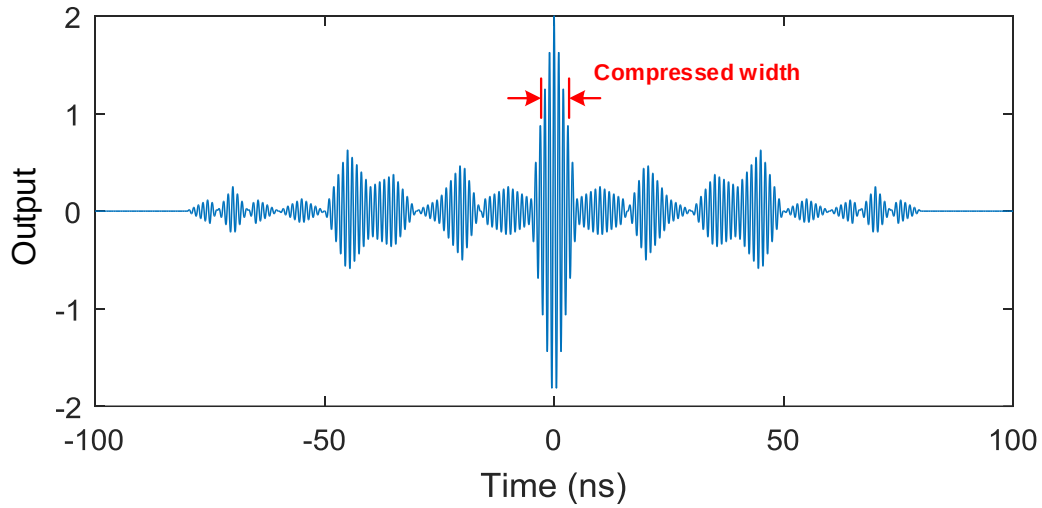


Fig. 2.11 The waveform achieved by compressing the phase coded waveform using cross-correlation technique.

It should be noted that pulse compressing based on matched filtering and signal cross-correlation are two identical processes mathematically. Matched filtering is realized in frequency domain, while the cross-correlation is performed in time domain.

Photonic techniques have been extensively investigated for the generation of spread-spectrum microwave waveforms, including LCMWs [70-85] and PCMWs [86-88]. On the other

hand, very few photonic techniques have been proposed for microwave pulse compression. For the compression of an LCMW in the electrical domain, a dispersive filter with its spectral response that is a complex conjugate version of the spectrum of the LCWM can be used as a matched filter, which can be implemented using a SAW device [89], a C-section delay line [90] or a synthesized microwave phaser [91]. However, the bandwidth of an electrical matched filter [89-91] is usually limited to less than a few GHz. A photonic matched filter has the potential to overcome the bandwidth limitation when used for pulse compression in a radar system. In [92], an MPF with a quadratic phase response was demonstrated for LCMW compression, in which the MPF was implemented by passing a single sideband modulated optical signal through an FBG that has a quadratic phase response. Thanks to optical phase to microwave phase conversion through single sideband modulation and heterodyne detection, an MPF with a quadratic phase response was achieved. The bandwidth of the MPF was 3 GHz, which can be much wider if the FBG is designed to have a wider bandwidth. In [93], a four-tap MPF was experimentally demonstrated to function as a matched filter for pulse compression of a binary PCMW with a carrier frequency of 6.75 GHz. The filter can be reconfigured by changing the wavelength spacing of the optical carriers to compress a microwave waveform with a different phase coding. Since the tap number is determined by the code length of the PCMW, which can be long, thus the system is complicated for long length code compression. In [94], an MPF with a quadratic phase response was demonstrated based on a broadband optical source sliced by a Mach-Zehnder interferometer (MZI). By passing the sliced optical wave through a nonlinear dispersive element, a finite impulse response (FIR) filter with nonuniform tap spacing corresponding to a quadratic phase response was implemented. A bandwidth of 2.5 GHz and a dispersion of 12 ns/GHz were experimentally achieved. A similar approach was proposed in [95].

To eliminate the dispersion induced power penalty, a phase modulator placed in one arm of the MZI was used instead of an intensity modulator that was placed at the output of the MZI. The bandwidth of the MPF was 4 GHz.

2.2.3 Temporal convolution

Another approach to realizing pulse compression is by signal correlation or convolution, in which a received signal is correlated with a reference signal. Signal correlation or convolution performs pulse compression in time domain and hence offers better configurability for microwave receivers. Here we use the convolution operation as an example. The temporal convolution of two signals is different from a filtering operation where a microwave signal is convolved with the impulse response of a microwave filter due to the multiplication between the spectrum of the signal and the frequency response of the filter. In many cases, the temporal convolution can provide better flexibility in signal processing as compared to a filtering operation since the spectral response of a filter is fixed, but for many signal processing applications, the spectra of the two microwave signals to be convolved need to be updated in real time. For example, the temporal convolution was used for the real-time distortion correction in imaging processing in [96]. Image deblur was achieved by the convolution between the image data and an impulse response function measured for the specific distortion. In [97], the detection of the phase information of a periodic signal in noise is realized by convolving the corrupted signal with its cumulant (accumulation with certain algorithm) version. The convolution in [96] and [97] are done by digital signal processing techniques. The convolution of two signals based on an analog system, especially a photonic analog system, has a potential to achieve a high operation speed for such applications.

Temporal convolution is more complex to implement compared to other signal processing functions that mentioned previously [37, 67, 68, 98-111], as it requires a combination of time reversal, time delay, signal multiplication and integration. However, thanks to the development of photonic signal processing in the past few decades, most of the operations have been demonstrated using photonic techniques.

Photonic microwave time reversal was first demonstrated in [67, 68], which has achieved an extremely long reversing time window of 6 μs using three photon echo effect. However, the operation bandwidth is only limited to 10 MHz. In [100, 101], we demonstrated that a wideband microwave waveform can be temporally reversed by a single LCFBG, in which we achieved a precise time reversal with an operation bandwidth of 4 GHz within a time window of 10 ns. The microwave time reversals reported in [67, 68, 100, 101] can be good candidates for a temporal convolution calculation.

On the other hand, photonic integrators for both optical and microwave signals are also widely reported, which can be realized using an FBG [102], a microring resonator [103, 104], an active Fabry-Perot cavity [105] or an optical dispersive device [106-109]. The multiplication of microwave signals can be easily achieved by cascading two electro-optical modulators. The only challenge in realizing a temporal convolution is to realize a changing time delay difference between the two signals to be convolved.

2.3 Signal Processing Based on a Dispersive Loop

The maximum dispersion of an LCFBG-based DDL is limited by its physical length when the operation bandwidth is fixed. However, a DDL with a large dispersion coefficient is required

in many signal processing systems. In this section, three signal processing systems implemented based on a dispersive loop with a large dispersion coefficient is investigated, including a microwave time-stretched sampling system, a large TBWP waveform generation system and true-time delay beamforming system.

2.3.1 Time-stretched sampling

For time reversal, pulse compression and temporal convolution introduced above, an optical DDL may be able to provide a sufficiently large time delay to the order of 10 ns within a bandwidth of a few hundred GHz. However, there are applications where an even larger dispersive group delay is required, such as time-stretched sampling.

The ever-increasing bandwidth of modern microwave sensing and communications systems has imposed new challenges on signal processors to operate at a very high sampling rate. The use of the conventional sampling techniques may not be able to meet the demand. To realize broadband sampling, a solution is to use photonic-assisted sub-Nyquist sampling. Numerous approaches have been proposed, such as optical down-sampling [112, 113], optical undersampling [114], optical pseudorandom sampling [115], compressive sampling [116], and optical time stretched sampling [41-49]. Among these techniques, the optical time-stretched sampling has been considered an effective solution for wideband microwave signal processing.

In an optical time-stretched sampling system, a microwave waveform is modulated on a pre-dispersed optical pulse, which, after the modulation, travels through another dispersive element for time stretching. The second dispersive element should have a much greater group delay dispersion (GDD) coefficient than the first element. A time-stretched microwave

waveform will be generated when the pulse is sent to a PD and a slow version of the original waveform is obtained [70]. This technique was first proposed by Coppinger et al. in [42], demonstrating a sampling rate that is $1/3.25$ of the Nyquist frequency. Since then, new efforts have been dedicated to further reduce the sampling rate by increasing the stretching factor. In [43], dispersion-based time stretched sampling with a stretching factor of 5 combined with compressive sensing was demonstrated to achieve to a sampling rate as low as only $1/40$ of the Nyquist frequency. The sampling rate is then further reduced to $1/80$ of the Nyquist frequency by using a time stretched sampling module with a time stretching factor of 20 [44]. In [45], time-stretched sampling using a coherent receiver to improve the detection sensitivity by cancelling the dispersion-induced impairments and optical nonlinearity was demonstrated. A stretching factor of 24 was achieved by using two dispersive elements with two GDD coefficients of 45 and 1045 ps/nm for the pre-dispersion and time stretching, respectively. In [46], time-stretched sampling of a continuous-time signal was demonstrated based on virtual time gating, where a stretching factor of 1.5 was achieved. To overcome the “big data” problem associated with time-stretched sampling, Asghari et al. demonstrated a sampling solution based on nonlinear time stretching [47]. An equivalent stretching factor of 200 was achieved. The system has been further enhanced to achieve real-time bandwidth suppression factor of 500 [48]. In the system, a programmable optical filter was used to modulate the microwave waveform to the optical pulse, which may make the system complicated and costly. In [49], an unprecedented time stretching factor of 250 was realized by using a pre-dispersion element with a GDD coefficient of 41 ps/nm and a double-pass dispersive element with a GDD coefficient of -10246 ps/nm. To achieve such a large dispersion, an extremely long DCF is used. To compensate for the loss in the long DCF, four stimulated Raman amplifiers pumped by four high-power laser diodes were employed.

Although the system could realize an effective sampling rate of 10 Tsamples/s, the use of a long DCF and multiple Raman amplifiers makes the system rather bulky and complicated. For a time stretching element with a fixed GDD coefficient, to achieve a large stretching factor, the pre-dispersion element could be selected to have a relatively small GDD coefficient. The consequence of using a pre-dispersion element with small dispersion is that the input optical pulse cannot be sufficiently pre-stretched to have a large time duration, to allow a microwave waveform with a long duration to be modulated on the pre-stretched input pulse. Therefore, the fundamental solution to have a large stretching factor for a long-duration microwave waveform is to use a time stretching element with a large GDD coefficient.

In fiber optics, a dispersive element can be an SMF, a DCF, or an LCFBG. Since an SMF has a relatively small dispersion coefficient, it is rarely used in a time stretched sampling system, especially as the time-stretching dispersive element. A DCF, on the other hand, can have a dispersion coefficient that is several times greater than that of an SMF. However, to achieve large time stretching, a DCF with a length of several tens [45] or even hundreds [49] of km is required. Thus, the system is still bulky and lossy. An LCFBG has been proved to be a highly effective dispersive element with low insertion loss and small nonlinear effects [27, 117]. The GDD of an LCFBG is proportional to its grating length and inversely proportional to its bandwidth. For time stretching applications, the bandwidth of an LCFBG is usually controlled to be equal to the bandwidth of the optical pulse. Hence, to have a large GDD coefficient, an LCFBG with a long length is needed. For example, a 10-cm long LCFBG with 1-nm bandwidth has a GDD coefficient of 1000 ps/nm. To further increase the GDD coefficient, the length of the LCFBG should be further increased. Although an LCFBG with a length greater than 1 m is commercially available, the size is large and the fabrication is complicated and costly.

The use of a microwave waveguide with a large GDD coefficient has been explored in the past few years for spectrum analysis [118, 119]. Recently, it was demonstrated [120] that by forming a dispersive loop that incorporates a microwave dispersive element and a microwave amplifier, an equivalent microwave dispersive element with a GDD coefficient that is several times greater than that of the original dispersive element can be achieved by recirculating the microwave waveform in the loop. Compared with a simple cascade of multiple dispersive elements to achieve an equivalent dispersive element with a large GDD coefficient, the recirculating dispersive loop has the advantages such as a smaller device footprint, lower insertion loss (less than 3 dB loss per round trip compared to tens of dB of a long optical fiber) and better signal-to-noise ratio (SNR) [120]. The major limitations of an electrical dispersive loop are the small bandwidth, usually below 1 GHz, and high loss. In addition, the maximum time delay is limited, although the constant time delay provided by a coaxial cable is already much larger compared to that of a waveguide. To implement a dispersive loop with a large GDD coefficient over a large bandwidth, a solution is to use photonic components. In addition to a broad bandwidth, an optical dispersive loop can generate a much longer time delay since a long loop length of several kilometers is possible due to the low loss of an optical fiber. A time-stretched sampling system using an optical dispersive loop can significantly increase the stretching factor over a broad bandwidth.

2.3.2 Large time-bandwidth product signal generation

An optical DDL with an extremely large dispersion coefficient may also be used in a photonic microwave AWG for the generation of a waveform with a large TBWP. Microwave waveforms with a large TBWP have been widely employed in microwave sensors [121, 122],

spread-spectrum communications [123], microwave computed tomography [124], and modern instrumentation. For example, in an active microwave sensor, an LCMW with a large TBWP can be used to improve the range resolution. An LCMW are usually generated electronically, and the temporal duration can be long. However, due to the limited speed of the currently available electronic circuits, the bandwidth and the central frequency of an LCMW generated electronically are usually limited to a few GHz [125-127]. An LCMW with a central frequency and a bandwidth up to tens or even hundreds of GHz may be required in a modern microwave sensor to improve the sensing resolution.

Numerous photonic approaches have been proposed and demonstrated for the generation of an LCMW with a large TBWP [70]. These approaches can be generally classified into three categories: space-to-time pulse shaping [71-73], SS-WTT mapping [74-76, 80-84], and temporal pulse shaping [77]. An LCMW can also be generated using an MPF [78] with a quadratic phase response, using an optoelectronic oscillator (OEO) [85], or optical heterodyning [79]. A space-to-time pulse shaping system is usually implemented using a spatial light modulator (SLM). The SLM has the flexibility in updating in real time the pattern on the SLM, which enables the generation of an arbitrary microwave waveform. The major limitations of using an SLM are the relatively high loss and large size [71], [72]. Although the SLM in a space-to-time pulse shaping system can be replaced by an arrayed waveguide grating, the maximum temporal duration of the generated waveform is still limited, in a range of tens of picoseconds, due to the relatively small channel number of an arrayed waveguide grating, developed usually for wavelength-division multiplexing communications applications [73]. Microwave waveform generation based on temporal pulse shaping, an MPF, or an OEO, also has the limitation of small temporal duration.

The phase stability of the waveforms generated by optical heterodyne technique is usually poor [79].

On the other hand, microwave waveform generation based on SS-WTT mapping has the advantages of simplicity, flexibility and relatively low cost. In an SS-WTT mapping system, an ultra-short pulse with a wide spectrum is spectrally shaped by an optical spectral shaper. The shaped spectrum is then mapped to the time domain by a dispersive element to produce a microwave waveform with a temporal shape that is a scaled version of the spectrum of the spectrally shaped pulse. Hence, an LCMW can be generated by designing an optical spectral shaper with a linearly increasing or decreasing free spectral range (FSR), and using a dispersive element with linear dispersion to perform linear WTT mapping. In [74], a fiber-optic spectral shaper with a Michelson interferometer structure using two LCFBGs as two reflectors was proposed and demonstrated for LCMW generation. Due to the wavelength-dependent length difference between the two arms of the Michelson interferometer, a linearly decreasing FSR is achieved which is needed for LCMW generation. In addition to the operation as two reflectors, the LCFBGs also function as a dispersive element to perform linear WTT mapping. An LCMW with a TBWP of around 15 was generated. In [80], an optical spectral shaper was realized by employing two LCFBGs with different chirp rates that are fabricated and superimposed in a fiber. The two LCFBGs form a Fabry-Perot (FP) cavity to have a spectral response with a linearly decreasing or increasing FSR due to the wavelength-dependent cavity length. The use of the spectral shaper to generate an LCMW was demonstrated. An LCMW with a TBWP of 37.5 was achieved. In [81], a spectral shaper implemented using a Sagnac loop mirror with an LCFBG in the loop was reported. Again, a spectral response with a linearly decreasing or increasing FSR was resulted due to the wavelength-dependent loop length. An LCMW with a TBWP of 44.8 was

experimentally generated. In [82], a tilted fiber Bragg grating (TFBG) fabricated in an erbium/ytterbium co-doped fiber was used as a spectral shaper. Since the spectral response of the spectral shaper can be tuned by optically pumping the TFBG, the generation of an LCMW with a tunable chirp rate from 1.8 to 7 GHz/ns was demonstrated. In [83], a configurable microwave waveform generator with a bandwidth up to 60 GHz based on a silicon photonic chip was demonstrated. However, the temporal duration of the waveforms generated in [82] and [83] are limited to 1 ns, with TBWPs of less than 60. In [84], a microwave waveform generator based on SS-WTT mapping was proposed, in which an electrically stabilized MZI was used as the spectral shaper and a DCF was used to perform WTT. A microwave waveform with a temporal duration of 16 ns and a TBWP of 589 was achieved. Using a similar scheme, an LCMW with a TBWP of 600 was generated thanks to the use of a near-ballistic uni-traveling-carrier photodiode with a large bandwidth of over 175 GHz [75]. However, the temporal duration of the microwave waveform is only 15 ns, which is still small for many applications. In [76], a microwave waveform with an arbitrarily long temporal duration was generated by synthesizing a series of LCMWs with different phases. An individual segment of the waveform is an LCMW generated by SS-WTT, in which a programmable optical filter and two modulators are needed which would increase the system complexity. In addition, the temporal duration of an individual segment of the LCMW is only 5 ns. Although the techniques in [74-76, 80-84] can be used to generate an LCMW with a wide bandwidth and a high carrier frequency of over tens of GHz, the temporal duration is usually small which is fundamentally limited by the achievable maximum time delay of the dispersive device used for WTT mapping. For example, the maximum time delay of an LCFBG is only a few nanoseconds, limited by its physical length [27]. Although a DCF with a length of tens of kilometers can achieve a larger dispersion, the high insertion loss needs to be

compensated by a fiber amplifier, such as a distributed Raman amplifier, making the system very complicated [49]. For many applications, a simple and cost-effective approach to generate a microwave waveform with a duration in the order of microseconds or even milliseconds is needed.

2.3.3 True-time delay beamforming

A PAA plays a key role in modern microwave sensing and communication systems as it can provide beam steering at a high speed without mechanical movement and with ultra-high directivity [122]. A beamforming network is required to produce progressive phase or time delays for a PAA, which can be implemented using phase shifters or true-time delay lines. The advantage of using true-time delay lines is that the beam is squint free, thus it is more suitable for broadband applications.

To illustrate the operation principle of a true-time delay beamforming network, we use a uniformly spaced four-element PAA as an example, as shown in Fig. 2.12. The PAA is fed with a wideband microwave signal, which experiences uniformly increasing or decreasing time delay before reaching each antenna element. Assume the time delay difference between two adjacent antenna elements is Δt , the relationship between the phase difference and the beam pointing angle at a given microwave frequency f is [128, 129]

$$\Delta\phi = 2\pi f \Delta t = 2\pi f \frac{d \sin \theta}{c} \quad (2-3)$$

where d is the spacing between two adjacent antenna elements and c is the light velocity in vacuum. The array factor of the PAA is given by

$$\begin{aligned}
G_a(\theta, f) &= \frac{\sin^2(N\Delta\phi/2)}{N^2 \sin^2(\Delta\phi/2)} \\
&= \frac{\sin^2[N\pi(fd/c)\sin(\theta)]}{N^2 \sin^2[\pi(fd/c)\sin(\theta)]}
\end{aligned} \tag{2-4}$$

It can be seen from (2-3) and (2-4) that beam steering can be realized by applying a phase shift $\Delta\phi$ or time delay change Δt to the microwave signals delivered to each antenna element. The traditional way of steering a beam from a PAA is by using microwave shifters. To steer the beam to a desired angle of θ_0 , the amount phase shift that is required can be given by

$$\Delta\phi_0 = \frac{2\pi df_0 \sin \theta_0}{c} \tag{2-5}$$

Note that for other microwave frequencies f , the required phase shift is different. The array factor can be achieved by replacing $\Delta\phi$ with $(\Delta\phi - \Delta\phi_0)$ in (2-4), we have

$$G_a^{phase}(\theta, f) = \frac{\sin^2[N\pi d/c(f \sin(\theta) - f_0 \sin(\theta_0))]}{N^2 \sin^2[\pi d/c(f \sin(\theta) - f_0 \sin(\theta_0))]} \tag{2-6}$$

Differently, if the beam steering is implemented with time delay, we first substitute the desired angle θ_0 to (2-3) to get Δt_0 and $\Delta\phi_0$. Again, replace $\Delta\phi$ with $(\Delta\phi - \Delta\phi_0)$ in (2-4), we have

$$G_a^{delay}(\theta, f) = \frac{\sin^2[N\pi df/c(\sin(\theta) - \sin(\theta_0))]}{N^2 \sin^2[\pi df/c(\sin(\theta) - \sin(\theta_0))]} \tag{2-7}$$

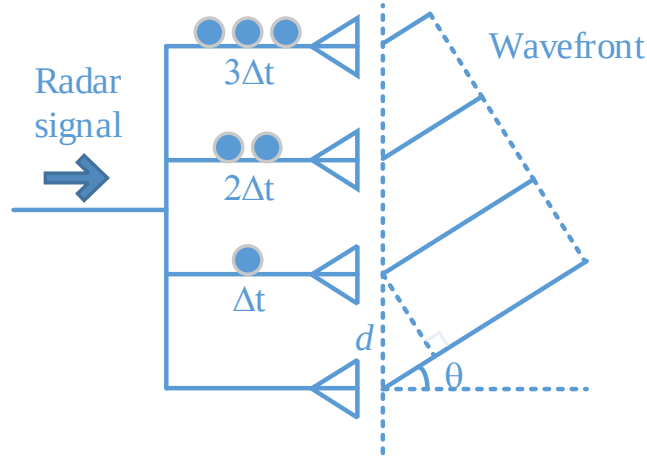


Fig. 2.12 Phased array antenna for beamforming.

Based on (2-6) to (2-7), the beam patterns are simulated for both phase shifter beamforming and time delay beamforming, as shown in Fig. 2.13. As can be seen from Fig. 2.13(a), beam forming based on phase shifters, the array factor is microwave frequency dependent. It leads to an effect called beam squint, in which different microwave frequency will be steered to different angles. The beam squint will get more significant as the bandwidth of the transmitted microwave signal increases, making a beam forming system based on phase shifter only suitable for narrowband microwave sensors. On the contrary, the array factor of a PAA based on true-time delay beam forming is independent of microwave frequency, as can be seen from Fig. 2.13(b), which means that it is beam-squint free and suitable for wideband operation.

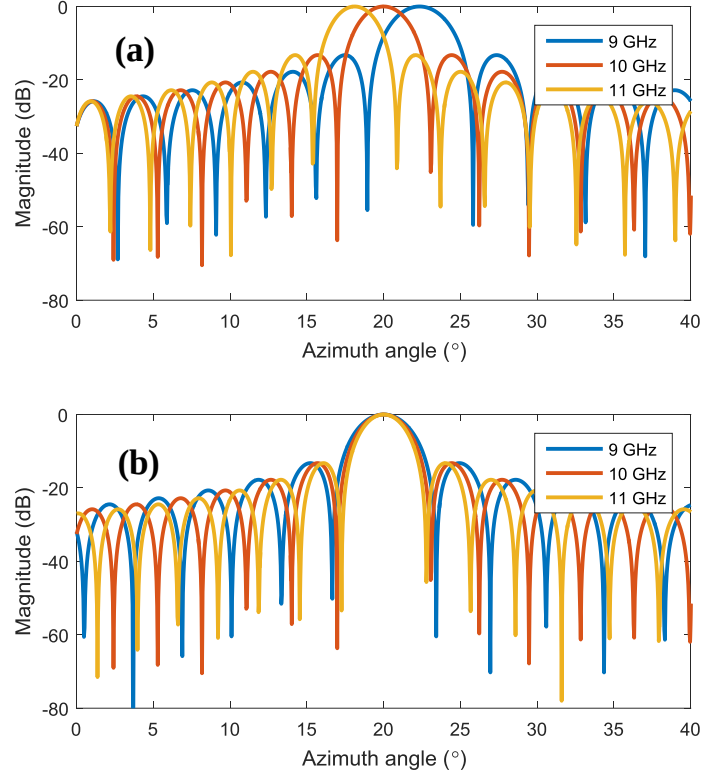


Fig. 2.13 (a) Beam pattern with phase shifter steering; (b) beam pattern with time delay beam forming. In the simulation: $N=40$; $d=1.5$ cm; $f_0=10$ GHz.

A true time delay beamforming network can be realized using electronic delay lines, but with a small bandwidth, a large size and a high loss [130, 131]. In the past few years, numerous photonic true-time delay beamforming networks have been demonstrated. In [132], a free space photonic true-time delay network was demonstrated. The time delay of a microwave signal modulated on an optical carrier is changed by changing the polarization state of the optical carrier so that it can travel through different optical paths. As a free-space optical system generally requires a large number of mirrors and lenses, it is usually very heavy and bulky. Fiber optics and photonic integrated circuits can be used to mitigate these limitations thanks to their small size and low loss. In [133], an FBG array was used to realize true-time delays. The tunability of the time delays was achieved by changing the optical carrier wavelengths. A

maximum true time delay of 233 ps was experimentally demonstrated. Since an FBG array contains a large number of FBGs, to reduce the complexity, in [134] a true-time delay network implemented using a single LCFBG was proposed. Since optical carriers with different wavelengths are reflected at different locations in an LCFBG, different time delays can be achieved for a microwave signal modulated on different optical carriers. In addition, tunable time delays can be achieved by changing the chirp rate of the LCFBG. A time delay tuning range of 100 ps, with a tuning step of 1 ps was demonstrated. In [135], a DCF was used instead of an LCFBG. Tunable true-time delays were achieved by changing the optical carrier wavelengths. Similarly, in [136] an SMF is used as a dispersion element to realize tunable time delay by changing the wavelengths of an optical frequency comb, which is used as the optical carriers for multiple microwave signals. A fiber-optics based beamforming network features a smaller size, but a TLS is usually required in [133, 135, 136] to achieve a large tunable time delay, making the system costly. In addition, the wavelength stability due to tuning may deteriorate, which will affect the time delay accuracy. In [137], instead of using TLSs, a tunable dispersive medium based on the similar effect in [134] was used to implement tunable true-time delay. A true-time delay with a tuning range of 200 ps was demonstrated, which is again very limited. In [138], a multicore fiber was employed to achieve tunable true-time delay. By designing the refractive index profile of the fiber, optical signals travels in different cores will experience different time delays. However, the time delay cannot be conveniently tuned. Photonic true-time delay can also be realized based on SBS [139]. In the gain SBS spectrum, the time delay is wavelength dependent. By changing the optical carrier wavelength which is placed in the SBS spectrum profile, the time delay is tuned. However, the spectral width of an SBS gain profile is very small, which limits the bandwidth of the microwave signal modulated on the optical carrier. Other

techniques to achieve true-time delays include the use of stack integrated micro-optical components [140]. Since movable prism groups are used, the reliability is poor. Recently, an on-chip microwave photonic beamformer based on $\text{Si}_3\text{N}_4/\text{SiO}_2$ waveguide technology was demonstrated [141, 142]. The stability is better than using fiber delay lines. Due to the small size of the chip, the achievable time delay is small, limited to a few hundreds of ps. In [143], a photonic microwave filter was designed to have a frequency response that is similar to a microwave delay line with a tunable time delay. An integrated-optics based beamforming network has an ultra-smaller size, but only small time delays are achievable due to the small size of the chip [140-143].

2.4 Summary

In this Chapter, signal processing functions based on FBG-based dispersive delay lines have been discussed. First, an introduction to an FBG was given. An LCFBG and a dispersive loop used to achieve a tunable time delay were also discussed. Then, signal processing functions that can be realized by an LCFBG-based DDL or a dispersive loop were introduced, including time reversal, pulse compression and temporal convolution that can be implemented with a single LCFBG as a DDL, and time-stretched sampling, large TBWP waveform generation and true-time delay beamforming that requires the use of a dispersive loop for a large dispersion coefficient.

CHAPTER 3 MICROWAVE TIME REVERSAL

Time reversal is one of the basic signal processing functions that can find numerous applications, such as increasing the resolution of an acoustic or microwave imaging system, solving the multipath problem in a microwave communication system, cancer detection and treatment. Current time reversal modules are usually implemented with digital electronics, which is limited to a bandwidth of a few GHz and cannot meet the requirement of broadband microwave imaging and communication systems

In this chapter, we propose and experimentally demonstrate a novel technique to achieve broadband and precise microwave time reversal using a single LCFBG as an optical DDL. In the proposed system, the time reversal is realized by the LCFBG that is operating in conjunction with a polarization beam splitter (PBS) to enable a triple use of the LCFBG with the microwave waveform entering the LCFBG from either the long or the short wavelength end. Since the LCFBG has a wide bandwidth and is used three times with exactly complementary and identical dispersion, broadband and precise microwave time reversal is ensured. A theoretical analysis is performed which is validated by simulations and an experiment. The time reversal of three different microwave waveforms with a bandwidth over 4 GHz and a time duration of about 10 ns is demonstrated.

3.1 Operation Principle

In this section, the operation principle of the proposed time reversal system is investigated. A mathematical model is developed to quantitatively describe the waveform distortion and bandwidth limit of the system.

3.1.1 System architecture

Fig. 3.1 shows the proposed microwave photonic system for broadband and precise microwave time reversal. A transform-limited optical pulse generated by a mode-lock laser (MLL) is filtered by a bandpass optical filter (OF) with a bandwidth of 4 nm, and sent to the LCFBG from its long wavelength end via a 3-port optical circulator (OC1). The optical pulse is then temporally stretched by the LCFBG. The LCFBG has a reflectivity of over 95% and a bandwidth of 4 nm which is equal to the spectral width of the optical pulse from the OF. Hence, the optical pulse from the OF is almost completely reflected by the LCFBG and the transmission is small and negligible. At the third port of OC1, a Mach-Zehnder modulator (MZM) is connected, to which a microwave waveform to be temporally reversed is applied. A polarization controller (PC1) is incorporated between OC1 and the MZM to align the polarization state of the optical pulse to the principal axis of the MZM, to minimize the polarization-dependent loss. At the output of the MZM, the optical pulse is directed into a 4-port optical circulator (OC2). A PBS is used to connect the short wavelength end of the LCFBG to the second and third ports of OC2. Two other PCs (PC2 and PC3) are employed between the PBS and OC2 to control the polarization directions of the light waves to the PBS, so that the light waves can be efficiently coupled to the LCFBG by the PBS. The optical pulse injected to the 1st port of OC2 is directed

to the second port, and then sent to the short wavelength end of the LCFBG through the PBS. The optical pulse is then dispersed by the LCFBG and returned to the PBS. Since there is no Faraday Effect involved in this process, the return light should have the polarization that perfectly matches the polarization of the lower arm of the PBS. Hence, the pulse is completely reflected to the second port of OC2. At the third port of OC2, an identical process occurs and the pulse is dispersed again at the short wavelength end of the LCFBG. The joint operation of OC2 and the PBS allows the optical pulse from the MZM to be independently and temporally dispersed by the LCFBG twice. The optical pulse is finally detected by a PD connected to the fourth port of OC2 and a time reversed microwave waveform is obtained at the output of the PD, which is monitored by a real-time oscilloscope.

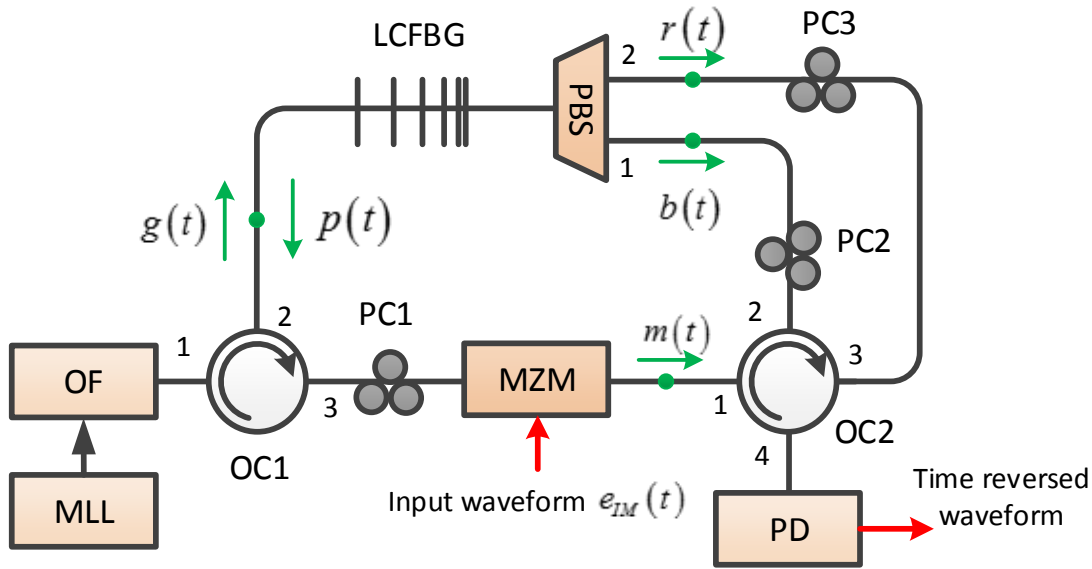


Fig. 3.1 Schematic of the proposed microwave time reversal system. MLL: mode-locked laser; OF: optical filter; LCFBG: linearly chirped fiber Bragg grating; PC: polarization controller; MZM: Mach-Zehnder modulator; PBS: polarization beam splitter; PD: photodetector; OC: optical circulator.

3.1.2 Time reversal modeling

Mathematically, the LCFBG can be modeled as a linear and time-invariant (LTI) system with a quadratic phase response and unity amplitude response. Assuming that the electrical field of the optical pulse from the OF is $g(t)$, after being dispersed by the LCFBG (entering from the long wavelength end), the electrical field of the optical pulse at the third port of OC1 is given by [70]

$$p(t) = g(t) * \exp\left(j \frac{t^2}{2\ddot{\Phi}}\right) \quad (3-1)$$

where $\ddot{\Phi}$ is the dispersion coefficient of the LCFBG looking into it from the long wavelength end, and $*$ denote the convolution operation.

Note that the dispersion coefficient looking into the LCFBG from the short wavelength end is $-\ddot{\Phi}$. Fig. 3.2 shows the reflection spectrum of an LCFBG fabricated based on the holographic method in a SMF with a length of 1 meter. The spectrum is measured by an optical vector analyzer (OVA, Luna Technologies). As can be seen the LCFBG has a bandwidth of 4 nm and a central wavelength of 1551.4 nm, which is approximately equal to the central wavelength of the pulse spectrum from the OF. The group delay responses of the LCFBG measured from its short and long wavelength ends are also shown in Fig. 3.2. As can be seen that the group delay responses are complementary and the dispersion coefficients are +2500 ps/nm and -2500 ps/nm, corresponding to the green and red lines, respectively. Since the LCFBG has a long length, which is more than the measurable length of the OVA, in the measurement the LCFBG is considered as four cascaded sub-sections, and each sub-section is measured independently. The

spectrum of the entire LCFBG is then achieved by synthesizing the four measurements corresponding to the spectra of the four sub-sections.

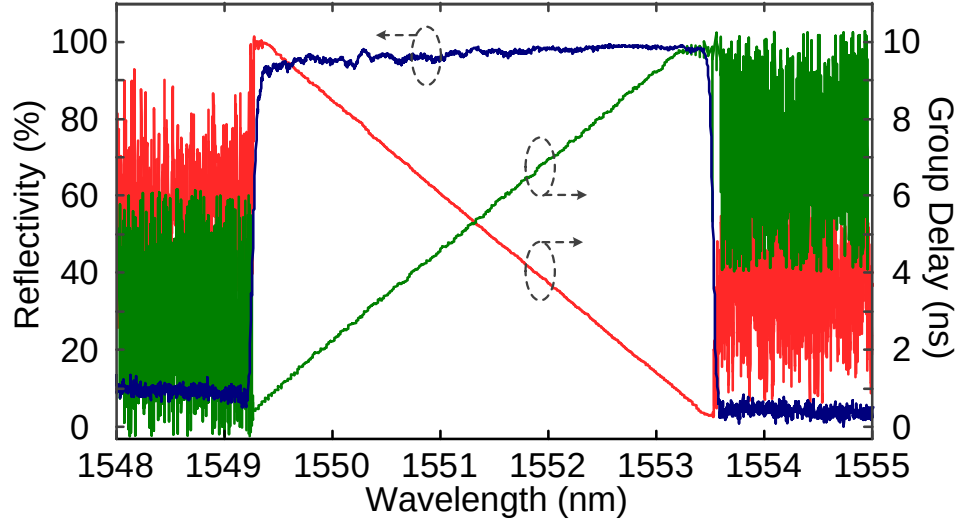


Fig. 3.2 The reflection spectrum and group delay responses of the LCFBG.

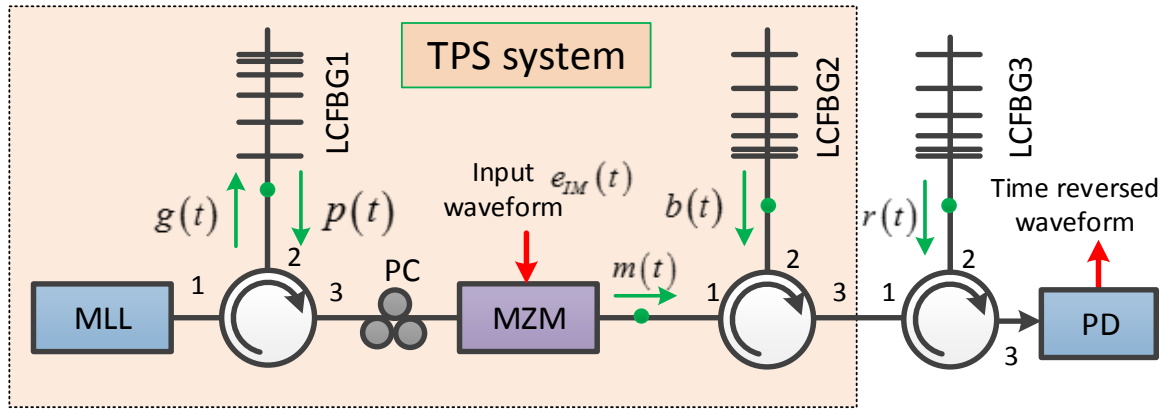


Fig. 3.3 The implementation of the proposed microwave time reversal system using three LCFBGs.

The microwave time reversal system can be modeled as a temporal pulse shaping (TPS) system [70] with a pair of complementary dispersive elements of $\ddot{\Phi}$ and $-\ddot{\Phi}$, followed by a residual dispersion of $-\ddot{\Phi}$, as shown in Fig. 3.3. As can be seen the implementation of the system

based on our proposed configuration in Fig. 3.1 to use only a single LCFBG significantly reduces the complexity and improve the performance (with no dispersion mismatch).

The electrical field at the output of the MZM is given by

$$m(t) = p(t) \times e_{IM}(t) \quad (3-2)$$

where $e_{IM}(t)$ is the microwave signal applied to the MZM.

After being dispersed by the LCFBG entering from the short wavelength end of the LCFBG for the first time, the electrical field becomes

$$b(t) = m(t) * \exp\left(j \frac{t^2}{-2\ddot{\Phi}}\right) \quad (3-3)$$

If the duration of the MLL pulse τ_0 and the dispersion of the LCFBG $\ddot{\Phi}$ (in ps²) satisfy the far field condition $|\tau_0^2 / \ddot{\Phi}| \ll 1$, the Fraunhofer approximation can be adopted. Substituting (3-1) and (3-2) into (3-3), we have [144, 145]

$$b(t) \propto g(t) * E_{IM}(\omega) \Big|_{\omega=t/\ddot{\Phi}} \quad (3-4)$$

where $E_{IM}(\omega)$ is the Fourier transform of $e_{IM}(t)$.

The electrical field at the output of the entire time reversal system $r(t)$ is obtained by propagating the optical signal $b(t)$ from the temporal pulse shaping system through a third dispersive element with a value of residual dispersion of $-\ddot{\Phi}$ via the fourth port of OC2 [70],

$$r(t) = \exp\left(\frac{jt^2}{-2\ddot{\Phi}}\right) B(\omega) \Big|_{\omega=\frac{t}{-\ddot{\Phi}}} \quad (3-5)$$

where $B(\omega)$ is the Fourier transform of $b(t)$.

Substitute (3-4) into (3-5), we have

$$\begin{aligned} r(t) &= \exp\left(\frac{jt^2}{-2\ddot{\Phi}}\right) F\left[g(t) * E_{IM}\left(\frac{t}{\ddot{\Phi}}\right)\right] \Big|_{\omega=\frac{t}{-\ddot{\Phi}}} \\ &= \exp\left(-\frac{jt^2}{2\ddot{\Phi}}\right) G(\omega) \frac{|\ddot{\Phi}|}{2\pi} e_{IM}(\ddot{\Phi}\omega) \Big|_{\omega=\frac{t}{-\ddot{\Phi}}} \\ &= \frac{|\ddot{\Phi}|}{2\pi} \exp\left(-\frac{jt^2}{2\ddot{\Phi}}\right) G\left(-\frac{t}{\ddot{\Phi}}\right) e_{IM}(-t) \end{aligned} \quad (3-6)$$

where $G(\omega)$ is the Fourier transform of $g(t)$.

The optical pulse at the fourth port of OC2 is detected by the PD. The generated photocurrent is given by

$$I(t) = \Re |r(-t)|^2 \propto G^2\left(-\frac{t}{\ddot{\Phi}}\right) e_{IM}^2(-t) \quad (3-7)$$

where $\Re$ is the responsivity of the PD.

3.1.3 Waveform distortion

As can be seen from (7) the microwave waveform at the output of the system is a time-reversed version of the input signal except for a multiplying term $G^2(-t/\ddot{\Phi})$. Ideally, the optical

pulse from the MLL is ultra-short, and its temporally dispersed version is ultra-wide and flat, which will have small impact on the generated waveform. To study the impact of $G^2(-t/\ddot{\Phi})$ on the generated time-reversed waveform, a simulation is performed, in which the dispersive element is a real LCFBG which has a spectral response given in Fig. 3.2. An up-chirped microwave waveform shown in Fig. 3.4 (solid line) is used as the input signal $e_{IM}(t)$. The microwave waveform at the output of the PD is frequency down-chirped, which is a time-reversed version of the input waveform, except for a slowly-varying envelope due to $G(-t/\ddot{\Phi})$. In the simulation, $G(-t/\ddot{\Phi})$ is the optical spectrum of an optical pulse from the MLL after being filtered by the bandpass OF and reflected three times by the LCFBG. From Fig. 3.4 it can be seen that the output waveform has exactly the same temporal duration and shape as compared with the input microwave waveform. Note that to ease the comparison, the generated time-reversed waveform is flipped horizontally and shown on the same figure. A correlation coefficient [146] of 0.998 is achieved between the original and the time reversed signal. The very small envelope distortion is induced by $G^2(-t/\ddot{\Phi})$, which can be further suppressed by using an MLL with a flatter optical spectrum and an LCFBG with more uniform reflectivity.

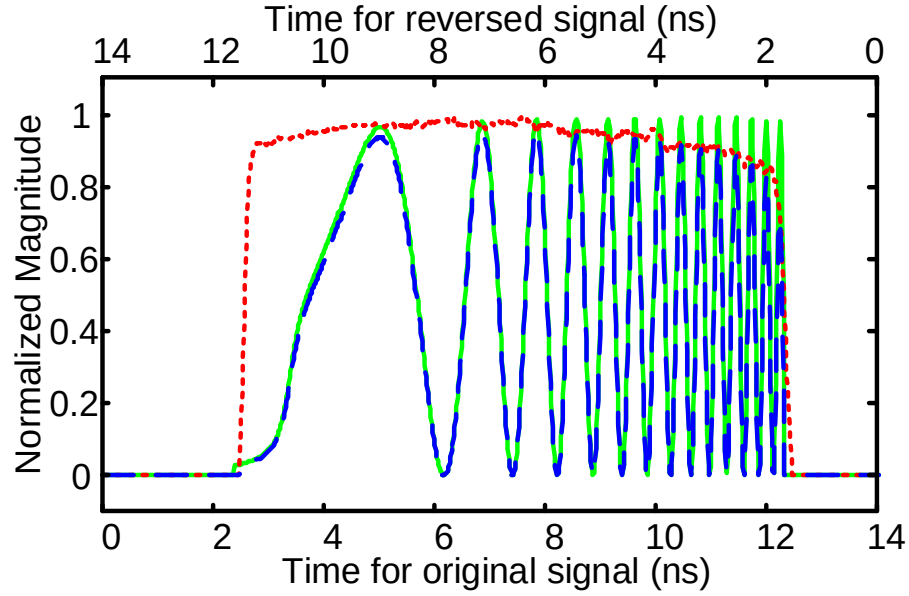


Fig. 3.4 The simulated time reversed waveform considering the impact from $G^2(-t / \ddot{\Phi})$. Dotted: input up-chirped waveform; dash: time-reversed output waveform with a frequency down-chirp; solid: the profile of $G^2(-t / \ddot{\Phi})$, determined by the spectrum of the optical pulse from the MLL and the dispersion of the LCFBG.

3.1.4 Electrical and optical bandwidth limit

The limited bandwidth is another factor that may influence the performance of the time reversal operation. In the proposed system, the bandwidth of the LCFBG is very wide, and the system bandwidth is mainly limited by the electronic components used, including the electronic amplifiers, MZM and the PD. Fig. 3.5 shows the distortion caused by the limited electronic bandwidth. Similar to Fig. 3.4, a frequency up-chirped waveform is used as the input waveform. At the output, a time-reversed waveform with a reduced amplitude is observed, especially for high frequency components, when a low-pass filter (3-dB cut-off frequency at 4 GHz) is employed to emulate the bandwidth limitation of the electronic components. The correlation coefficient between the original and the time-reversed waveform is 0.981. As the central frequency of the input signal increases, the correlation coefficient drops drastically, indicating a

largely degraded performance of the system. Again, to ease the comparison, the generated time-reversed waveform is flipped horizontally and also shown on Fig. 3.5. Since high-speed MZMs and PDs with a bandwidth up to 100 GHz or higher are now commercially available, the electronic bandwidth of the time reversal system may not be limited by the electronic components. Then, the bandwidth of the time reversal system will be determined by the optical components. Theoretically, the LCFBG is the only optical component with a finite bandwidth. Its bandwidth may limit the bandwidth of the time reversal system.

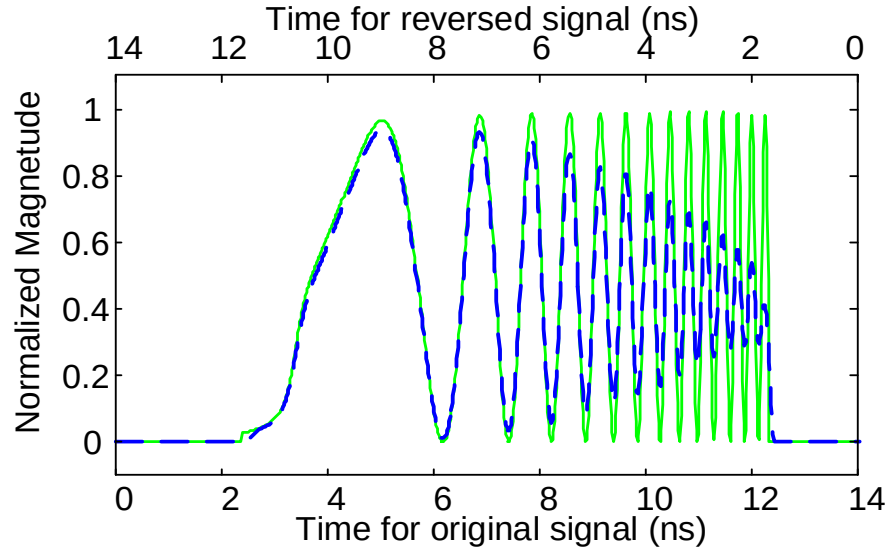


Fig. 3.5 The simulated time reversed waveform when the limited bandwidth of the electronic components is considered. Solid: input chirped signal; dash: output time-reversed signal for a limited electronic bandwidth of 4 GHz.

For an intensity-modulation and direct-detection (IM/DD) system, when a microwave signal is modulated on an optical carrier, first-order optical sidebands will be generated. The beating between the optical carrier and the sidebands at a PD will recover the modulation microwave waveform. When a temporally stretched MLL pulse is used as an optical carrier, however, the modulation process will broaden the optical spectrum. When the broadened

spectrum exceeds the bandwidth of the LCFBG, the microwave power detected at the PD will decrease since some of the spectral components will not be reflected, resulting in a limited bandwidth of the time reversal system.

Fig. 3.6 illustrates the impact of the limited bandwidth of the LCFBG on the microwave detection. First, we assume a single-frequency optical carrier at ω_c that is modulated by a microwave signal at a relatively low frequency Ω_1 . Due to the low microwave frequency, the two sidebands are within the reflection band of the LCFBG, as shown in Fig. 3.6(a). Both of them will be reflected and beat with the optical carrier at the PD, thus a maximum microwave power is achieved. When the microwave signal is increased to a higher frequency Ω_2 , only one sideband is within the LCFBG reflection band, the power of the microwave beat signal at the PD will be reduced by half. Finally, as the modulation frequency is increased to Ω_3 , no sidebands fall within the LCFBG reflection band, thus no microwave signal will be detected. Therefore, the frequency response of the system corresponds to a low pass filter with its frequency response shown in Fig. 3.6(b).

We then quantitatively calculate the bandwidth of the system due to the finite bandwidth of the LCFBG. Since intensity modulation is used, the detected microwave power should be the summation of the powers of the beat signals between the optical carrier and its two first-order sidebands,

$$P(\omega_c, \Omega) \propto R(\omega_c) \cdot R(\omega_c + \Omega) + R(\omega_c) \cdot R(\omega_c - \Omega) \quad (3-8)$$

where $R(\omega)$ is the reflectivity of the LCFBG and $P(\omega_c, \Omega)$ represents the optical power generated by an optical carrier with an angular frequency of ω_c and a unit spectral width. Since all the spectral components of the optical pulse from the MLL over a frequency range of $\delta\omega$ contribute to the optical carrier, the total microwave power detected at the PD is

$$P(\Omega) = \int_{\delta\omega} P(\omega_c, \Omega) d\omega_c \quad (3-9)$$

Based on (3-9) and using the measured LCFBG reflection spectrum shown in Fig. 3.2, the microwave spectral response of the time reversal system is calculated. As shown in Fig. 3.7, the system is a low-pass filter with the 3-dB cut-off frequency at 273 GHz, which is approximately equal to half of the optical bandwidth of the LCFBG.

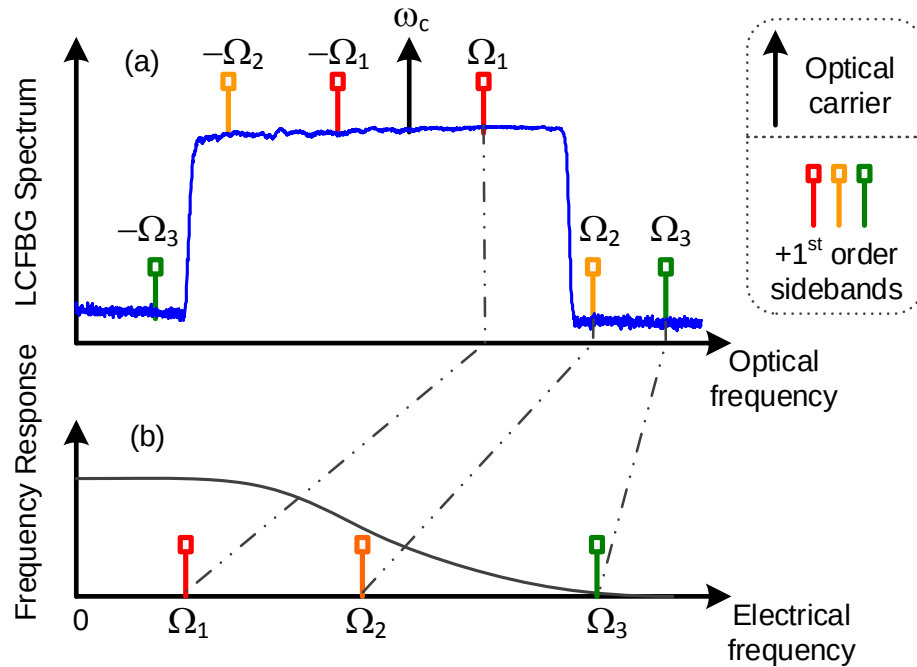


Fig. 3.6 The mechanism for the bandwidth limit of the optical part. (a) Optical carrier ω_c and sidebands reflected by the LCFBG. As modulation frequency increases from Ω_1 to Ω_3 , the sidebands may locate outside the reflection band of LCFBG; (b) the corresponding frequency response of the LCFBG.

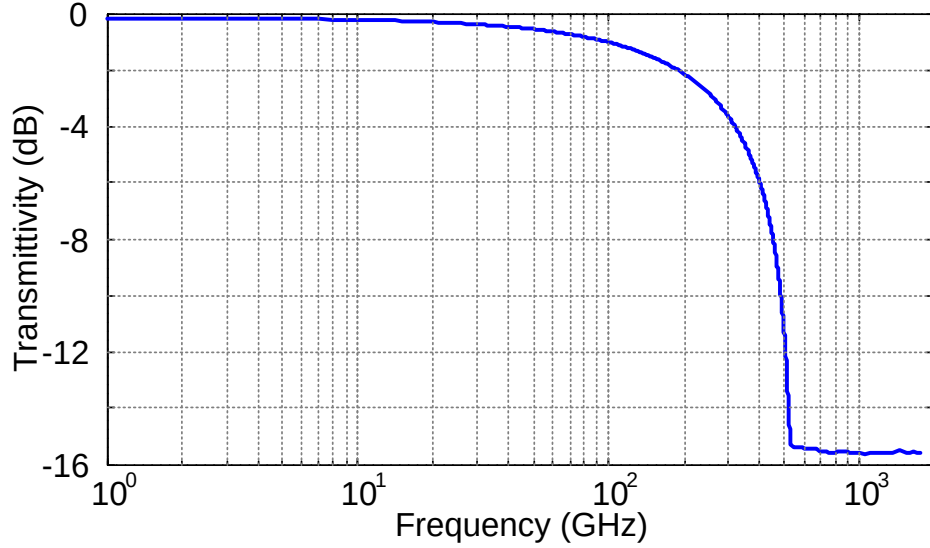


Fig. 3.7 Microwave spectral response of the time reversal system due to the finite bandwidth of the LCFBG.

3.2 Experimental Implementation

An experiment based on the setup shown in Fig. 3.1 is performed. An optical pulse from the wavelength tunable MLL (PriTel FFL-1550-20) with a 3-dB bandwidth of 8 nm and a pulse width less than 600 fs is used as the light source. The repetition rate of the pulse train from the MLL is 20 MHz. An OF with a bandwidth of 4 nm centered at the spectrum of the MLL is employed to achieve a flat spectrum and, at the same time, to ensure that the pulse can be completely reflected by the LCFBG. The LCFBG was fabricated using the holographic method. A microwave AWG (Tektronix AWG7102) with a sampling rate of 10 Gb/s is used to generate a microwave waveform that is applied to the MZM (JDSU OC-192, bandwidth of 10 GHz) after amplified by an electrical amplifier (MTC5515, bandwidth of 10 GHz). The microwave waveform and the optical pulse from the MLL are synchronized by applying a trigger signal from the MLL to the arbitrary waveform generator. The optical pulse at the fourth port of OC2 is

a time-reversed optical waveform as compared with the optical waveform at the output of the MZM. The time-reversed optical waveform is applied to the PD (New Focus 1414, 25 GHz). The detected waveform is monitored by a 32-GHz real-time oscilloscope (Agilent 93204A). A photograph of the experimental setup is shown in Fig. 3.8.

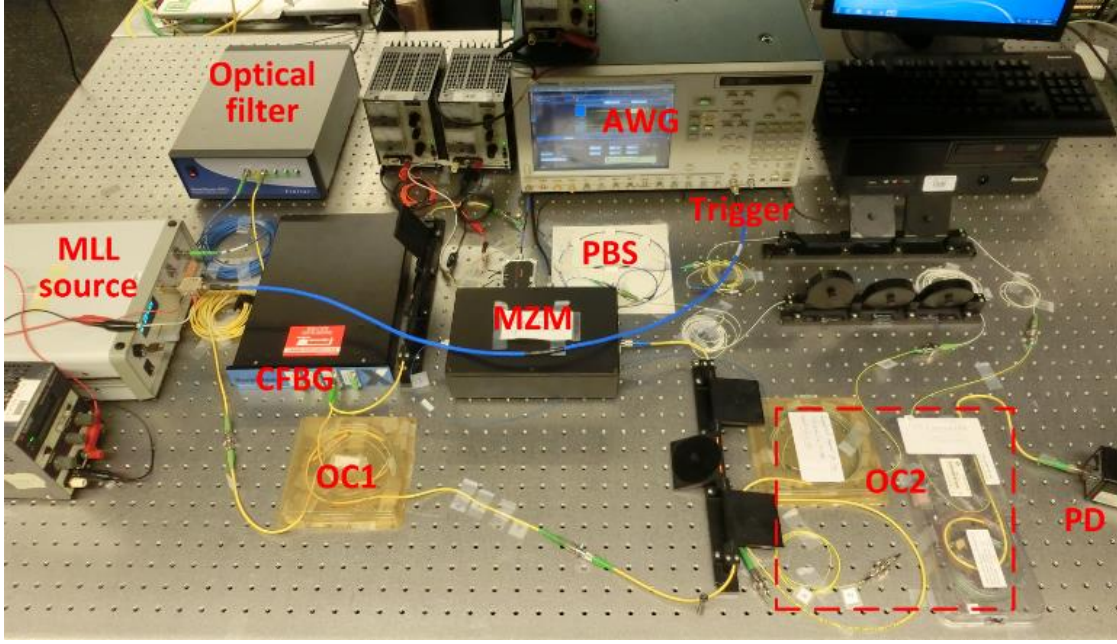


Fig. 3.8 Photograph of the experiment setup. Two 3-port circulators are cascaded to function as a 4-port circulator OC2.

The key device to achieve the time reversal is the LCFBG, which is fabricated in a single-mode fiber with a length of 1 meter. The reflection spectrum and the group delay responses are shown in Fig. 3.2.

It can be calculated that the optical pulse from the OF is stretched to have a time duration of $\tau = \ddot{\Phi} \times \Delta\lambda = 10 \text{ ns}$ at the input of the MZM. Hence, the time duration of the input microwave signal should be limited to 10 ns in order to be carried by the temporally dispersed optical pulse. The time duration of the pulse at the output of the OF is estimated to be 0.88 ps, which satisfies

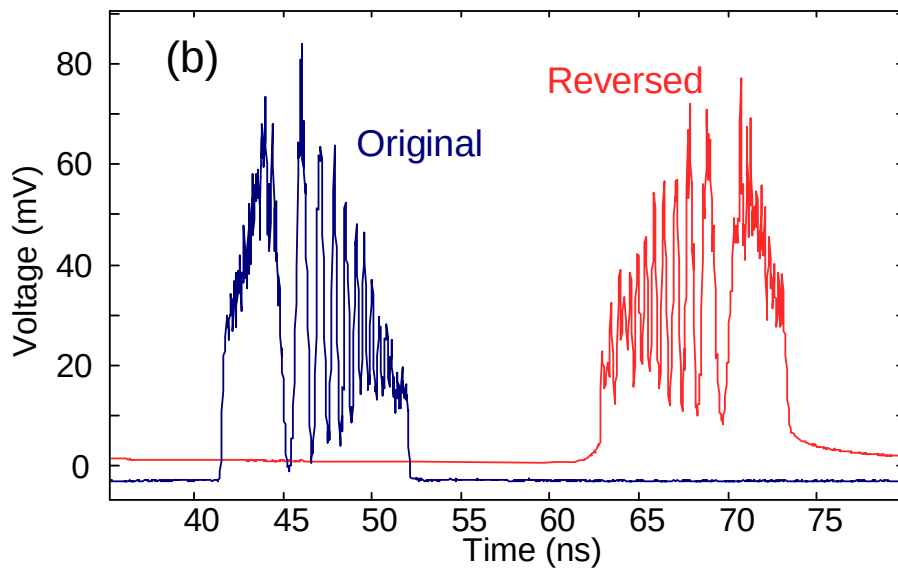
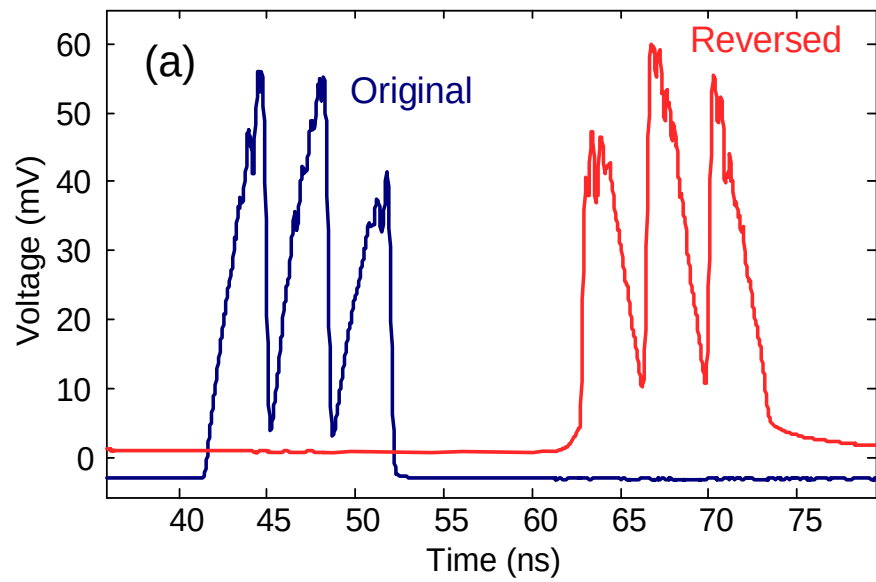
the far-field condition for a TPS system [70]. Note that the far-field condition does not need to be considered when the pulse passing through the LCFBG for the second and third times [41].

3.3 Performance Evaluation

Three different waveforms are generated by the arbitrary waveform generator to test the operation of the proposed microwave time reversal system. The three waveforms are a sawtooth wave, a chirped wave, and an arbitrary waveform. To compare an original waveform and a time-reversed waveform simultaneously, a 3-dB coupler was used after the MZM to direct part of the modulated optical pulse to a PD and sampled by another channel of the real-time oscilloscope.

Fig. 3.9 shows the microwave waveforms from the two separate channels of the real-time oscilloscope, which correspond to the waveforms before and after the time reversal. Specifically, in Fig. 3.9(a), a 3-cycle up-ramp sawtooth is time reversed to become a down-ramp sawtooth. A small amplitude change in the 3-cycles can be observed, which is caused by the non-ideally flat spectrum shape of the MLL pulse, as confirmed by the simulation. The amplitude change can be reduced by improving the flatness of the optical pulse. In Fig. 3.9(b), a frequency up-chirped microwave waveform with a time duration of 10 ns and a frequency range from DC to 4 GHz is time reversed to become a frequency down-chirped waveform. Note that the input microwave waveform is not an ideal frequency-chirped pulse due to the limited sampling rate (10 Gb/s) of the arbitrary waveform generator, and the limited bandwidth of the electrical amplifier (EA) and the MZM. In Fig. 3.9(c), an arbitrary waveform is generated by the arbitrary waveform generator which is also time reversed. It can be seen from Fig. 3.9(a)-(c) that the time-reversed waveforms have exactly the same time duration and the same shape with the original waveforms, except for

very small amplitude distortions caused by the limited bandwidth of the electronic components and the non-flat spectral shape of the optical pulse. By correlating the original waveforms with a flipped version of the time reversed waveforms, we have three correlation coefficients of 0.930, 0.939 and 0.951, which are slightly smaller than the theoretical values of 1 due to the existence of system noise. Nevertheless, precise and single shot time reversal of a microwave waveform with a bandwidth up to 4 GHz and a time duration of 10 ns has been achieved.



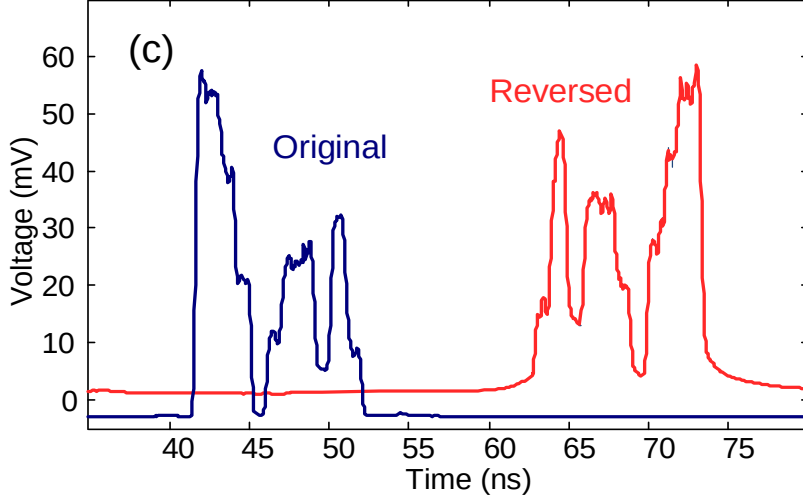


Fig. 3.9 Comparison between the original and the time reversed waveforms. (a) sawtooth wave; (b) chirped wave; (c) arbitrary waveform. The corresponding correlation coefficients are calculated to be 0.930, 0.939, 0.951.

3.4 Conclusion

We have proposed and experimentally demonstrated a novel technique to achieve broadband and precise real-time microwave time reversal using a single LCFBG. The key advantage of the proposed technique was the use of only a single LCFBG, which was used three times, thus the system was greatly simplified. More importantly, the triple use of the LCFBG enabled the complete elimination of the dispersion mismatch existing in a time-reversal system using three independent dispersive elements. The proposed technique was studied theoretically and validated by an experiment. The time reversal of three different microwave waveforms with a bandwidth of 4 GHz and a time duration of about 10 ns was demonstrated. To further increase the time duration, a dispersive element with a greater time delay is needed, for example, a longer LCFBG, or a dispersive filter near atomic resonance in rare earth ion-doped crystals.

CHAPTER 4 ARBITRARY WAVEFORM GENERATION AND PULSE COMPRESSION

In modern radar systems, high frequency and large bandwidth signal generation and processing are of great importance to achieve a high spatial resolution. To achieve a large bandwidth, phase-coded or frequency chirped signal are generally used. The signal is transmitted into free space and reflected by a target. At the receiver, a matched filter is usually required to extract the signal that is immersed in noise. The generation and detection of the signals can be realized using both analog and digital electronic systems, but with a limited operation bandwidth due to the speed of electronic system.

In this Chapter, we present an approach for simultaneous generation and compression of a microwave waveform based on an MPF. The pulse compression involves two operations, spread-spectrum microwave waveform generation at a transmitter and matched filtering at a receiver. Assume a radiated microwave waveform is $x(t)$ and its Fourier transform is $X(\omega)$, a matched filter to compress this waveform should have a spectral response given by $X^*(\omega)$, which is a complex conjugate version of the spectrum of the radiated waveform, or an impulse response $x(-t)$, which is a time reversed version of the radiated signal. Based on the convolution commutative property, if the radiated sign is time reversed, $x(-t)$, the impulse response of the matched filter should be $x(t)$. As can be seen, to achieve pulse compression, we may first generate a time reversed microwave waveform $x(-t)$, and the pulse compression can be done by passing the received time-reversed signal through a matched filter with an impulse response $x(t)$. Based on this concept, a microwave photonic signal processor to achieve spread-spectrum arbitrary microwave waveform generation and pulse compression is proposed and demonstrated.

4.1 Operation Principle

Fig. 4.1 shows the schematic diagram of the proposed microwave photonic signal processor for spread-spectrum microwave waveform generation and pulse compression. The signal processor consists of an MPF and a TRM. A light wave from a broadband optical source (BOS) is sent to a fiber-optic MZI, with the two arms connected by two 3-dB optical couplers (C1 and C2). A waveshaper (WS) as a programmable optical filter is incorporated in the upper arm to change the transmission spectrum of the MZI by applying a phase coding signal to the light wave travelling in the upper arm and an optical tunable delay line (TDL) is incorporated in the lower arm to adjust the length difference between two arms of the MZI. An MZM is connected at the output of the MZI at which the optical carrier is modulated by an ultra-short microwave pulse for waveform generation or by a received microwave waveform for pulse compression. The optical signal from the MZM is reflected by an LCFBG (LCFBG1) via an OC (OC1) and sent through a 2×2 switch to a TRM. When the switch is in the cross state, the processor is configured for waveform generation. When the switch is in the bar state, the processor is configured for pulse compression. The setup can be considered as an MPF when the switch is in the bar state. The spectral response of the MPF can be reconfigured by applying a phase coding signal to the WS, to make the MPF operate as a reconfigurable matched filter for a pre-defined microwave signal.

First, we investigate the generation of an LCMW in which the system is operating as an MPF and a TRM. To generate an LCMW, the MPF is configured to have a group delay response with a linearly increasing time delay. If an ultra-short microwave pulse is applied to the MPF, an LCMW will be generated. The chirp rate of the generated LCMW is determined by the group

delay response of the MPF. As shown in Fig. 4.1, an ultra-short microwave pulse generated by an electrical pulse generator (PG) is applied to the MZM via a 2×1 microwave combiner (MC). The receiving antenna is also connected to the MZM via the 2×1 MC. To generate an LCMW that can be compressed by the MPF, the waveform should have a spectral response that is a complex conjugate version of the spectral response of the MPF, which is done by passing the waveform through a TRM. In this case, the optical switch is in the cross state. The redirected signal is amplified by an erbium-doped fiber amplifier (EDFA) and sent to port 1 of a 4-port OC (OC2). The second and third ports of OC2 are connected to a polarization beam combiner (PBC), at the output of which LCFBG2 is incorporated. This configuration allows the light wave from port 1 of OC2 to be reflected by LCFBG2 twice, and when the dispersion coefficient of LCFBG2 is opposite to that of LCFBG1, the optical signal carrying the impulse response of the MPF will be temporally reversed [101]. Two PCs (PC1 and PC2) are employed between the second and third ports of OC2 and the PBC to ensure a maximum coupling efficiency to LCFBG2. The optical signal is finally detected by PD2 to generate a microwave waveform, which is a time reversed version of the impulse response of the MPF. The microwave signal can then be amplified, sent to an antenna Tx and radiated to the free space. After being reflected by a target, the waveform will be received by an antenna Rx and compressed by the MPF, which functions as a matched filter. It should be noted that, in a radar system, the transmitter and receiver share one antenna, which can be realized by a duplexer switch in the system.

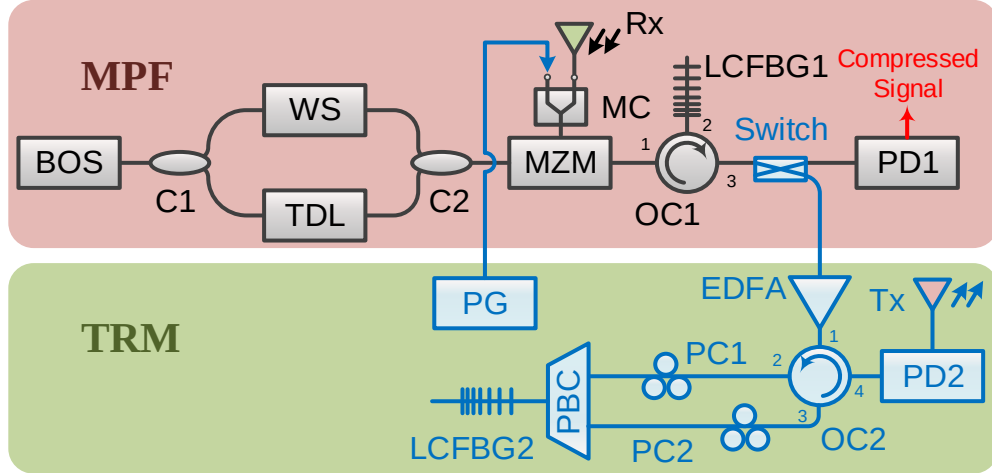


Fig. 4.1 Schematic diagram of the microwave photonic signal processor. MPF: microwave photonic filter; TRM: time reversal module; BOS: broadband optical source; C1, C2: 3-dB optical couplers; WS: waveshaper; TDL: tunable delay line; MZM: Mach-Zehnder modulator; Rx: receiving antenna; MC: microwave combiner; OC: optical circulator; LCFBG: linearly chirped fiber Bragg grating; PD: photodetector; EDFA: erbium doped fiber amplifier; PC: polarization controller; PBC: polarization beam combiner; PG: pulse generator; Tx: transmitting antenna.

4.2 Theoretical Analysis

Assume that the BOS has a broadband flat spectrum with a unity magnitude, the optical spectrum at the output of the MZI can then be denoted as $s(\omega)$, where $s(\omega)$ is also the frequency response of the MZI and ω is the optical angular frequency. Note that $s(\omega)$ can be seen as the spectrum of the optical carrier for the microwave signal modulated at the MZM. For the microwave signals modulating on different optical angular frequencies ω , different time delays will be resulted when detected at PD1 due to the dispersion of LCFBG1. The signal at the output of PD1 should be the summation of all the time delayed signals carried by all the optical carrier frequencies. First, we consider a microwave signal $e(t) = \exp(j2\pi\Omega t)$ with an angular frequency of Ω , the signal at the output of PD1 can be written as

$$\begin{aligned}
y(t) &= \int_0^{+\infty} \exp[j2\pi\Omega(t - \beta\omega)] s(\omega) d\omega \\
&= \exp(j2\pi\Omega t) \int_0^{+\infty} s(\omega) \exp(-j2\pi\Omega\beta\omega) d\omega
\end{aligned} \tag{4-1}$$

where $\beta\omega$ is a carrier frequency dependent time delay induced by LCFBG1 and β (in ps²) is the dispersion coefficient of LCFBG1. Here, the beat signals between the optical carriers are ignored, as they are not phase-correlated. Note that in (4-1), $\exp(j2\pi\Omega t)$ is the input microwave signal, the integration is time-independent and thus is the response of the system to the input microwave signal. The frequency response of the MPF is then given by

$$\begin{aligned}
H(\Omega) &= \int_0^{+\infty} s(\omega) \exp(-j2\pi\Omega\beta\omega) d\omega \\
&= S(\xi) \Big|_{\xi=\beta\Omega}
\end{aligned} \tag{4-2}$$

where $S(\xi)$ is the Fourier transform of $s(\omega)$. Since the frequency response of the MPF is simply the Fourier transformation of the optical spectrum at the output of the MZI, we can program the WS to have a certain phase response, which would lead to a frequency response of the MPF that can be used to compress an input microwave waveform.

Here, the MPF is also used in conjunction with the TRM for the generation of an arbitrary microwave waveform. To do so, we apply a short pulse to the MZM, a microwave signal that is the impulse response of the MPF will be achieved at the output of PD1. The impulse response of the system can be derived by the inverse Fourier transformation of its frequency response, given by

$$h(t) = s\left(\frac{t}{\beta}\right) \tag{4-3}$$

According to (4-3), an electrical signal with a shape identical to the spectrum of the optical carrier will be generated at PD1 when a short pulse is applied to the MZM. The system can be seen as an SS-WTT mapping system that is commonly used for the generation of microwave arbitrary waveforms [74, 75]. For example, if the MZI has a linearly increasing or decreasing FSR, an LCMW will be generated if a short pulse is applied to the MZM.

It is known that the frequency response of a matched filter should be the complex conjugate of the spectrum of the input signal. In our system, the TRM is employed to perform complex conjugation. First, an electrical short pulse is applied to the MZM, the optical signal containing the impulse response of the MPF is directed to LCFBG2 by setting the 2×2 switch at the cross state, and reflected twice due to the use of the PBC and the 4-port OC (OC2). The dispersion coefficient of LCFBG2 is chosen to be opposite to that of LCFBG1. A time reversed version of the MPF impulse response will be obtained at the output of PD2, which can be expressed as $g(t)=h(-t)=s(-t/\beta)$ [101]. It is easy to prove that for the signal $g(t)$ that contains only real values, the Fourier transform is

$$G(\Omega) = \int_{-\infty}^{+\infty} h(-t) \exp(-j2\pi\Omega t) dt = [H(\Omega)]^* \quad (4-4)$$

As can be seen, the Fourier transform (spectrum) of $g(t)$ is complex conjugate to the frequency response of the MPF $H(\Omega)$. If $g(t)$ is a radar signal being transmitted into the free space, the MPF can be used as a matched filter for the detection and compression of the returned signal.

4.3 Experimental Evaluation

The proposed processor is experimentally evaluated. In the experiments, the BOS is a spectrally flattened amplified spontaneous emission (ASE) source using an EDFA. A WS (Finisar 4000s) is employed in the upper arm of the MZI. The MZM (JDSU OC-192) has a bandwidth of 10 GHz. A microwave arbitrary waveform (Keysight M8195A) is used to generate a 62.5-ps electrical pulse. The electrical pulse and the received microwave signal are both applied to the MZM via a microwave power combiner. The dispersion coefficients of LCFBG1 and LCFBG2 are +2500 ps/nm and -2500 ps/nm, respectively, within an identical bandwidth of 40 nm centered at 1545 nm. The two PDs, PD1 (New Focus 1414) and PD2 (New Focus 1014), are used to measure the compressed electrical waveform and generate the microwave waveform to be transmitted into the free space, respectively. For simplicity, we use a microwave cable to replace the receiving and transmitting antennas by connecting the output port of PD2 to the power combiner, of which the output is connected to the microwave port of the MZM. A real-time oscilloscope (Agilent DSO-X 93204A) is used to sample the output signal from PD1 (receiver mode) or PD2 (transmitter mode). A digital high pass filter with a cutoff frequency at 50 MHz is connected to the output of PD1, to remove the strong DC component in the compressed pulse. The sample data after the digital filtering are converted to its absolute value digitally.

We firstly configure the system to generate an LCMW. To do so, the 2×2 switch is in the cross state. The WS is configured to have a quadratic phase response centered at 1543 nm with a maximum phase of 10π , which is shown as red dotted line in Fig. 4.2. The MZI has an arm length difference of 1.7 mm. The transmission spectrum of the MZI is then measured to have a

linearly increasing FSR, as also shown in Fig. 4.2. The frequency response of the MPF is measured by a vector network analyzer (Agilent E8364A) with the switch in the bar state. Fig. 4.3 shows the frequency response of the MPF, which has passband from 1.55 to 9.22 GHz, and a group delay dispersion of -0.691 ns/GHz.

When the electrical pulse with a duration of 62.5 ps is applied to the MZM and the switch is set at the cross state, a chirped microwave waveform with a shape similar to the transmission spectrum of the MZI is generated at the output of PD2, as shown in Fig. 4.4(a), which has a frequency range from around 2 to 10 GHz and a chirp rate of 1.44 GHz/ns.

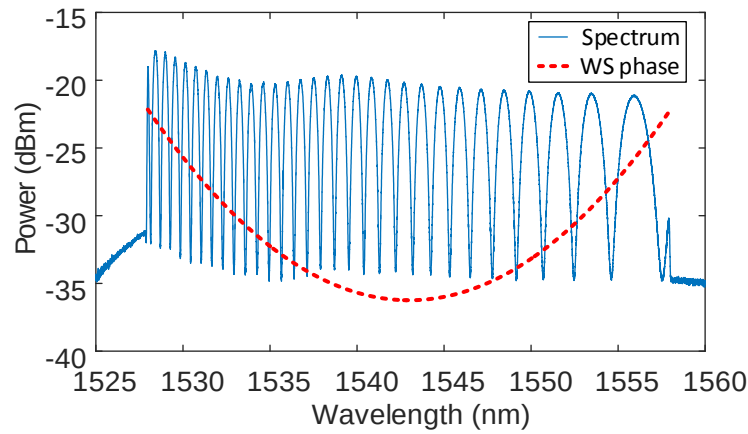


Fig. 4.2 The spectrum of the optical carrier measured at the output of the MZI when a quadratic phase is applied to the WS.

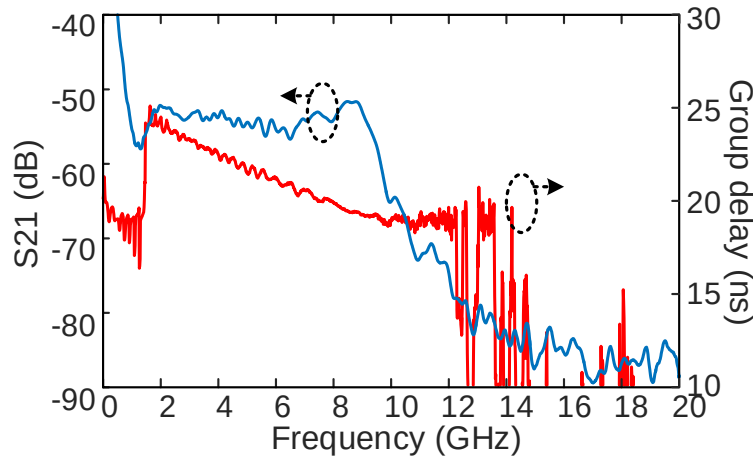


Fig. 4.3 The magnitude and group delay response of the MPF when a quadratic phase is applied to the WS.

We then configure the system to perform pulse compression. To do so, the switch is changed to the bar state and the generated chirped microwave waveform is applied to the MZM as a received signal. The chirped microwave waveform is then compressed by the matched filter. A compressed pulse is measured at the output of PD1, as shown in Fig. 4.4(b), which has a temporal width of 0.27 ns, corresponding to a compression ratio of 20.7 considering the duration of the original pulse of 5.57 ns. Theoretically, perfect matched filtering can compress the chirped pulse to a temporal width of 0.20 ns or a compression ratio of 27.9. The slightly poorer pulse compression is caused by the limited bandwidths of the electro-optic components and the measurement equipment, which makes the generated chirped microwave waveform slightly different from an ideal waveform (smaller amplitude for the high frequency components). To verify that the MPF is able to reject a microwave waveform that is different from the transmitted waveform. Here, for simplicity, a different waveform is generated by simply disconnecting the TRM, which is done by connecting PD2 directly to the output of the EDFA. In this case, the microwave waveform generated at the output of PD2 is no longer a time reversed impulse response of the MPF, but the impulse response itself $h(t)$, as shown in Fig. 4.4(c), which should not be compressed by the MPF. The signal at the output of PD1 when $h(t)$ is applied to the MZM and the switch is set at the bar state is shown in Fig. 4.4(d). No compressed pulse is observed, which confirms that the MPF is a matched filter which is able to reject a microwave signal that is different from the transmitted signal.

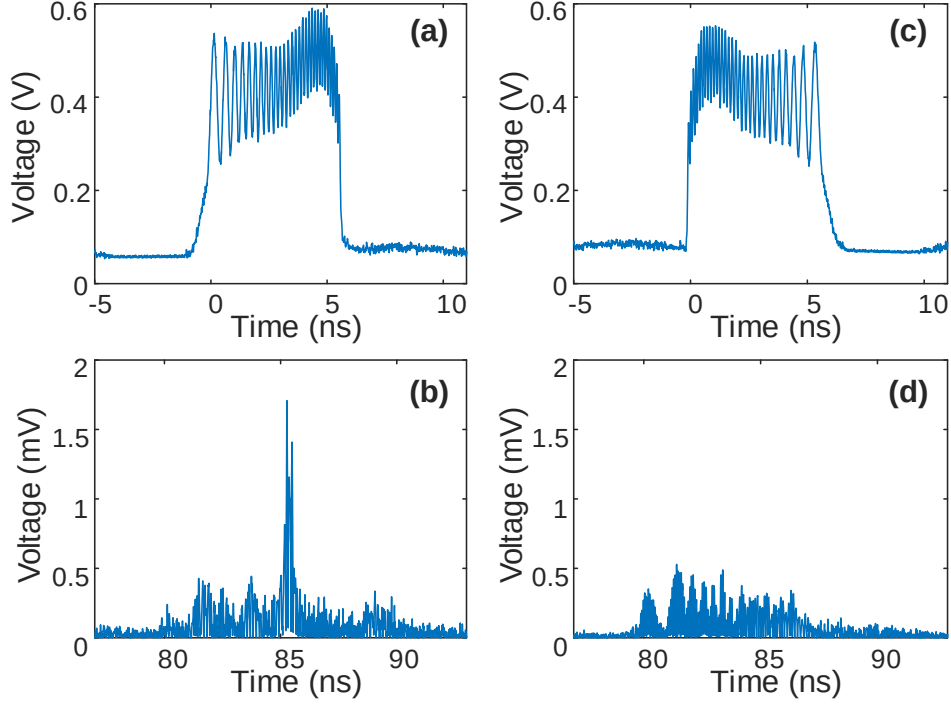


Fig. 4.4 (a) The LCMW generated at the output of PD2 with the TRM connected when a short pulse is applied to the MZM. (b) The signal at the output of PD1. The LCMW is highly compressed. (c) The LCMW at the output of PD2 with the TRM disconnected. (d) The signal at the output of PD1. No pulse compression is observed.

A significant advantage of the proposed signal processor is that it can generate and compress not only a chirped microwave waveform, but a truly arbitrary waveform by simply changing the phase response of the WS. Here we verify the operation of the system for a PCMW generation and compression. Instead of a quadratic phase, here we configure the WS to have a 7-bit Barker phase, as indicated in the red dotted line in Fig. 4.5. The MZI then has an optical transmission spectrum that corresponds to the desired PCMW (blue solid line).

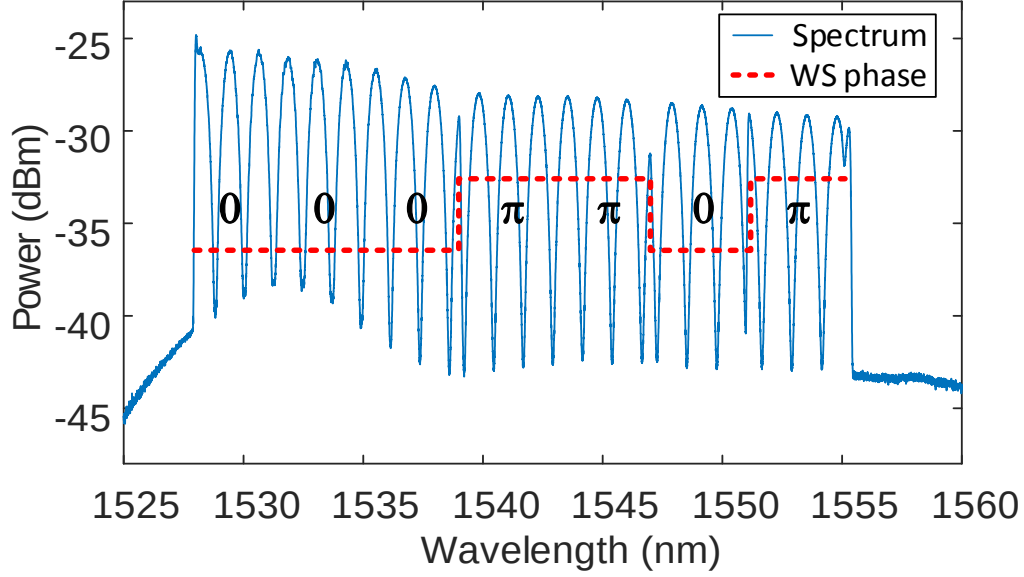


Fig. 4.5 The spectrum of the optical carrier measured at the output of the MZI when a 7-bit binary phase code is applied to the WS.

Fig. 4.6(a) shows the generated PCMW when a short electrical pulse is applied to the MZM and that the switch is set at the cross state. The signal is measured to have a carrier frequency of 4.08 GHz and a duration of 5.4 ns. The PCMW is then fed to the MZM and the switch is set at the bar state. Fig. 4.6(b) shows the compressed pulse measured at the output of PD1, in which a peak with a temporal width of 0.58 ns is observed. The compression ratio is calculated to be 9.3. Theoretically, a perfect matched filter can compress the PCMW to a temporal width of 0.42 ns or a compression ratio of 12.9. Again, the slightly poorer pulse compression is caused by the limited bandwidths of the electrical components and the measurement equipment. Similarly, here we also test the ability of the MPF to reject a signal that is different from the transmitted signal. With PD2 connected to the EDFA directly, we get a generated waveform which is shown in Fig. 4.6(c). The microwave waveform is then applied to the MZM and the switch is set at the bar state. Fig. 4.6(d) shows the measured signal at the output of PD1. No pulse compression is observed, which confirms again that the MPF is a

matched filter which is able to reject a microwave signal that is different from the transmitted signal.

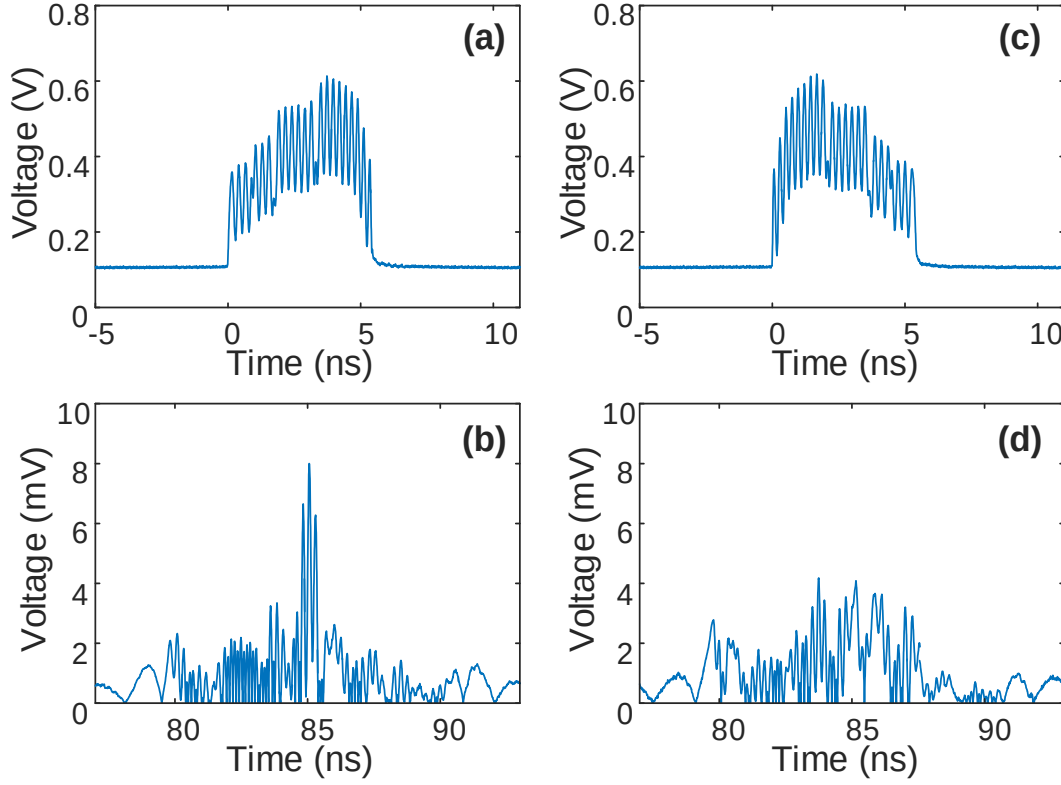


Fig. 4.6 (a) and (b): the phase-coded waveforms generated at the output of PD2 with and without time reversal when a short pulse is applied to the MZM and the switch is at cross state; (c) and (d): responses of the MPF measured at the output of PD2 when (a) and (b) is applied to the MZM, and the switch is at bar state.

5.4 Conclusion

A microwave photonic signal processor for arbitrary microwave waveform generation and pulse compression based on an MPF and a TRM was proposed and experimentally demonstrated. An arbitrary microwave waveform was generated by allowing an ultra-short microwave pulse to pass through an MPF and a TRM, to get a microwave waveform to have a spectrum that is the complex conjugate of the spectral response of the MPF. When the generated microwave

waveform was transmitted and received, by passing the received microwave waveform through the same MPF, matched filtering was performed and the microwave waveform is compressed. The proposed microwave photonic signal processor was verified by two experiments, in which a 7-bit PCMW with a carrier frequency of 4.08 GHz, and an LCMW with a bandwidth of 7.7 GHz were generated and compressed. The durations of the generated LCMW and PCMW were 5.57 and 5.4 ns, respectively. The widths of the compressed pulses were 0.27 and 0.58 ns and the pulse compression ratios were 20.6 and 9.3. The proposed microwave photonic signal processor can find applications in radar systems to generate and compress wideband and high speed microwave signals.

CHAPTER 5 TEMPORAL CONVOLUTION OF MICROWAVE SIGNALS

Temporal convolution is another approach to realizing pulse compression, it can also find other applications such as image deburring. The implementation of temporal convolution between two signals requires a combination of time reversal, time delay, signal multiplication and integration.

Based on the time reversal shown in Chapter 4, we propose and experimentally demonstrate a photonic system that can perform temporal convolution calculation of two microwave signals. The time reversal, multiplication and integration of the inputs signals are based on existing optical signal processing techniques. A changing time delay between the two input signals to be convolved is achieved by generating two sequences of replicas of the two signals with two slightly different repetition rates. The convolution results for different input signals are then derived at the output of a PD, which measures the energy of each pulse in a pulse train. Since the convolution result is recovered from the energy of the pulses, a PD and a sampling system with relatively small bandwidths will be sufficient to perform convolution between two wideband signals.

5.1 Convolution Basics

In order to detect an instance of a reflected microwave pulse, a widely used analog approach is to cross-correlate the received signal with the transmitted signal. Since the reflected signal should have a similar shape with that of the transmitted signal, this approach can essentially find out if there is a pattern that matches the transmitted waveform in the received

signal, which usually contains strong noise and interference. The cross-correlation is defined in the time domain as

$$(f \otimes g)(\tau) = \int_{-\infty}^{\infty} f(t)g(\tau+t)dt \quad (5-1)$$

which can be seen as the integration of the multiplication of two signals with a changing time shift. If we compared it to the definition of convolution

$$(f * g)(\tau) = \int_{-\infty}^{\infty} f(t)g(\tau-t)dt \quad (5-2)$$

we notice the only difference is that convolution requires a time reversal of one of the two signals. It is easy to understand, similar to convolution, cross-correlation of two signals in time domain is equivalent to multiplying the spectrum of one signal with the complex conjugate of the spectrum of the other signal. Using this concept, a microwave filter with a spectral response the same as the complex conjugate of the spectrum of the reference (transmitted) signal is usually designed and implemented on the received signal to achieve the signal cross-correlation. This process is widely known as matched filtering, which can be realized using photonics but not investigated extensively.

According to (5-2), temporal convolution can be calculated in the time domain by the following three steps:

- 1) Time reversal of one of the input signal, $g(\tau)$ for instance, to get $g(-\tau)$;
- 2) Multiplying $f(\tau)$ and a time-delayed $g(t-\tau)$;

3) Integration of the multiplication result, which gives the convolution result for a certain time delay t .

To get a full convolution result, the steps 2) to 3) should be repeated with different time delays t .

We propose and experimentally demonstrate a photonic system that can be used simultaneously as a wideband microwave waveform generator and as a matched filter for the detection of the generated microwave waveform. The matched filter is realized with a broadband optical source (BOS), an optical interferometer and a dispersive element. The wideband microwave signal is generated by temporally reversing the impulse response of the matched filter, so that the frequency response of the matched filter is always conjugate to the spectrum of the generated waveform, no matter what kind of wideband signal is generated. A linearly chirped microwave signal and phase-coded signal are used to test the system. An operation bandwidth as large as 7.7 GHz is demonstrated for a waveform with a duration of 5.57 ns.

5.2 Experimental Implementation

Fig. 5.1(a) shows microwave temporal convolution for two input microwave signals of $f(t)$ and $g(t)$, which involves three operations, microwave time reversal, multiplication and integration. The three operations can be performed using three subsystems in the optical domain, as shown in Fig. 5.1(b). The first subsystem is used for achieving time reversal, which is similar to the approach we introduced in [101], where a mode-locked laser (MLL) is employed to generate a transform-limited pulse train. An optical pulse in the pulse train is first reflected by an LCFBG (LCFBG1) through an OC (OC1), and then spectrally shaped by a programmable optical

filter (POF), to encode a microwave waveform to make the spectral response of the POF have a shape that is identical to the microwave waveform. The second subsystem is for achieving multiplication, which is implemented by simply using an MZM, to which a second microwave signal is applied via its electrical port. The integration is performed by a third subsystem that consists of a second LCFBG (LCFBG2) and a low-speed PD. Since the input signals are faster than the response time of the PD, the output of the PD is in fact proportional to the optical energy that it receives within its response time window, i.e., the integration of the power of the fast input signal. To get the integration for the amplitude of the signals as indicated in (5-2), the input signals $f(t)$ and $g(t)$ should be preprocessed to have only positive values, and then converted to $\sqrt{f(t)}$ and $\sqrt{g(t)}$. Note that if the integration is performed in the optical domain (without photo-detection), the preprocessing is not needed. In the following, for simplicity, assume we have two input signals given by $\sqrt{f(t)}$ and $\sqrt{g(t)}$.

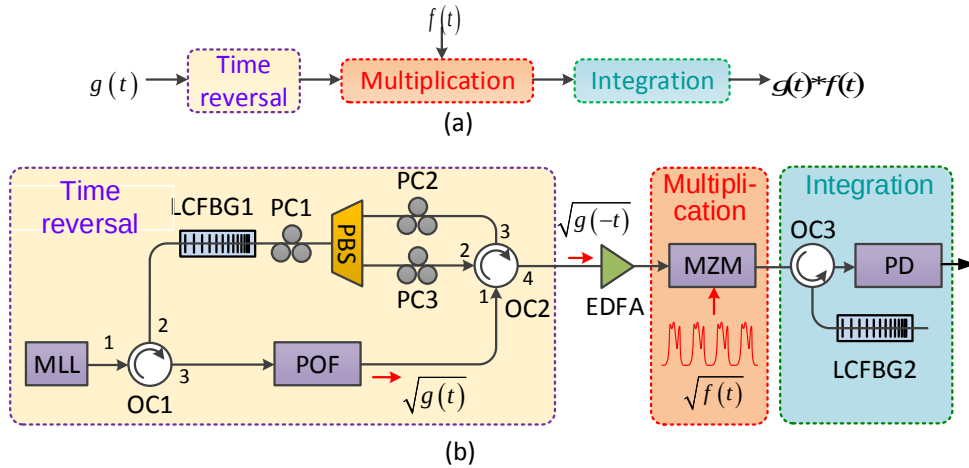


Fig. 5.1 (a) Illustration for the operation of the proposed temporal convolution system; (b) Schematic diagram of the temporal convolution system consisting of three sub-systems. MLL: mode-locked laser; OC: optical circulator; POF: programmable optical filter; LCFBG: linearly chirped fiber Bragg grating; PC: polarization controller; PBS: polarization beam splitter; EDFA: erbium-doped fiber amplifier; MZM: Mach-Zehnder modulator; PD: photodetector.

For a microwave signal $\sqrt{g(t)}$, the POF can be configured to have a spectral response that has the same shape as the microwave signal, $\sqrt{g(\omega)}$, where ω is the optical angular frequency given by $\omega = t / \ddot{\Phi}$ and $\ddot{\Phi}$ is the dispersion coefficient of LCFBG1 when looking into it from the left end [70]. The spectrum is linearly mapped to the time domain through wavelength-to-time mapping at LCFBG1 from the left end [70]. After reflected by LCFBG1 the second time from its right side, wavelength-to-time mapping is performed and a time-reversed microwave signal $\sqrt{g(-t)}$ is obtained [101]. Note that the dispersion coefficient of LCFBG1 when looking into it from the right end is $-\ddot{\Phi}$. $\sqrt{g(t)}$ can also be encoded to a pulse in the pulse train with the temporal pulse shaping approach [70], where an MZM is used instead of the POF, but the microwave signal applied to the MZM must be synchronized to the pulse in the pulse train. The two techniques are equivalent. Here we choose the wavelength-to-time mapping approach using a POF as it does not require any synchronization between the waveform from the MLL and the waveform from an AWG, thus it is simpler experimentally. Although $\sqrt{g(-t)}$ can be generated by letting the MLL pulse reflected only once by the left end of LCFBG1 after been filtered by the POF, we used the LCFBG1 three times as it is the configuration that can also perform time reversal of a signal encoded using the temporal pulse shaping approach. Three PCs are used to ensure that the pulse can be reflected by LCFBG1 via the right end twice and to achieve a maximum coupling efficiency [101]. Since the spectral response of the POF is not updated on a pulse-by-pulse basis [72], $\sqrt{g(-t)}$ is repeating at a repetition rate identical to that of the pulse train from the MLL. An EDFA is used after the time reversal subsystem to compensate for the losses of the POF, the PBS, and LCFBG1.

The amplified pulses in the pulse train encoded by $\sqrt{g(-t)}$ are then sent to the multiplication subsystem, which is simply the MZM. The second microwave signal to be convolved, $\sqrt{f(t)}$, is generated by the AWG with a repetition rate slightly different from that of the pulse train from the MLL, and is applied to the MZM. The multiplied signal at the output of the MZM is then launched into LCFBG2 for integration. LCFBG2 has a dispersion coefficient that is identical to that of LCFBG1 when looking into it from the left end. The signal at the output of LCFBG2 is converted to the electrical domain at the PD. Integration will be performed at the same time thanks to the small bandwidth of the PD.

To illustrate the operation of the system, the convolution between a rectangular waveform $f(t)$ and an inverse sawtooth waveform $g(t)$ is used as an example, as shown in Fig. 5.2. First, $\sqrt{f(t)}$ and $\sqrt{g(-t)}$ are generated with a repetition rate of T_1 and T_2 , respectively. There is a slight difference of Δt between T_1 and T_2 . Due to the difference between the repetition rates of the two signals, a changing time delay difference can be achieved between the replicas of $\sqrt{f(t)}$ and $\sqrt{g(-t)}$ after a different number of periods n . The two signals are then multiplied and integrated. The output of the integration subsystem is a series of short pulses with different peak powers. The convolution result $I'(n)$ can be reconstructed from the amplitudes of the peaks. Note that $I'(n)$ is the integration of the n -th pulse in the pulse train. Therefore, it is discrete, and the corresponding unit time increase along the horizontal axis is $|T_1 - T_2|$ for the convolution result.

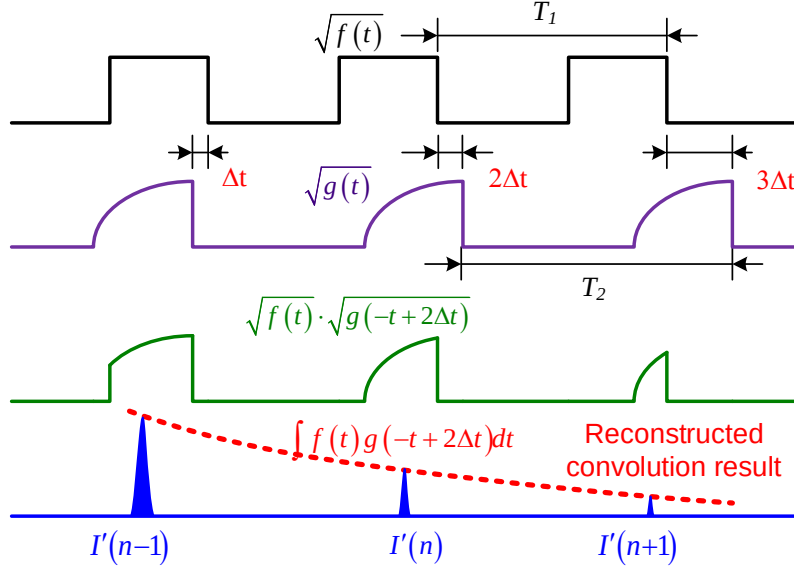


Fig. 5.2 Operation principle of the proposed temporal convolution system. A rectangular waveform $f(t)$ and a sawtooth waveform $g(t)$ are used as the two signals to be convolved.

Mathematically, a time reversed signal on an optical pulse at the output of the time reversal subsystem (port 4 of OC2) can be expressed as [101],

$$r_i(t) = \exp\left(\frac{jt^2}{2\ddot{\Phi}}\right) \sqrt{g(-t)} \quad (5-3)$$

where $\sqrt{g(t)}$ is seen as the pre-processed input microwave signal. The quadratic phase term in (5-3) is induced by LCFBG1, which will be eliminated at the PD after photo-detection. Considering that the time-reversed microwave signal is carried by the optical pulse train, we have

$$r(t) = r_i(t) * \sum_{n=0}^{N-1} \delta(t - nT_1) = \sum_{n=0}^{N-1} r_i(t - nT_1) \quad (5-4)$$

where δ is the Dirac delta function, T_1 is the period of the pulse train from the MLL, n is an integer and $N = 1/|T_1 - T_2|$. Similarly, the other pre-processed microwave signal, $\sqrt{f(t)}$, is repeating at a slightly different repetition rate with a period of T_2 . The signal applied to the MZM is expressed as

$$s(t) = \sum_{n=0}^{N-1} \sqrt{f(t - nT_2)} \quad (5-5)$$

The two signals $r(t)$ and $s(t)$ are then multiplied at the MZM. The time intervals T_1 and T_2 are chosen to be much larger than the temporal duration of the input waveforms to avoid overlapping between any two adjacent waveforms, and the difference between T_1 and T_2 is chosen to be small, so that we only need to consider the terms with the same value of n in (5-4) and (5-5) to overlap in time within the summation range of $N-1$, i.e., the multiplication will only take place for the terms with the same value of n . At the output of the MZM, the signal can be expressed as

$$m_n(t) = r_i(t - nT_1) \times \sqrt{f(t - nT_2)} \quad (5-6)$$

The time delay difference between $r_n(t)$ and $\sqrt{f(t - nT_2)}$ is $n \times (T_2 - T_1)$. As n changes, a different time delay difference between the two waveforms is resulted, which is required by the temporal convolution.

The multiplication output is then directed to LCFBG2 for the first step of the integration operation. After propagating through LCFBG2 with a dispersion coefficient of $\ddot{\Phi}$, which is

identical to that of LCFBG1 when looking into it from the left end, we obtain the output signal as a convolution between $m_n(t)$ and the impulse response of LCFBG2, given by

$$y(t) = m_n(t) * \exp\left(-\frac{jt^2}{2\ddot{\Phi}}\right) \quad (5-7)$$

By using the wavelength-to-time mapping relationship [70], we get

$$y(t) = \exp\left(-\frac{jt^2}{2\ddot{\Phi}}\right) \times F[m_n(t)] \quad (5-8)$$

where F denotes Fourier transform. The signal is then detected at the PD, which is the second step of the integration operation, generating an output current given by

$$\begin{aligned} I(t) &= \Re |y(t)|^2 \\ &= \Re \left| \exp\left(-\frac{jt^2}{2\ddot{\Phi}}\right) \times F[m_n(t)] \right|^2 \bigg|_{\omega = -\frac{t}{\ddot{\Phi}}} \\ &= \Re |F[m_n(t)]|^2 \bigg|_{\omega = -\frac{t}{\ddot{\Phi}}} \end{aligned} \quad (5-9)$$

where $\Re$ is the responsivity of the PD. It can be seen that the signal at the output of the PD is actually the power spectrum of the multiplication result in (5-6), rather than its integration. However, it should be noted if the bandwidths of the input signals are small compared to the optical carrier frequency, $I(t)$ becomes a very short optical pulse with a pulse width given by $\delta t = \delta\omega \cdot \ddot{\Phi}$, where $\delta\omega$ is the electrical bandwidth of the multiplication result in (5-6). If δt is smaller than the response time of the PD, the output current will be proportional to the energy of a pulse, which is the integration of the pulse spectrum, thus we have

$$I'(n) = \Re \int_{\delta\tau} |y(t)|^2 dt = \Re \int_{\delta\omega} |F[m_n(t)]|^2 d\omega \quad (5-10)$$

According to the Parseval's theorem, for each pulse, we have the output given by

$$I'(n) = \Re \times \frac{1}{2\pi} \int_{\delta t} |m_n(t)|^2 dt \quad (5-11)$$

Substitute (5-3) and (5-6) into (5-11), we get

$$I'(n) = \Re \times \frac{1}{2\pi} \int_{\delta t} g(-t + nT_1) f(t - nT_2) dt \quad (5-12)$$

Compare (5-12) with (5-2), $I'(n)$ can be seen as the convolution between signals $g(t)$ and $f(t)$, with a time delay difference of $n \times (T_2 - T_1)$. For a different n , the convolution result provides a value corresponding to a different time delay difference. It can be seen that the convolution process imposes strong requirement for the period of the two signals to be convolved. However, a fiber optic loop may be used to convert a pulsed signal into a periodic signal with a repetition rate determined by the loop length. Convolution can then be performed between a periodic signal and a non-periodic signal.

In the proposed system, the PD is used to measure the energies rather than the temporal shapes of the pulses, thus the required bandwidth can be much smaller than the bandwidths of the input signals. In fact, it is only required that the response time of the PD is faster than T_1 and T_2 . It is also required that the response time is slower than the duration of the pulses that arrive at the PD. If these two requirements are satisfied, integration can be realized without the use of

LCFBG2. Unfortunately, the pulse duration, which is the duration of $\sqrt{f(t)}$ and $\sqrt{g(-t)}$ without using LCFBG2, is comparable to T_1 and T_2 . A practical PD may not satisfy the response time requirement, considering that a steep slope is difficult to achieve at the cut-off frequency of a PD or even that of an electrical filter. The use of LCFBG2, which reduces the duration of the pulses that arrive at the PD, significantly increases the high frequency limit of the PD and allows the implementation of temporal convolution with a practical PD.

The preprocessing that converts the input signals to their square roots is required since the final step of integration is realized by the small-bandwidth PD, and the signal at the output of the PD, which is in the electrical domain, is proportional to the power of the input optical signal. If an all-optical integrator is implemented instead, the preprocessing will not be needed.

5.3 Experimental Evaluation

An experiment based on the system shown in Fig. 5.1 is performed. A wavelength tunable MLL (PriTel FFL-1550-20) is used as the optical source, which generates an optical pulse train with a repetition rate of 20 MHz or a period of 50 ns. The 3-dB spectral width and temporal width of a pulse in the pulse train is 8 nm and 600 fs, respectively. LCFBG1 and LCFBG2 are fabricated to have an identical bandwidth of 4 nm and a dispersion coefficient of ± 2500 ps/nm. A POF (Finisar WaveShaper 4000s) is used to encode one of the input signal to the MLL pulses. The other input signal is generated by an AWG (Tektronix AWG7102) with a sampling rate of 10 Gb/s and applied to an 10-GHz MZM (JDS-U OC-192) to perform signal multiplication. The AWG is configured to generate a waveform with a period 1% longer than that of the pulse train

from the MLL, i.e., 50.5 ns. The convolution result at the output of the PD (New Focus 1414, 25 GHz) is sampled by a real-time oscilloscope (Agilent 93204A).

The POF has a spectral resolution of 10 GHz, which can generate $\sqrt{g(t)}$ at an equivalent sampling rate of 5 Gb/s when working in conjunction with LCFBG1. Hence, both $f(t)$ and $g(t)$ have an analog bandwidth of less than 5 GHz [70]. According to (5-9), the pulse width is around 200 ps after integration, which is larger than the response time of the PD and that of the oscilloscope. To satisfy the condition given in (5-10), a digital low-pass filter with a cutoff frequency at 1 GHz is adopted for the signal sampled by the oscilloscope. In fact, it is only required that the PD has a response time faster than the period of the waveforms to be convolved, which is 50 ns in our experiment. Although the system performs convolution for two signals with relatively large bandwidths, only a low-speed PD and a low-speed sampling system are required to acquire the convolution result, which can be a great advantage for the proposed system. It should be noted that, each MLL pulse is temporally stretched to have a duration of 10 ns by LCFBG1, indicating that the system can only process an input signal with a temporal duration less than 10 ns.

Then, we use different waveform pairs to test the operation of the proposed temporal convolution system. The waveform pairs include two rectangular waveforms, a rectangular waveform and an inverse sawtooth waveform, and an arbitrary waveform and a short pulse. Fig. 5.3 shows the two rectangular waveforms with temporal widths of 10 ns that are generated by the POF and the AWG, respectively. Although a rectangular $g(\omega)$ is applied to the POF, some ripples can be found in the generated waveform shown in Fig. 5.3(a) due to the uneven optical spectrum of the MLL pulse and the uneven gain spectrum of the EDFA. The rectangular

waveform generated by the AWG is very close to an ideal rectangular waveform. It is known that the convolution of two rectangular waveforms with an identical temporal width is a triangular waveform, and the rise time of the triangular waveform should be equal to the width of one of the input rectangular waveform. Fig. 5.4 shows the experimentally obtained convolution output (blue line). An ideal convolution result (red-dotted line) is also shown for comparison. The output signal is a series of short pulses, with the peak amplitude profile nicely fitting to the ideal convolution. It should be noted that two time scales for the horizontal axes are used in Fig. 5.4, where the lower horizontal axis represents the time for the measured output and the upper horizontal axis represents the time for the convolution, which is recovered by using $n \times (T_2 - T_1)$ with n from 0 to $N-1$. As we have discussed, the convolution results are discrete values given by the measured the energies of the pulses. The corresponding time axis should also be discrete, with a unit time increment given by $(T_2 - T_1)$. In our case, $(T_2 - T_1) = 0.01 \times T_1$. The upper horizontal axis corresponding to the convolution is simply obtained by multiplying the real time in the lower horizontal axis by 0.01.

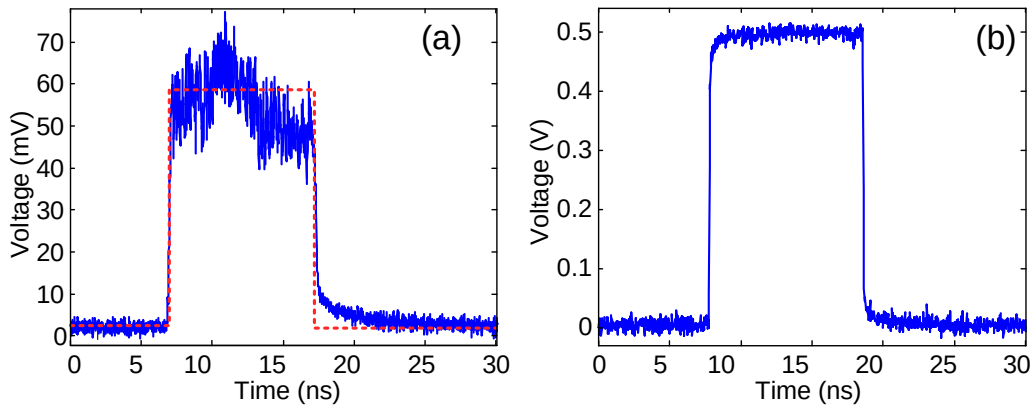


Fig. 5.3 Two rectangular waveforms used as the input waveforms for temporal convolution. (a) Square root of $g(t)$ encoded by the POF. Blue line: the measured waveform at the output of the POF; red dotted line: an ideal rectangular waveform. (b) Square root of $f(t)$ generated by the AWG.

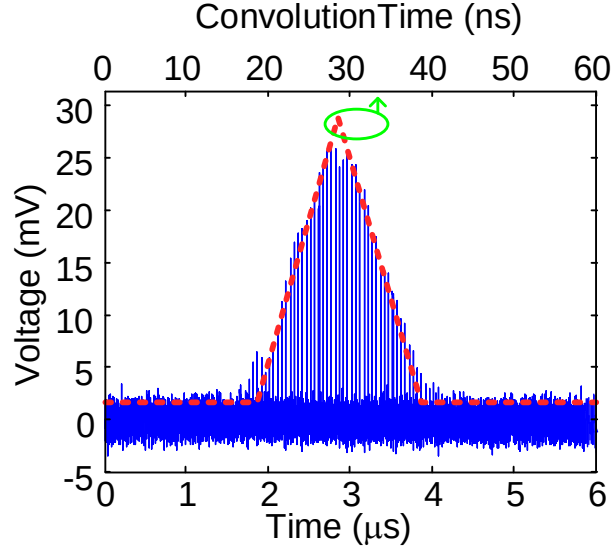


Fig. 5.4 The convolution between two rectangular waveforms. Red-dotted line: the theoretical convolution output of the two rectangular waveforms with the upper horizontal axis; blue line: the measured convolution output with the lower horizontal axis, which is a series of pulses with the peak amplitudes representing the convolution result.

An asymmetric waveform which is an inverse sawtooth waveform is then used to test the temporal convolution system. Again, by configuring the POF to have a spectral response of $\sqrt{g(\omega)}$, where $g(\omega)$ has an inverse sawtooth shape, the square root of an inverse sawtooth waveform with a temporal duration of 10 ns is obtained at the output of the time reversal subsystem, as shown in Fig. 5.5(a). The waveform is then convolved with the rectangular waveform shown in Fig. 5.3(b). Fig. 5.5(b) shows the convolution result. A good agreement between the theoretical and the measured results is achieved. For convolution operation, we know that $f * g = g * f$, i.e., no matter which function is temporally reversed, the convolution result should be the same. In our system, however, the convolution output may be temporally reversed if a different input signal is temporally reversed. But the sign of $(T_2 - T_1)$ will also be changed for $f * g$ and $g * f$. The time in the horizontal axis for convolution $n \times (T_2 - T_1)$ will then be reversed, which results in a consistent convolution results for both $f * g$ and $g * f$.

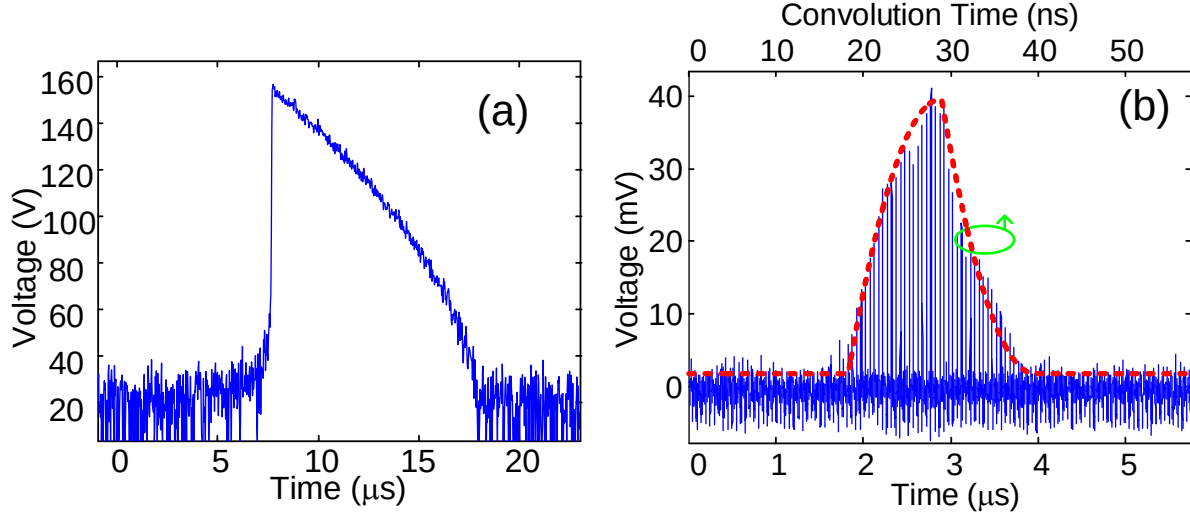


Fig. 5.5 (a) The square root of an inverse sawtooth waveform achieved at the output of the POF; (b) the convolution between a rectangular waveform and an inverse sawtooth waveform. Red dotted line: the theoretical convolution output of a rectangular waveform with an inverse sawtooth waveform, blue line: the measured convolution output of the system.

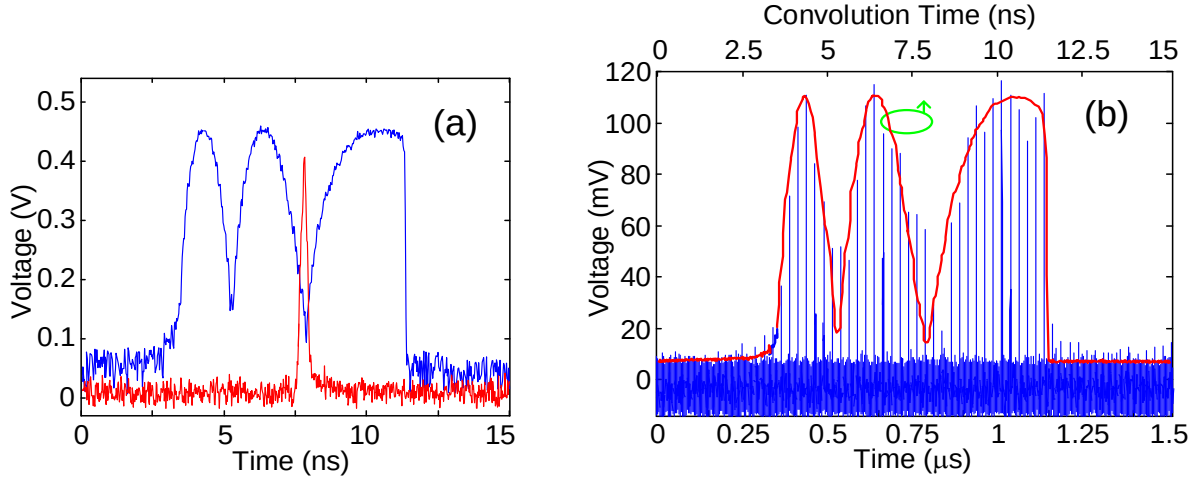


Fig. 5.6 (a) The square root of a short pulse achieved at the output of the POF (red) and the square root of a three-cycle chirped waveform generated by the AWG (blue); (b) the convolution between a three-cycle chirped waveform and a short pulse. Red line: theoretic convolution result; blue line: the output of the convolution system, when the three-cycle chirped waveform is convolved with a short pulse with a temporal width of 400 ps.

Finally, we investigate the convolution between a complex waveform and a short pulse. The complex waveform is a three-cycle chirped waveform, which is generated by the AWG. The POF is configured to have a narrow passband of 20 GHz which leads to the generation of a short

pulse with a temporal width of 400 ps after wavelength-to-time mapping by LCFBG1. The generated square root of the three-cycle chirped waveform and the short pulse are shown in Fig. 5.6(a). The convolution of a waveform and an ultra-short pulse (ideally a unit impulse function) should be the waveform itself. Fig. 5.6(b) shows the ideal convolution result and the measured convolution output of the system. Note that the vertical axis does not represent the actual voltage level of the three-cycle chirped waveform generated by the AWG, which has a peak voltage of 0.5 V (refer to Fig. 5.3(b)). Again, the measured result is in good agreement with theoretical result. For a complex waveform with more details, to get a more smooth convolution result, one may use a smaller value of $|T_2 - T_1|$, so that the convolution can be calculated with a higher time resolution.

5.4 Conclusion

We have proposed and experimentally demonstrated a photonic system that can perform temporal convolution of two microwave waveforms, which was realized by three photonic subsystems to perform the time reversal, signal multiplication, and integration. The key challenge in performing temporal convolution was to realize a variable time delay difference between the two microwave waveforms, which was achieved by generating two sequences of replicas of the two microwave waveforms with two slightly different repetition rates. The two sequences were multiplied at the MZM and integrated by LCFBG2 followed by the photo-detection at the PD, with the convolution result obtained at the output of a PD. Since the PD here is used to detect the pulse energy, a small bandwidth of the PD will be sufficient to perform the proposed temporal convolution in which the two microwave waveforms could be wideband. The

proposed approach was experimentally evaluated, in which the calculations of three temporal convolutions between two rectangular waveforms, between an inverse sawtooth waveform and a rectangular waveform, and between an arbitrary waveform and a short pulse were experimentally demonstrated.

CHAPTER 6 TIME STRETCHED SAMPLING BASED ON A DISPERSIVE LOOP

The ever-increasing bandwidth of modern microwave sensing and communications systems has led to new challenges on signal processors to operate at a very high sampling rate. Using conventional sampling techniques may not be able to meet the demand. To realize broadband sampling, the optical time-stretched sampling has been considered an effective solution. So far, the TBWP and the stretching factor of a time-stretched sampling system are mainly limited by the maximum available dispersion coefficient of the DDL used to perform time stretching.

In this Chapter, we propose a novel technique to achieve time-stretched microwave sampling with a significantly increased stretching factor. In the proposed system, a microwave waveform is modulated on a pre-dispersed optical pulse which is sent to a recirculating dispersive loop consisting of an LCFBG and an EDFA. The LCFBG is used to achieve repetitive pulse stretching and the EDFA is used to compensate for the loss in the loop. By controlling the gain of the EDFA to compensate for the loop loss, the optical waveform can recirculate in the loop and a repetitive use of the LCFBG for accumulated pulse stretching is realized. The proposed technique is experimentally demonstrated. An LCFBG with a GDD coefficient of 1500 ps/nm is fabricated and incorporated in the recirculating dispersive loop. An equivalent GDD coefficient of 12000 ps/nm is achieved, which, to the best of our knowledge, is the largest dispersion ever reported for time-stretched sampling. The corresponding stretching factor is 36. The use of the system to sample a microwave waveform is demonstrated. For a sampling system with a bandwidth of 32 GHz, the use of the proposed recirculating dispersive loop can extend the

bandwidth by 36 times or 1.15 THz (or a time resolution of 347 fs) with a frequency resolution of 4.93 GHz.

6.1 Operation Principle

The schematic of the proposed time stretched sampling system is shown in Fig. 6.1. An optical pulse from an MLL is sent to a DCF serving as a pre-dispersion element. The pre-dispersed optical pulse is then sent to an MZM through an optical bandpass filter (OBPF) and an EDFA (EDFA1). A microwave waveform is modulated on the pre-dispersed optical pulse at the MZM. The modulated signal is then sent to the recirculating dispersion loop, in which an LCFBG and a second EDFA (EDFA2) are incorporated. Note that the bandwidth of the OBPF is identical to the bandwidth of the LCFBG, so the pre-dispersed optical pulse at the output of the OBPF has a spectral width that is identical to that of the LCFBG. The microwave waveform to the MZM is generated by mixing an electrical gate signal from an AWG with a sinusoidal microwave signal from a microwave generator (SG). The modulated waveform is launched into the recirculating dispersive loop through a 2×2 3-dB coupler. In the loop, the LCFBG is used as a dispersive element and EDFA2 is used to compensate for the round-trip loss. An attenuator (Att) is also included in the loop to provide a fine control of the loop gain, to maintain a full compensation of the loss while avoiding optical lasing in the loop. The optical pulse is recirculating in the dispersive loop. The time-stretched optical pulse at the output of the loop is sent to a PD. The stretched microwave waveform is sampled by a real-time oscilloscope.

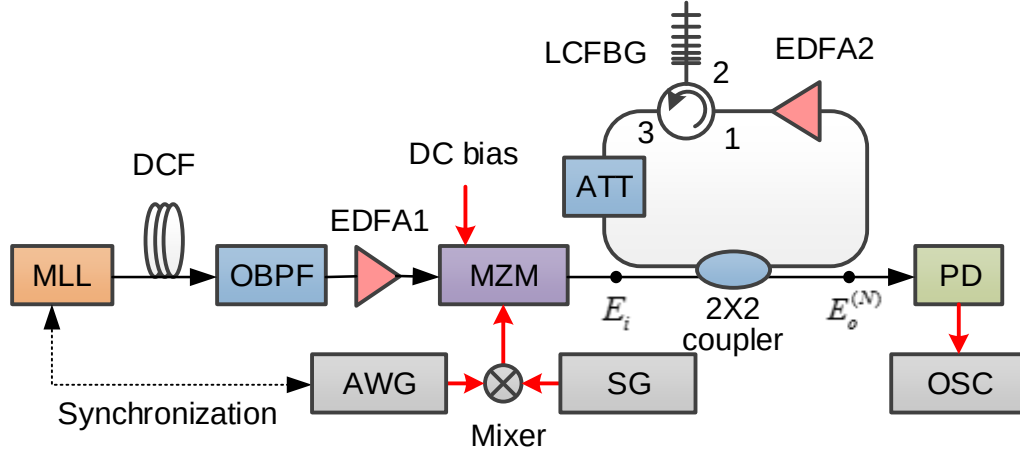


Fig. 6.1 Schematic of the time stretched sampling system. MLL: mode locked laser, OBPF: optical bandpass filter, MLL: mode-locked laser, DCF: dispersion compensating fiber, EDFA: erbium-doped fiber amplifier, MZM: Mach-Zehnder modulator, ATT: attenuator, LCFBG: linear chirped fiber Bragg grating, PD: photodetector, AWG: arbitrary waveform generator, SG: signal generator, OSC: oscilloscope.

It is known that an LCFBG has a quadratic phase response within its passband. Its transfer function can be written as [147]

$$H_{LCFBG}(\omega) = \exp\left(-j \frac{\beta_C}{2} \omega^2\right) \quad (6-1)$$

where ω is the optical angular frequency and β_C is the GDD coefficient of the LCFBG (in ps²).

Assuming that the optical spectrum at the input of the dispersive loop is $E_i(\omega)$, the output spectrum after recirculating for N round trips in the loop can be written as

$$E_o^{(N)} = \left(\frac{\sqrt{2}}{2}\right)^{N+1} g^N E_i(\omega) H^N(\omega), \quad (6-2)$$

where g is the net gain of the loop, which can be changed by tuning the gain of EDFA2 or the loss of the attenuator. If we make g close to but slightly less than $\sqrt{2}$, we then have $(\sqrt{2}g/2)^N \approx 1$. The transfer function of the loop can be expressed as

$$H_{loop}(\omega) = \frac{E_o^{(N)}}{E_i} = \frac{\sqrt{2}}{2} \exp\left(-j \frac{N\beta_c}{2} \omega^2\right) \quad (6-3)$$

By comparing Eqs. (6-1) and (6-3), we can see that the recirculating dispersive loop acts as a dispersive element that has an equivalent GDD coefficient of $N\beta_c$. It should be noted that, the ripples in the group delay of the LCFBG will also be magnified when a pulse recirculates for more round trips. It is preferable that an LCFBG to be used in a dispersive loop has a group delay ripple N times as small as a single time used LCFBG.

In (6-2), $\sqrt{2}g/2$ should always be smaller than unity to prevent the loop from lasing. As a result, the amplitude of $H_{loop}(\omega)$ should decay with the increase of N . The maximum number of N is determined by the minimum SNR required to detect the time-stretched signal. If the 2×2 coupler is replaced by an optical switch, then the number of round trips can be controlled by the optical switch. In this case, the equivalent GDD coefficient of the recirculating dispersive loop can be tunable by letting the waveform recirculate in the loop for a certain number of round trips.

The stretching factor of the time stretched sampling system is given by

$$M = 1 + N\beta_c/\beta_D \quad (6-4)$$

where β_D is the GDD coefficient of the pre-dispersion element. Since the second term in Eq. (6-4) is much greater than 1, it can be seen that the stretching factor increases proportionally to the number of the round trips N . Again, if an optical switch is employed in the system, the stretching factor can then be adjusted to improve the performance of the sampling system according to the frequency band of the input waveform. For example, N should be large for a fast microwave waveform so that all the details of the microwave waveform can be revealed, while for a relatively slow microwave waveform, N should be small to avoid over sampling and data redundancy [47, 48].

6.2 Experimental Implementation

An experiment based on the setup shown in Fig. 6.1 is performed. In the experiment, the MLL (IMRA femtolite 780) with a repetition rate of 48 MHz and a central wavelength of 1558 nm is employed to produce an optical pulse train. An individual pulse in the pulse train is nearly transform-limited with a 3-dB spectral bandwidth of 8 nm. The pre-dispersion element is a DCF with a dispersion coefficient of $\beta_C=432 \text{ ps}^2$ (or -339 ps/nm). The LCFBG used in the dispersive loop has a dispersion coefficient of $\beta_C=1912 \text{ ps}^2$ (or -1500 ps/nm) within a reflection passband of 0.6 nm centered at 1558 nm. The OBPF (Finisar WaveShaper 4000S) is configured to have a near rectangular passband with a bandwidth identical to that of the LCFBG. It can be calculated that, after the pre-dispersion by the DCF and the filtering by the OBPF, the MLL pulse is stretched to have a time duration of $\Delta\tau=203 \text{ ps}$. The optical pulse train at the output of the OBPF is amplified by EDFA1 and sent to the MZM. The MZM has a bandwidth of 20 GHz and is biased at its minimum transmission point. A microwave waveform generated by mixing an 18-

GHz microwave signal from the SG with a rectangular pulse train with a repetition rate of 286 kHz serving as a gate signal from the AWG is applied to the MZM. The repetition rate of the rectangular pulse train is smaller than that of the MLL (48 MHz) to reduce the duty cycle of the modulated optical pulse train, thus allowing pulse stretching with a large stretching factor without creating overlap between adjacent pulses. Note that in the experiment, the AWG and the MLL are synchronized. The modulated optical waveform at the output of the MZM is sent to the recirculating dispersive loop via the 2x2 coupler. The length of the recirculating dispersive loop is estimated to be 61 m (corresponding to a time delay of 305 ns). The time stretched optical pulse from the recirculating dispersive loop is sent to the PD (25-GHz, New Focus). The electrical waveform at the output of the PD is sampled by a real-time oscilloscope (Agilent DSO-X 93204A).

The modulation process is depicted in Fig. 6.2. As can be seen from Fig. 6.2(a) the gate signal with a repetition rate of 286 kHz and a gate duration of 20.8 ns is mixed with the 18-GHz microwave generated by the SG and sent to the MZM. Since the MZM is biased at its minimum transmission point, the pre-dispersed pulse train corresponding to the low voltage level of the modulation waveform will not be able to pass through the MZM, and the pre-dispersed pulse train corresponding to the high voltage level of the modulation waveform is modulated by a microwave waveform with twice the frequency of the microwave signal generated by the SG (i.e., 36 GHz), as shown in Fig. 6.2(b). Therefore, the number of microwave cycles in each MLL pulse is 7. In addition, there will be only one MLL pulse that is modulated by the microwave waveform in every period of the gate. The resulting pulse train with a reduced repetition rate is illustrated in Fig. 6.2(c). It should be noted that, for practical applications, one can use an MLL with a lower repetition rate so that the gate signal is not needed. Then, the MZM can be biased at

the quadratic point and the waveform carried by the MLL pulse will be the same as the modulation signal.

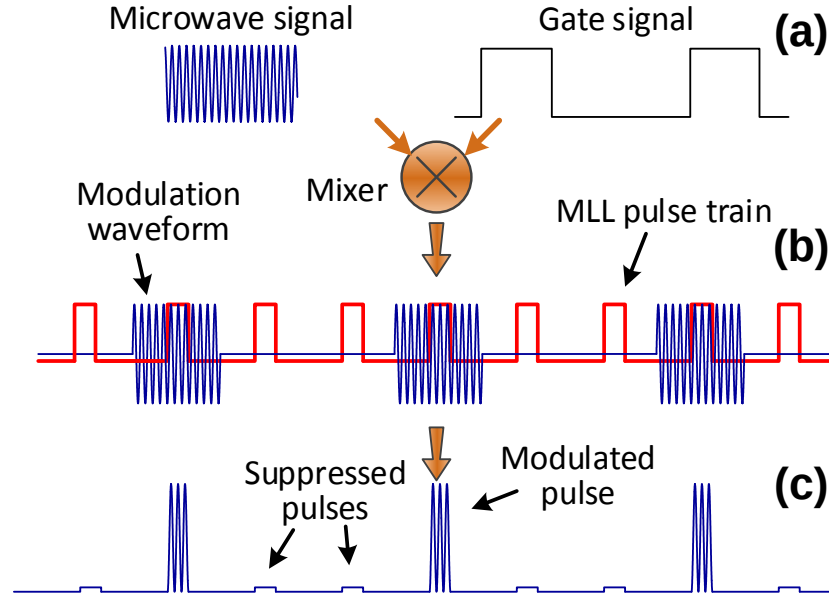


Fig. 6.2 The modulation process. (a) A 18-GHz microwave signal generated by the SG (solid-green line) and a gate signal generated by the AWG (black); (b) Waveform applied to the MZM (blue) and the MLL pulse train after pre-dispersion and filtering (red); (c) the resulted optical pulse train carrying the microwave waveform with a reduced repetition rate.

The modulated pulses are then injected to the recirculating dispersive loop. In every round trip, part of the optical pulse is coupled out of the loop by the 2x2 coupler and detected by the PD.

6.2 Experimental Results

Fig. 6.3 shows the measured MLL pulse at the output of the MZM. The full width at half maximum of the pulse is measured to be around 230 ps. Compared to the theoretical pulse width of $\Delta\tau = 203$ ps, the 27-ps difference could be caused by the relatively large sampling interval of 12.5-ps of the oscilloscope. The microwave waveform modulated on the pre-dispersed pulse

cannot be correctly sampled since the doubled microwave frequency of 36 GHz exceeds the highest frequency of the oscilloscope.

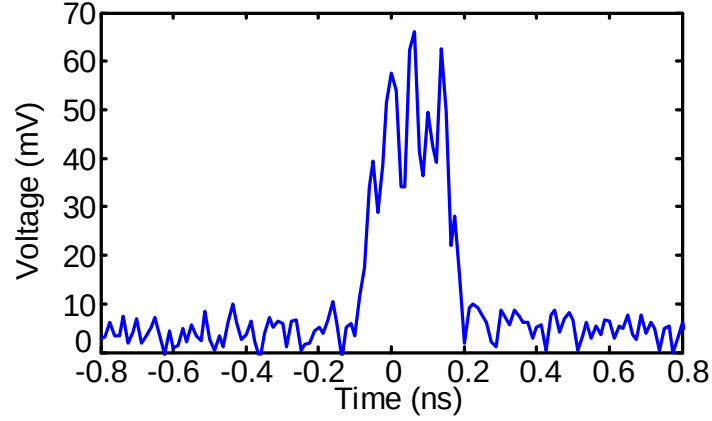


Fig. 6.3 The waveform of the modulated MLL pulse measured at the output of the MZM.

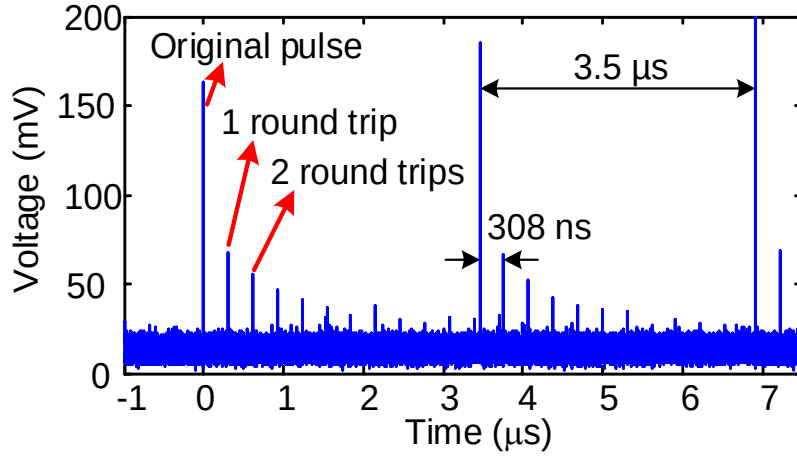


Fig. 6.4 Measured optical waveform at the output of the recirculating dispersive loop.

The waveform at the output of the dispersive loop is then measured by the real-time oscilloscope, as shown in Fig. 6.4. As can be seen when a modulated optical waveform is launched into the recirculating dispersive loop, a pulse burst with a decaying amplitude is generated. The quick decay in amplitude is due to the loss in the loop. To avoid lasing in the loop the gain of EDFA2 is controlled smaller than the loss. The time duration between two adjacent

pulses is 308 ns, which corresponds to the time delay of the recirculating dispersive loop. The time duration between two large pulses is 3.5 μ s, corresponding to the repetition time of the optical pulse train at the output of the MZM.

Theoretically, the decaying can be reduced by increasing the loop gain. However, due to the uneven magnitude response of the LCFBG and the uneven gain spectrum of EDFA2, the recirculating dispersive loop may start lasing at certain wavelength when the loop gain is increased, while at the other wavelengths, the loop gain is still smaller than 1. The relatively high noise floor is mainly caused by the amplified spontaneous emission of EDFA2 and the occasional lasing of the loop as the loop gain is very close to 1 at some wavelengths. To increase the SNR of the system, a LCFBG with specially design amplitude response or an EDFA gain flattening filter should be included in the loop to avoid lasing. One may also use an optical switch to replace the optical couple, so that the loop loss can be reduced, and a lower gain for EDFA2 will be required.

The detailed waveforms after the pulse is stretched in the recirculating dispersive loop for 1 to 8 round trips are shown in Fig. 6.5(a)-(h). After one round trip, the pulse duration is stretched to around 1 ns and all the seven microwave cycles with a temporal separation between two adjacent cycles of around 140 ps can be identified, as shown in Fig. 6.5(a). The optical pulse is stretched with a stretching factor of around 5.

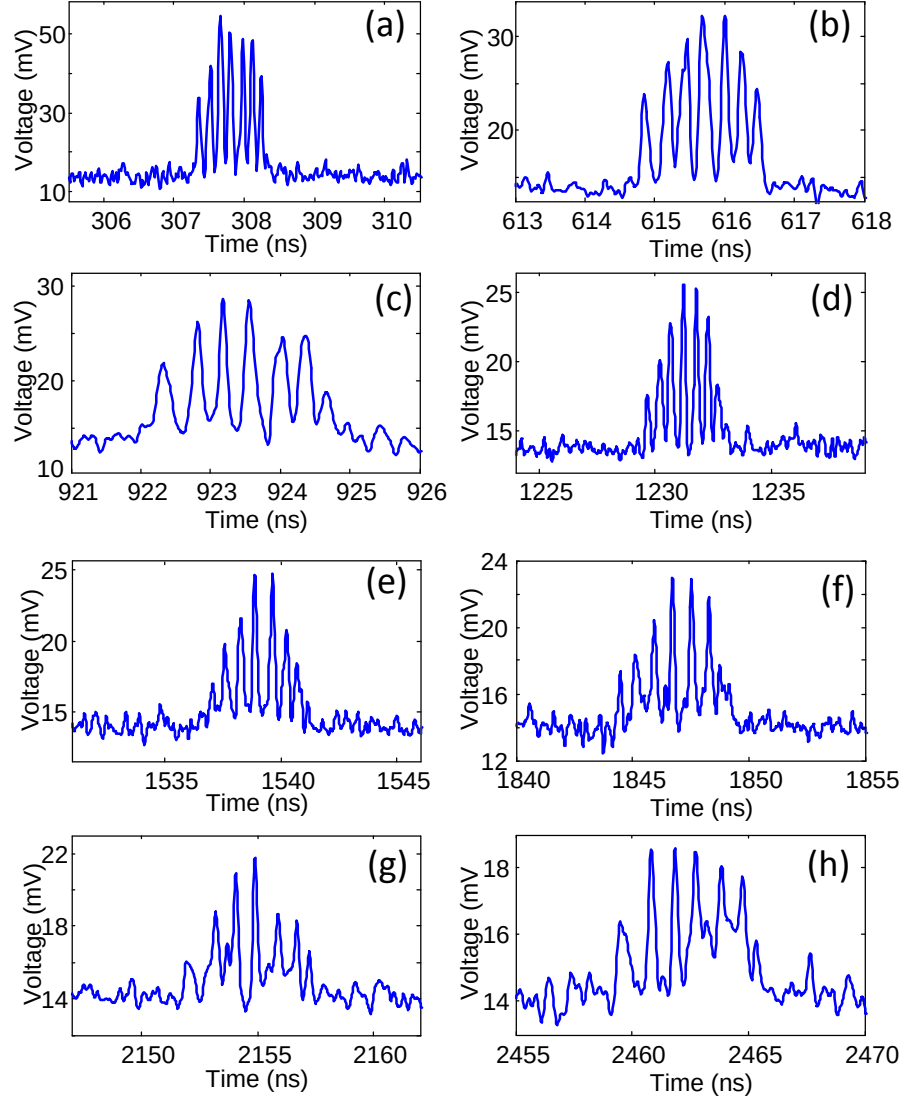


Fig. 6.5 The output waveforms after different number of round trips. (a) 1 round trip, (b) 2 round trips, (c) 3 round trips, (d) 4 round trips, (e) 5 round trips, (f) 6 round trips, (g) 7 round trips, and (h) 8 round trips. Note that the time scale is 1 ns/div in (a) to (c), and 5 ns/div in (d) to (h).

Then, the optical pulse keeps on recirculating in the loop, with the stretched pulses shown in Fig. 6.5(b)-(h). Since the round trip loss cannot be completely compensated by the gain of EDFA2, and the amplified spontaneous emission of EDFA2 introduces a significant amount of noise, the SNR drops after each round trip. By a fine control of the loop gain using the tunable attenuator, we are able to make the pulse circulate for 8 round trips before it is fully imbedded in

the noise. For the pulse after the 8th round trip, the equivalent GDD is $8 \times 1500 \text{ ps/nm} = 12000 \text{ ps/nm}$. The measured waveform after the 8th round trip shows that the pulse duration is round 7 ns and the average temporal separation between each microwave cycle is 1 ns. It indicates that a stretching factor of 36 is obtained, which is close to the theoretically calculated stretching factor of 36.4 based on Eq. (6-4). Assuming that the bandwidth of the system is limited by the oscilloscope used in our experiment, which is 32 GHz, the bandwidth of the sampling system can be as large as $36 \times 32 = 1.15 \text{ THz}$, corresponding to a time resolution of 347 fs. The frequency resolution, on the other hand, is limited by the time duration of the optical pulse used to carry the microwave waveform, which is $(203 \text{ ps})^{-1} = 4.93 \text{ GHz}$.

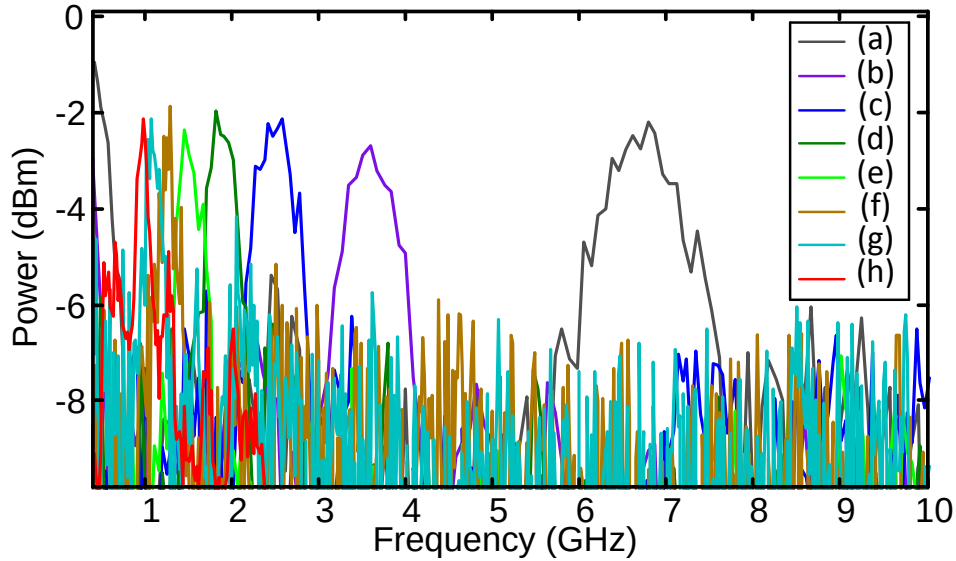


Fig. 6.6 The electrical spectra of the measured time-stretched waveforms for different number of round trips. (a)-(h) corresponds to the waveforms given in Fig. 6.5 (a)-(h).

Fig. 6.6 shows the electrical spectra of the measured waveforms for different number of round trips given in Fig. 6.5. The spectra show that the SNR decreases as the number of round trips increases, which agrees with our discussion. It can also be seen that, for a single frequency input, there is only one output frequency component, which means that the signal distortion

effect usually encountered by a time-stretched system has a weaker effect compared to the SNR deterioration, and is negligible.

6.3 Conclusion

A novel time-stretched sampling system with a large stretching factor has been demonstrated by a repetitive use of an LCFBG in a recirculating dispersive loop. An equivalent GDD as large as 12000 ps/nm with a large stretching factor of 36 was achieved. This is the 2nd largest dispersion-based stretching factor for a time-stretched sampling system ever reported. Although the stretching factor of 250 reported in [49] is much greater than the stretching factor of 36 in this work, we have demonstrated a dispersive element with a greater dispersion than that in [49]. If we use a pre-dispersion element that has a similar dispersion coefficient as the one in [49], we would be able to achieve a much greater stretching factor than 250. Note that for pulse stretching with a very large stretching factor, the input pulse applied to the MZM should be very short, thus the system can only be able to sample a microwave waveform with a narrow width.

It should also be noted that the stretching factor can be further increased by using a low noise optical amplifier. In addition, by flattening the magnitude response of the LCFBG and the gain of EDFA2, the net gain in the loop can be controlled to be very close to 1 but with no lasing, thus an input pulse can recirculate in the loop for more times, which would lead to a much greater stretching factor. In [29], an MPF with an ultra-even magnitude response was used to achieve pulse recirculation in an active cavity for 270 round trips. If this can be realized for a wideband LCFBG, the equivalent GDD can be as large as 4.05×10^5 ps/nm, which corresponds to a stretching factor of 1195.

CHAPTER 7 LINEARLY CHIRPED MICROWAVE WAVEFORM GENERATION

Microwave waveforms with a large TBWP have been widely employed in microwave sensors, spread-spectrum communications, microwave computed tomography, and modern instrumentation. Photonic generation of microwave signal, especially the one based on SS-WTT mapping technique, has attracted increasing interests due to its ability to achieve a waveform with a large bandwidth. However, the temporal duration of a signal generated using the SS-WTT mapping technique is limited by the maximum available dispersion coefficient of the DDL used to perform WTT mapping.

In this Chapter, a microwave waveform generator to generate an LCMW with an extended temporal duration by a repetitive multi-time use of an LCFBG in a dispersive fiber-optic recirculating loop that we demonstrated in the previous Chapter is proposed and experimentally demonstrated. A comprehensive analysis is provided, which is then verified by more detailed experiments. In addition to the increase in the temporal duration, we also demonstrate that the central frequency of the generated LCMW can be tuned. In the proposed system, the spectral shaper is a Fabry-Perot interferometer (FPI) incorporating two LCFBGs with complementary chirps to form an FP cavity with a linearly decreasing or increasing FSR. The spectrum of an ultra-short optical pulse is shaped by the FPI. The pulse is then directed into a dispersive loop consisting of a third LCFBG. Since the optical pulse is temporally stretched multiple times when reflected by the third LCFBG multiple times, an LCMW with an extended temporal duration that exceeds the physical length of the third LCFBG can be generated. Note that although a similar

dispersive loop has been used in [24], it is for a different application where fast signal sampling is implemented. The proposed technique is analyzed theoretically and validated experimentally.

7.1 Operation Principle

Fig. 7.1 shows the schematic diagram of the microwave waveform generation system. An ultra-short optical pulse train is generated by an MLL source. A repetition-rate-reduction module consisting of an MZM and an arbitrary waveform generator is used to realize the repetition rate reduction of the pulse train to avoid the overlapping of adjacent pulses when temporally stretched by the LCFBG in the dispersive loop. A gate signal is generated by an AWG with a repetition rate equals to that of the repetition-rate reduced pulse train. The pulse train is then sent via an OC (OC1) to an FPI formed by two complementary LCFBGs (LCFBG1 and LCFBG2), which is used as the spectral shaper. An EDFA (EDFA1) is employed after the MZM to compensate for the loss of the repetition-rate-reduction module. The pulse train is then launched into a dispersive loop, in which a third LCFBG (LCFBG3) is incorporated via a second OC (OC2). In the dispersive loop, a second EDFA (EDFA2) is employed to provide an optical gain, followed by an attenuator (ATT) to balance the gain to be slightly less than 1 to avoid lasing. The temporally stretched pulse is finally detected by a PD via a 2 x 2 optical coupler. An LCMW with an extended temporal duration is obtained at the output of the PD.

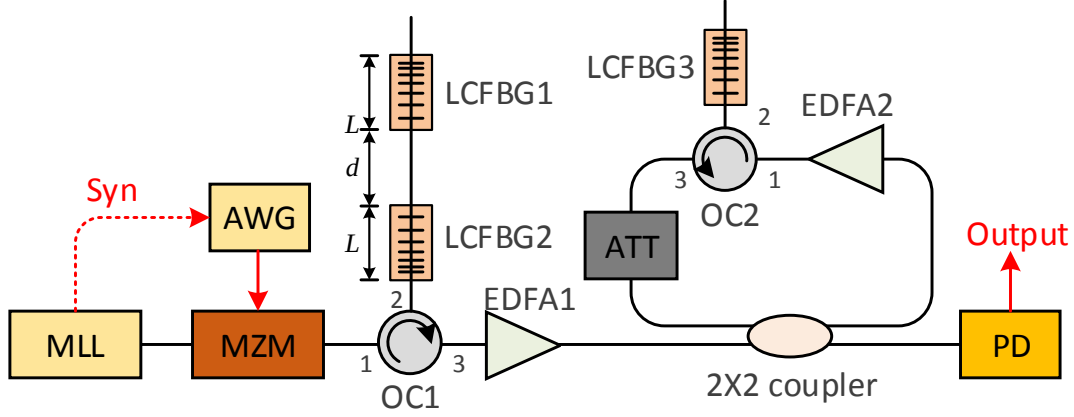


Fig. 7.1 Schematic diagram of the microwave waveform generation system. Syn: synchronization; MLL: mode-locked laser; AWG: arbitrary waveform generator; MZM: Mach-Zehnder modulator; OC: optical circulator; LCFBG: linearly chirped fiber Bragg grating; ATT: attenuator; EDFA: erbium-doped fiber amplifier; PD: photodetector.

Assuming that the dispersion coefficients of LCFBG1 and LCFBG2 are, respectively, $\ddot{\Phi}_1$ and $-\ddot{\Phi}_1$ -when looking into from the 2nd port of OC1, the cavity length of the FPI for a light wave with an angular frequency of ω is given by [27]

$$L(\omega) = d + \frac{2c\tau(\omega)}{n_{eff}} = d + \frac{2c\ddot{\Phi}_1(\omega - \omega_s)}{n_{eff}} \quad (7-1)$$

where $\tau(\omega)$ is the time delay caused by LCFBG1 and LCFBG2 for a light wave with an angular frequency of ω resonating in the FPI; c is the light velocity in vacuum; n_{eff} is the effective refractive index of the optical fiber; ω_s denote the lowest optical angular frequencies within the reflection bands of LCFBG1 and LCFBG2.

The FSR of the FPI can then be calculated by

$$\omega_{FSR} = \frac{2\pi c}{2n_{eff}L(\omega)} = \frac{\pi c}{n_{eff}d + 2c\ddot{\Phi}_1(\omega - \omega_s)} \quad (7-2)$$

Since both LCFBG1 and LCFBG2 are fabricated with low reflectivities, the reflection spectrum of the FPI should have an interference pattern within the reflection spectrum of LCFBG1 and LCFBG2. A simulated spectrum of an FPI formed by two identical LCFBGs, with an identical reflectivity of 10% and a bandwidth of 4 nm centering at 1551 nm, is given in Fig. 7.2. For comparison, an ideal linearly chirped sinusoidal function is also shown (in log scale). It can be seen that such an FPI has a spectral response that is similar to the shape of a sinusoidal function with an increasing period (or FSR) given by (7-2). The spectral response of the FPI can thus be written as

$$\begin{aligned}
 R(\omega) &= \sin\left(\frac{2\pi}{\omega_{FSR}} \omega + \varphi\right) \\
 &= \sin\left[\left(\frac{2n_{eff}d}{c} - 4\ddot{\Phi}_1\omega_s\right)\omega + 4\ddot{\Phi}_1\omega^2 + \varphi\right]
 \end{aligned} \tag{7-3}$$

where φ is an initial phase that will be interpreted as a microwave phase in the generated waveform. It can be seen from (7-2) and (7-3) that the FSR of the FPI is linearly increasing or decreasing, depending on the sign of $\ddot{\Phi}_1$.

After spectral shaping by the FPI and amplification by EDFA1, the spectrally shaped pulse is directed into the dispersive loop via the optical coupler. It has been theoretically proved in the previous Chapter that the dispersive loop has an equivalent dispersion coefficient of $N\ddot{\Phi}_3$ thanks to the multi-time use of LCFBG3, where $\ddot{\Phi}_3$ and N are the dispersion coefficient of LCFBG3 and the round trip number that the optical pulse travels in the loop, respectively. If the gain of EDFA2 can be controlled to fully compensate for the round-trip loss of the dispersive loop, N

can be an extremely large number, which would result in a very large equivalent dispersion coefficient and hence allowing a highly extended temporal duration for the generated LCMW.

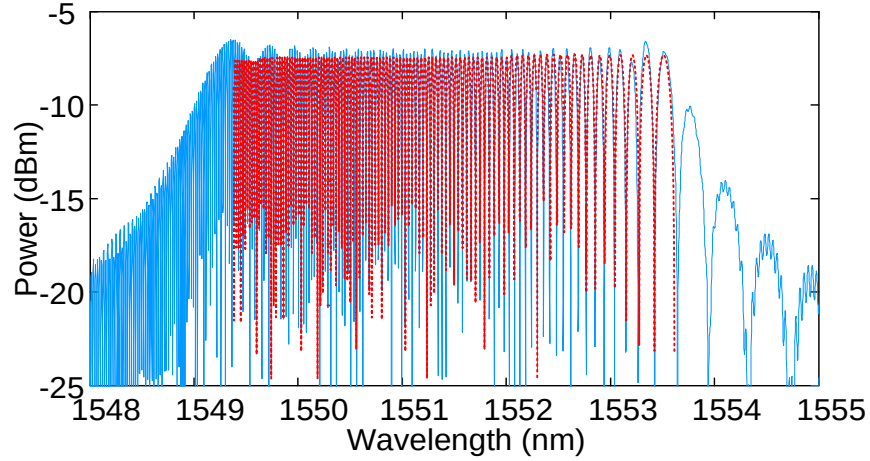


Fig. 7.2 Simulated reflection spectrum of an FPI formed by two LCFBGs with complementary dispersion (blue). The central wavelength and bandwidth of the two LCFBGs are 1551 nm and 4 nm. They are fabricated to have a uniform reflectivity of 10% and physically separated by 2 mm. The red dotted line is an ideal LCMW.

When the pulse recirculates in the loop, WTT mapping is performed. After N round trips, the electrical field at the output of the dispersive loop is given by [70]

$$y(t) = \exp\left(j \frac{1}{2N\ddot{\Phi}_3} t^2\right) X(\omega) \Bigg|_{\omega = \frac{t}{N\ddot{\Phi}_3}} \quad (7-4)$$

where $X(\omega) = G(\omega) \cdot R(\omega)$ is the optical spectrum of the pulse after spectrally shaped by the FPI, and $G(\omega)$ is the spectrum of a pulse from the MLL. In our case, the bandwidth of the pulse from the MLL is significantly larger than that of the optical spectral shaper, we can let $G(\omega) = 1$ for simplicity. In addition, the phase term in (7-4) will be eliminated by photo-detecting at a PD. Substituting (7-3) into (7-4), we get

$$y(t) = \sin \left(\frac{2n_{eff}d - 4c\ddot{\Phi}_1\omega_s}{cN\ddot{\Phi}_3}t + \frac{4\ddot{\Phi}_1}{N^2\ddot{\Phi}_3^2}t^2 + \varphi \right) \quad (7-5)$$

which precisely represents an LCMW with an instantaneous frequency of

$$f(t) = \frac{n_{eff}d - 2c\ddot{\Phi}_1\omega_s}{\pi cN\ddot{\Phi}_3} + \frac{4\ddot{\Phi}_1}{\pi N^2\ddot{\Phi}_3^2}t \quad (7-6)$$

The first term of (7-6) determines the central frequency of the LCMW, while the second term corresponds to the linear frequency chirping. The central frequency of the LCMW can be changed by adjusting the spacing between LCFBG1 and LCFBG2.

In our system, the optical bandwidth is limited by the FPI. Thus, it is required in (7-4) that

$$\omega_s \leq \frac{t}{N\ddot{\Phi}_3} \leq \omega_l \quad (7-7)$$

where ω_l denotes the upper frequency limit of the reflection bands of LCFBG1 and LCFBG2.

Then, the temporal duration of the LCMW can be deduced from (7-7)

$$\Delta\tau = N\ddot{\Phi}_3\Delta\omega \quad (7-8)$$

where $\Delta\omega = \omega_l - \omega_s$ is the bandwidth of the FPI. Substitute (7-7) into (7-6), the bandwidth of the generated waveform is derived which is given by

$$\Delta f = \frac{4\ddot{\Phi}_1\Delta\omega}{\pi N\ddot{\Phi}_3} \quad (7-9)$$

Multiplying (7-8) and (7-9), we get the TBWP,

$$TBWP = 4\ddot{\Phi}_1 \Delta\omega^2 / \pi. \quad (7-10)$$

It is seen that the TBWP of the LCMW is a constant even when the temporal duration is extended, since the bandwidth of the waveform is reduced when the waveform is temporally extended. The product between the two remains constant. This conclusion is true for linear temporal stretching. However, the use of the dispersive loop allows us to generate an LCMW with a time duration that is N times as long as the one without a dispersive loop, and the TBWP can be controlled to be large by designing an FPI with a wider bandwidth.

The central frequency of the generated LCMW can be tuned by adjusting the physical spacing between LCFBG1 and LCFBG2. A greater spacing corresponds to a smaller FSR, which would generate a waveform with a higher central frequency after WTT mapping. On the other hand, the bandwidth of the LCMW can be increased if the two LCFBGs in the FPI are designed to have larger dispersion coefficients, which leads to an FPI with a faster varying FSR.

7.2 Experimental Implementation

The LCMW generation system shown in Fig. 7.1 is then implemented. Fig. 7.3 gives a photograph of the experimental setup. An ultra-short optical pulse train is generated by an MLL (PriTel FFL-1550-20). The repetition rate and the central wavelength of the pulse train are 20 MHz and 1551.5 nm, respectively. The 3-dB spectral bandwidth of an individual pulse is 6 nm, with a transform limited temporal width of 550 fs. The gate signal with a repetition rate of 1.18 MHz or a period of 850 ns produced by the AWG (Tektronix AWG7102) provides a 50-ns long time window to reduce the repetition rate of the pulse train from 20 MHz to 1.18 MHz. The

MZM is configured to operate as an optical switch, by biasing it at its minimum transmission point (switch off) and the maximum transmission point (switch on), corresponding to the gate is close and open, respectively. Note that if the MLL has a smaller repetition rate, the AWG and the MZM will not be needed and the system can be simplified. LCFBG1 and LCFBG2 forming the FPI are fabricated to have a bandwidth of 4 nm centered at 1551.5 nm and a dispersion coefficient of $\pm 25 \text{ ps}^2/\text{rad}$. Two grating pairs with two different physical separations of 2 mm and 2 cm between LCFBG1 and LCFBG2 are fabricated to generate LCMWs with two different central frequencies. The reflection spectra of the two FPIs are shown in Fig. 7.4(a) and (b). A linearly increasing FSR is observed for both FPIs. The FPI with a larger physical separation, i. e., longer FPI cavity, has a smaller FSR that can be used for the generation of an LCMW with a higher central frequency. In the reflection spectra shown in Fig. 7.4(a) and (b), strong amplitude ripples are observed especially for the smaller the FSR end. The ripples are introduced by the limited wavelength sampling interval of the OVA (LUNA Technologies) used to measure the spectra of the FPIs. The wavelength sampling interval is 2.4 pm, while the smallest FSR is 13 pm. LCFBG3 in the dispersive loop is fabricated to have a 4-nm reflection bandwidth with a center wavelength of 1551.5 nm and an in-band dispersion coefficient of $-3188 \text{ ps}^2/\text{rad}$. Thanks to the multi-time use of LCFBG3 in the loop, a large equivalent dispersion coefficient can be achieved. For example, if the pulse is recirculating in the loop for five round trips, the equivalent dispersion coefficient would be as large as $15940 \text{ ps}^2/\text{nm}$. A PD (New Focus 1414, 20-GHz bandwidth) is used to detect the temporally stretched optical pulse to get a microwave waveform.

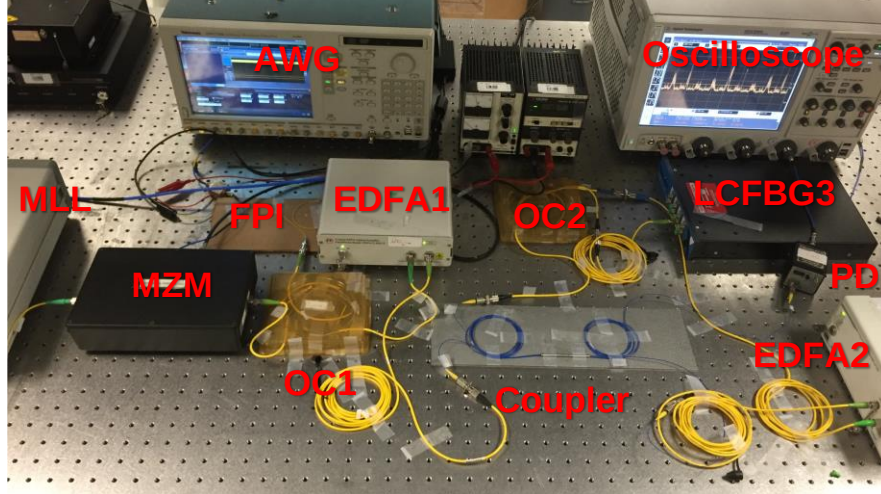


Fig. 7.3 Photograph of the experimental setup.

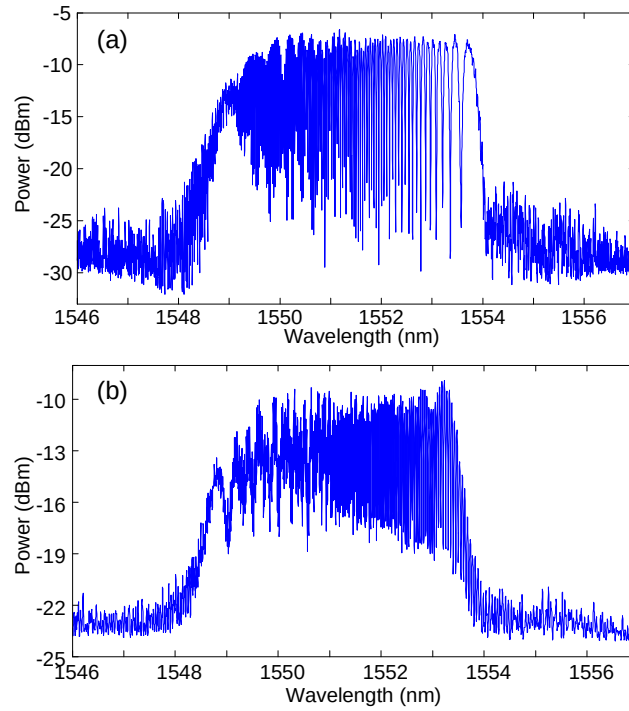


Fig. 7.4 Reflection spectra of the FPIs with a physical spacing between LCFBG1 and LCFBG2 of (a) 2 mm and (b) 2 cm.

It should be noted that two EDFAs are used in the system. The first EDFA is required by the repetition-rate-reduction module to compensate for the insertion loss in the module. Since only 1 out of 17 pulses is selected by the MZM, the reduction in the repetition rate would

introduce 12.3 dB insertion loss. If the insertion loss of the MZM of 5 dB is included, the total insertion loss is 17.3 dB. If the pulse train generated by the MLL has a longer repetition period, the first EDFA and the MZM will not be needed, and the system will be simplified. The insertion loss of the FPI in the dispersion loop is 7.5 dB. The second EDFA is also required to compensate for the round trip loss, to allow the pulse to recirculate for more round trips in the dispersive loop.

7.3 Experimental Results

The LCMW generated at the PD is monitored by an oscilloscope (Agilent DSO-X 93204A). First, we use the FPI with a separation of $d = 2$ mm as the optical spectral shaper. Fig. 7.5 shows two LCMWs after the pulse recirculates for three and five round trips in the loop. The LCMWs have decreasing periods, indicating a frequency up-chirp. Compared with a single-time use of LCFBG3 that would generate an LCMW with a duration of less than 10 ns, extended temporal durations of around 25 and 42 ns are obtained for the two LCMWs. It should be noted that the waveforms should have temporal durations of 30 and 50 ns calculated theoretically based on (8). The differences in the temporal durations are due to the errors in the fabrication of the LCFBGs, which would cause a reflection band mismatch between the LCFBGs. The amplitude ripples shown in the LCMW spectra are due to the ripples in the spectrum of an MLL pulse, the non-flat gain spectra of the EDFAs, and the ripples in the reflection spectra of the LCFBGs. The ripples can be mitigated by adding an optical gain-flattening filter in the dispersive loop.

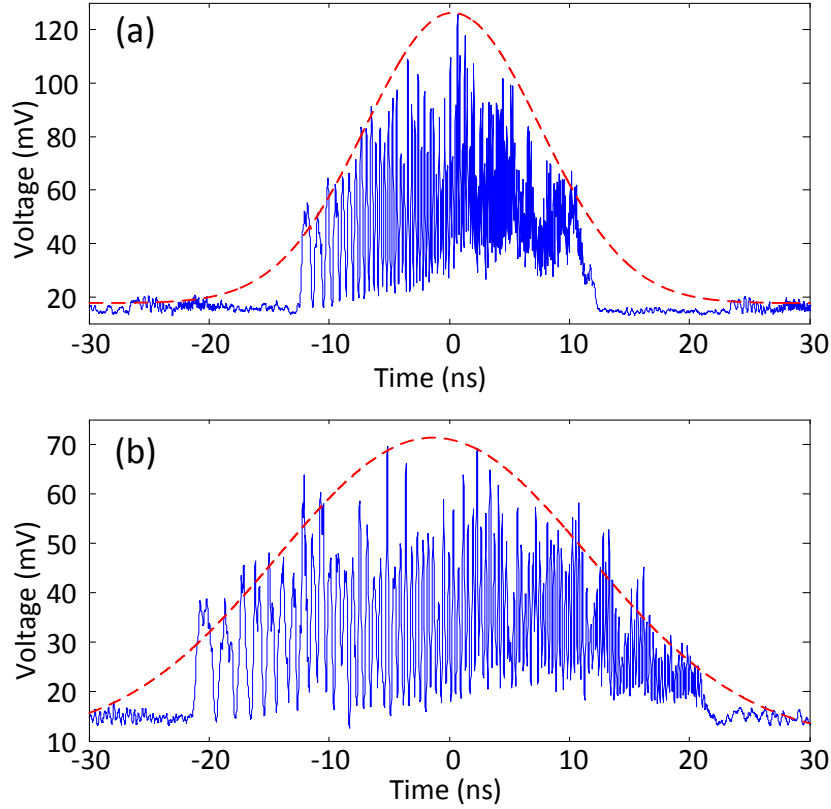


Fig. 7.5 Generated LCMWs using the FPI with a physical spacing between the two LCFBGs of 2 mm with (a) three and (b) five round trips. The red dash lines represent the pulse profiles induced by the spectral shape of a MLL pulse.

The spectrograms of the generated LCMWs shown in Fig. 7.5 are calculated and shown in Fig. 7.6. Linearly increasing instantaneous frequencies can be observed for the two generated LCMWs, which indicate a good linearity of the frequency chirping of the waveforms. The two LCMWs have bandwidths of 8.4 and 5.0 GHz with an identical TBWP of around 210. However, the temporal durations are extended thanks to the greater equivalent dispersion coefficient of the dispersive loop. According to (7-10), the theoretical TBWP of the system is estimated to be 315. Since WTT mapping is only performed to part of the spectrum (82%, in our case) shown in Fig. 7.4 due to the mismatch between the reflection bandwidths of the LCFBGs, the temporal durations and bandwidths of the LCMWs are reduced.

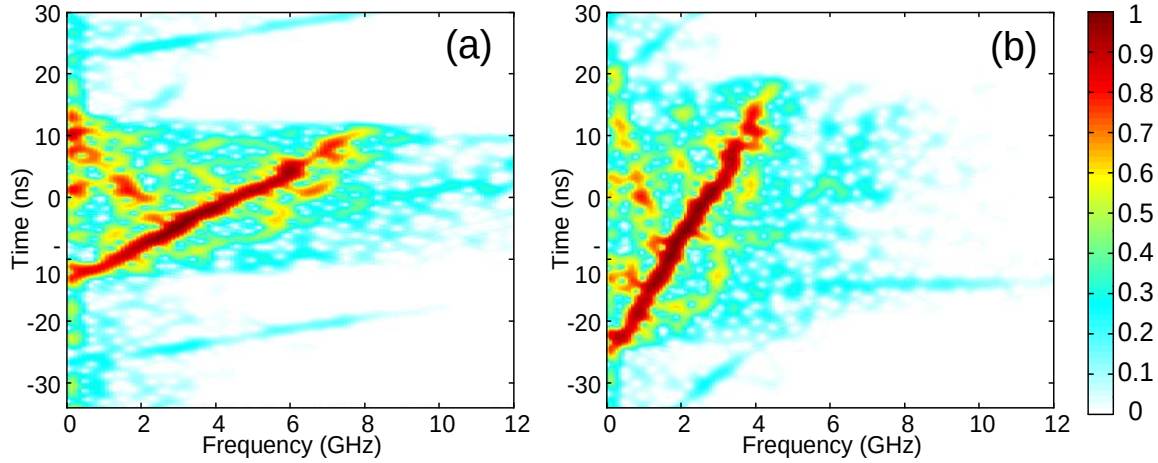


Fig. 7.6 Spectrograms of the LCMWs for (a) three and (b) five round trips. The color scale represents the normalized amplitude of the spectrogram.

In a microwave receiver, an LCMW is compressed by a matched filter to improve the range resolution. The calculated correlation results between an LCMW and its reference are presented in Fig. 7.7(a) and (b). The widths of the correlation peaks are 100 ps and 160 ps for the two LCMWs after three and five round trips, which correspond to two suppression ratios of 250 and 262, respectively.

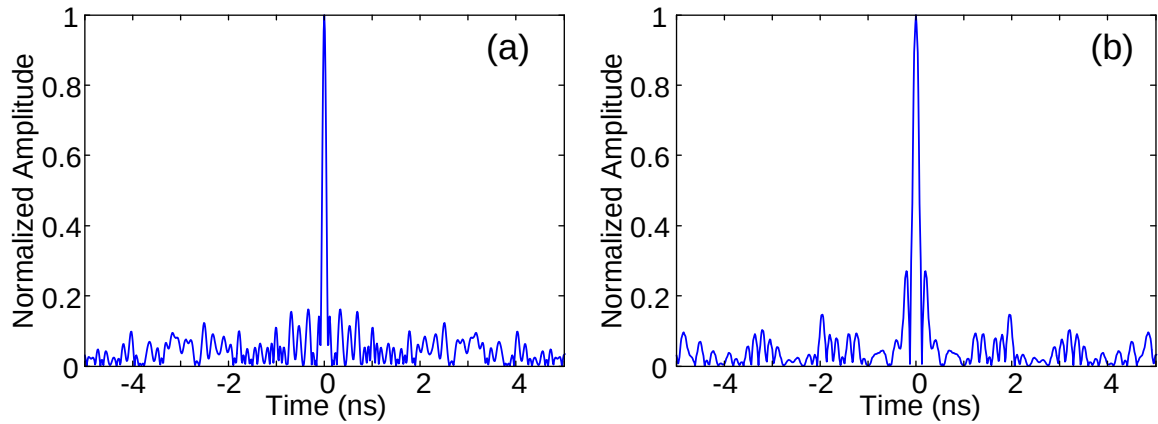
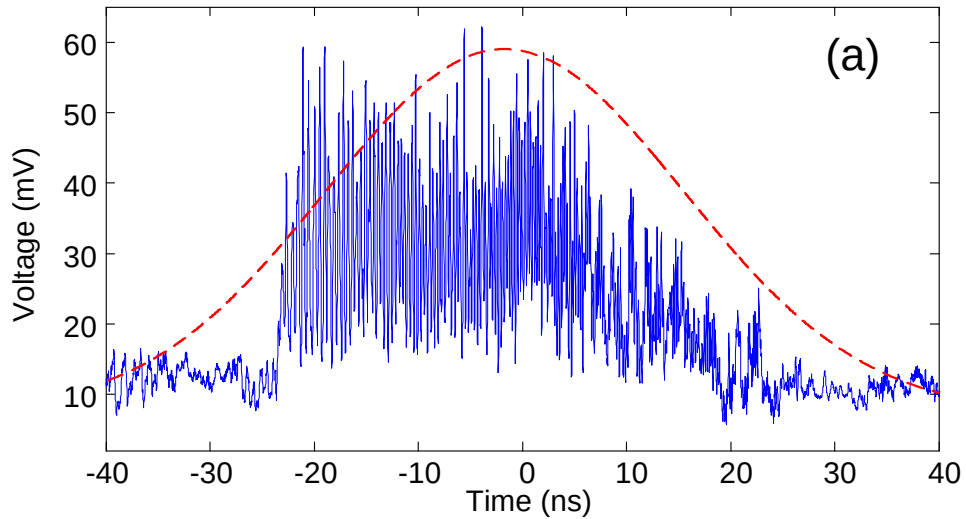


Fig. 7.7 Calculated autocorrelation between the LCMWs and their references. For the FPI with a spacing of (a) 2 mm, and (b) 2 cm.

To generate an LCMW at a different frequency band, a second FPI with a physical separation of $d = 2$ cm is then employed as the optical spectral shaper. The LCMW for five round trips are shown in Fig. 7.8(a). A 45-ns long LCMW is achieved. The minimum frequency is 1.5 GHz, instead of around DC for the LCMW shown in Fig. 7.5. However, strong attenuation can be observed for the high frequency components due to a lower responsivity of the PD at a higher frequency band. The spectrogram in Fig. 7.8(b) indicates a TBWP of only 180, which is also caused by the lower responsivity of the PD at the higher frequency band. The calculated autocorrelation of the LCMW shows a width of the correlation peak of 358 ps. A compression ratio of 125 is achieved for the 45-ns long LCMW. It should be noted that amplitude ripples in the spectra of the LCMWs shown in Fig. 7.5 and Fig. 7.8 are observed. The ripples are caused due to the lasing in the dispersive loop since the gains at certain wavelengths are near the lasing threshold. To reduce the ripples, in the dispersive loop, a gain-flattening filter may be used to flatten the gain spectrum of the EDFA, so that the lasing can be suppressed.



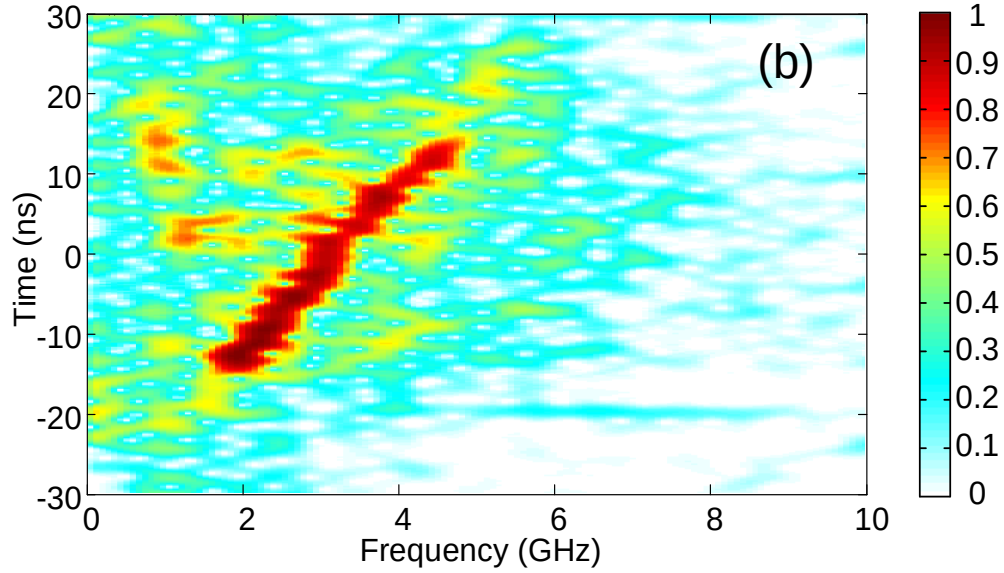


Fig. 7.8 (a) Generated LCMW using the FPI with a spacing of 2 cm after the optical pulse recirculates for five round trips and (b) the corresponding spectrogram. The color scale represents the normalized amplitude of the spectrogram. The red dash line in (a) represents the pulse profile induced by the spectral shape of a MLL pulse.

The stability of the proposed LCMW generator is also studied. We first investigate the short term stability. To do so, two LCMWs that are separated in time by 60 cycles ($45 \mu\text{s}$) are sampled and compared. The LCMWs have very similar shapes, including the amplitude ripples and the phase responses, indicating good stability and repeatability of the operation of the system. The cross-correlation between the two LCMWs is also calculated which is identical to the auto-correlation of one of the LCMWs. This again demonstrates a stable and repeatable operation of the system. The long term stability is strongly affected by the ambient temperature change, as the FPI is temperature sensitive. By using a temperature control unit, the long term stability can be improved.

7.4 Conclusion

An approach to the generation of an LCMW with an extended temporal duration implemented by an FPI for spectral shaping and a dispersive loop for wavelength-to-time mapping was proposed and experimentally demonstrated. Long temporal duration for the generated LCMW was enabled by multi-time use of an LCFBG in a dispersive loop to perform WTT mapping. Two LCMWs with two temporal widths of 25 and 42 ns were generated at two different frequency bands. A further increase in the temporal durations of the LCMWs is possible by allowing the optical pulse recirculate for more round trips in the loop. The TBWPs of the two LCMWs were both 210 and the extension of the temporal duration of an LCMW will not increase the TBWP for a given FPI. To increase the TBWP, an FPI with two LCFBGs having larger dispersion coefficients may be used. For example, if two LCFBGs with two opposite dispersion coefficients of $\pm 3188 \text{ ps}^2$ are used to form the FPI and a PD with a bandwidth of over 100 GHz is used to perform photodetection [75], an LCMW with a TBWP as large as 4200 can be generated.

CHAPTER 8 PHOTONIC TRUE-TIME DELAY BEAMFORMING

A beamforming network is required to produce progressive phase or time delays for a PAA, which can be implemented using phase shifters or true-time delay lines. In the past few years, numerous photonic true-time delay beamforming networks have been demonstrated due to the large operation bandwidth of a photonic system. However, all of the photonic true-time delay beamforming networks requires a TLS array to achieve multi-channel tunable time delay, which makes the system expensive and instable.

In this Chapter, we introduce a fiber-optic true-time delay beamforming network using a switch-controlled recirculating wavelength-dependent dispersive loop incorporating an LCFBG that only requires a laser array with fixed wavelengths. In the proposed system, a microwave signal to be radiated to the free space is modulated on the multi-wavelength carrier from the laser array, which is sent to a switch-controlled recirculating wavelength-dependent dispersive loop. Since the optical signals with different wavelengths are reflected at different locations of the LCFBG, different time delays are achieved. The tuning of the time delays are realized by controlling the number of round trips the optical signals recirculate in the loop, which is done by using a 2×2 optical switch to direct the optical signals back to the loop for additional time delays or output of the loop. The proposed true-time delay beamforming network is experimentally demonstrated. A four-channel true-time delay beamforming network using two different recirculating dispersive loops with a true-time delay of 2.5 ns and 160 ps per round trip incorporating two LCFBGs with different dispersion coefficients are demonstrated. The use of

the beamforming network to achieve arrayed beamforming that can cover -90° to 90° is demonstrated.

8.1 Photonic True-Time Delay Based on a Dispersive Loop

The schematic diagram of the proposed true-time delay beamforming network is shown in Fig. 8.1. The light waves from four laser diodes (LDs) with different wavelengths of λ_1 to λ_4 are combined at a wavelength-division multiplexer (WDM1) and sent to an MZM, where a microwave signal, generated by an electrical AWG, is modulated on the four wavelengths. The modulated optical signals at the output of the MZM is then launched into a switch-controlled wavelength-dependent optical dispersive recirculating loop through a 2x2 optical switch. An LCFBG is incorporated in the loop via an OC to provide a wavelength-dependent time delay. An EDFA1 is also incorporated in the loop to compensate for the loss in the loop so that an optical signal can recirculate in the loop for multiple round trips. A programmable optical filter (OF) with four passbands centered at λ_1 to λ_4 is also employed to suppress the ASE noise from the EDFA. The number of round trips is controlled by the 2x2 optical switch. At the output of the loop, a second EDFA (EDFA2) is used to further amplify the optical signal and a second WDM (WDM2) is used to demultiplex the time-delayed optical signals in the four channels, which are converted to four time delayed microwave signals at the PDs.

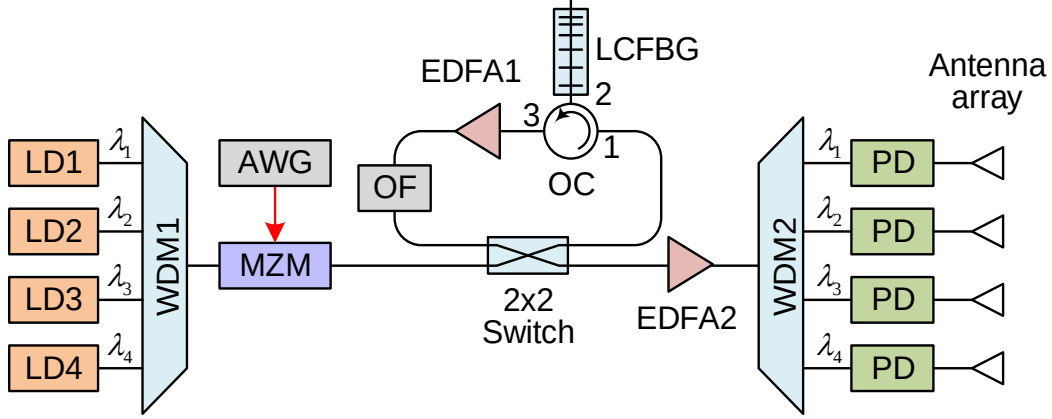


Fig. 8.1 Schematic diagram of the true-time delay beamforming network using a recirculating wavelength-dependent dispersive loop. LD: laser diode; WDM: wavelength-division multiplexer; AWG: arbitrary waveform generator; MZM: Mach-Zehnder modulator; OF: optical filter; EDFA: erbium-doped fiber amplifier; OC: optical circulator; LCFBG: linearly chirped fiber Bragg grating; PD: photodetector.

The feed signal $x(t)$ from the AWG is modulated on the optical carriers. The optical switch is configured at the cross state, so that the optical signals at different wavelengths can be directed into the recirculating dispersive loop. After the signals enter the loop, the switch is configured at the bar state. Thus, the optical signals will be recirculating in the loop until the state of the switch is changed from bar to cross. Depending on the number of round trips in the loop, the optical signals will experience different time delays.

Due to wavelength-dependent time delay resulted from the LCFBG in the loop, the time delay for a wavelength λ_i ($i=1, 2, 3, 4$) is given by

$$T_i = t_0 + \beta(\lambda_i - \lambda_r) \quad (8-1)$$

where t_0 is the fixed time delay of the loop excluding the LCFBG; β is the dispersion coefficient of the LCFBG and λ_r is a reference wavelength corresponding to the wavelength of the end of the LCFBG connected to the optical circulator. Assuming the feed signal $x(t)$

recirculates for N round trips, a time delay signal $x_i(t - NT_i)$ can be achieved at the output of the loop. The time delay difference between the signals carried by two adjacent wavelengths, $x_i(t - NT_i)$ and $x_{i+1}(t - NT_{i+1})$, is given by

$$\Delta t = NT_{i+1} - NT_i = N\beta\Delta\lambda \quad (8-2)$$

where $\Delta\lambda$ is the wavelength spacing between the two carriers. Note that the wavelengths are uniformly spaced in our system. It can be seen that progressive true-time delays can be obtained for the optical signals as the number of round trips n increases. The relationship between the time delays, the number of round trips and the carrier wavelengths are sketched in Fig. 8.2. By increasing the number of round trips, a time-delay increment of $\beta\Delta\lambda$ can be achieved. A relationship between the beam steering angle and the number of round trips that the signal recirculates in the loop can then be written as [19]

$$\theta = \sin^{-1}\left(\frac{cN\beta\Delta\lambda}{d}\right) \quad (8-3)$$

where d is the spacing between two adjacent antenna unit in a PAA; c is the light velocity in vacuum. It can be seen that, as N increases, the beam pointing direction will be scanning, with a scan step determined by $\beta\Delta\lambda$ and a scan range determined by the maximum value of N . In our system, N is a large finite number as only part of the round trip loss in the loop can be compensated by the EDFA.

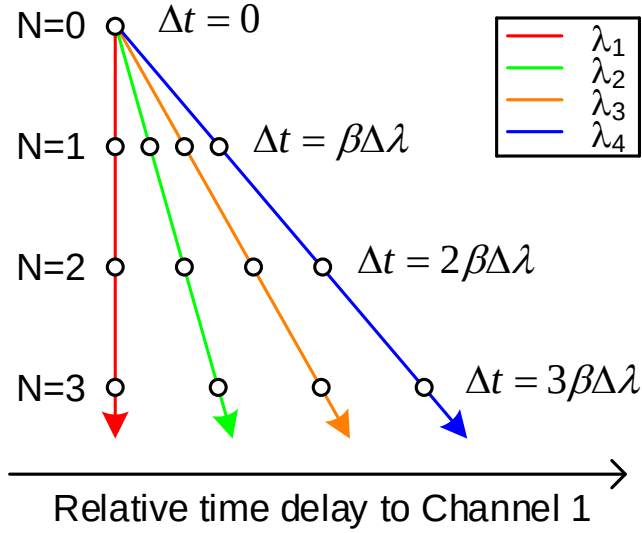


Fig. 8.2 The time delay of the signal in each channel relative to channel 1 as the number of round trips N increases.

In microwave sensing applications, the temporal duration and the bandwidth of the feed signal determine its spatial resolution and measurement range, and therefore are of great importance [122]. In our system, it is required that the duration of the original microwave signal should be smaller than the round trip time of the loop, so that the signal will not overlap with itself when recirculates in the loop, i.e., the duration of the original microwave signal $\tau < t_0$. Thanks to the small loss of optical fibers, t_0 can be easily increased by using a long optical fiber in the loop. Note that the loop length cannot be chosen arbitrarily long, since a longer dispersive loop will slow down the beam steering speed. In addition to radial resolution, the angular resolution of a radar is determined by the beam steering step, which can be calculated from (8-3) for our system.

The bandwidth limit, on the other hand, is mostly determined by the opto-electronic devices used in the system, such as the MZM and the PD. However, the dispersion-induced power penalty [19] should also be considered as it may get significant due to the large equivalent

dispersion of a recirculating loop including an LCFBG, which is deduced to be $N\beta$ in Chapter 6. To avoid the first zero-response point in the power distribution function of a system suffering from dispersion-induced power penalty, it is required that the microwave signal frequency [19]

$$f < \sqrt{\frac{c}{2N\beta\lambda_i^2}} \quad (8-4)$$

It should be noted that the dispersion-induced power penalty can be completely eliminated by employing single-sideband with carrier (SSB+C) modulation [19], which is an effective way to overcome the bandwidth limitation imposed by (8-4).

8.2 Experimental Implementation

Fig. 8.3 shows a photograph of the experimental setup, which is based on the schematic diagram in Fig. 8.1. First, the experiment is carried out to have a large true-time delay step, i.e., $\Delta\beta\lambda$ is chosen to be large. The central wavelengths of the four LDs (Agilent N7714A) are set to be 1548.9, 1549.9, 1550.9 and 1551.9 nm, which have a uniform wavelength spacing of 1 nm. The OF is then programmed to have four 10-GHz-wide passbands at the same wavelengths. The optical spectrum when the output of the four LDs are combined with the WDM presented in Fig. 8.4. Since the output power of the LDs are independently tunable, a flat optical comb is generated. The LCFBG is fabricated to have a dispersion coefficient of 2500 ps/nm within its 4-nm reflection band centered at 1550.5 nm (Fig. 8.5). It can be calculated that $\beta\Delta\lambda = 2.5$ ns, i.e., a true-time delay of 2.5 ns can be achieved between two adjacent channels when the signal recirculates in the loop for one round trip. The AWG (Tektronix 7102), which has a sampling rate of 10 Gb/s, is configured to generate the feed microwave signal. A 2x2 coupler is used

instead of the 2x2 switch to simplify the experiment and to study the round-trip-by-round-trip behavior of the system. Although the coupler cannot actively control the number of round trips that the optical signal recirculates in the loop, part of the optical signal will be coupled out of the loop after each round trip, generating a stepped increasing true-time delay that corresponds to a scanning steering beam angle for a PAA. Additionally, the extra 3-dB loss induced by the coupler can be compensated by the EDFA in the loop. The time-delayed signals are detected by four PDs with bandwidths of over 20 GHz and sampled by a 4-channel oscilloscope (Agilent 93204A).

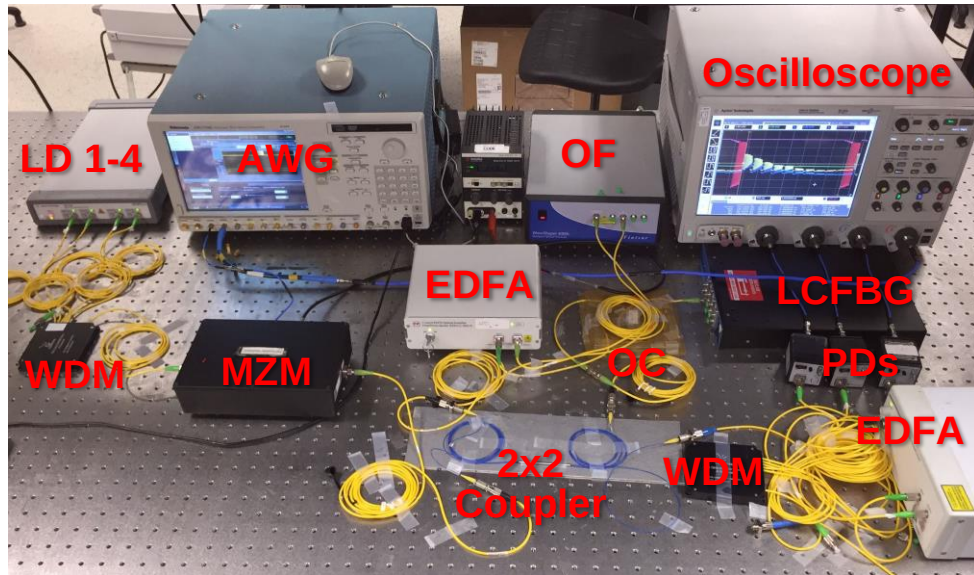


Fig. 8.3 The photograph of the experimental setup.

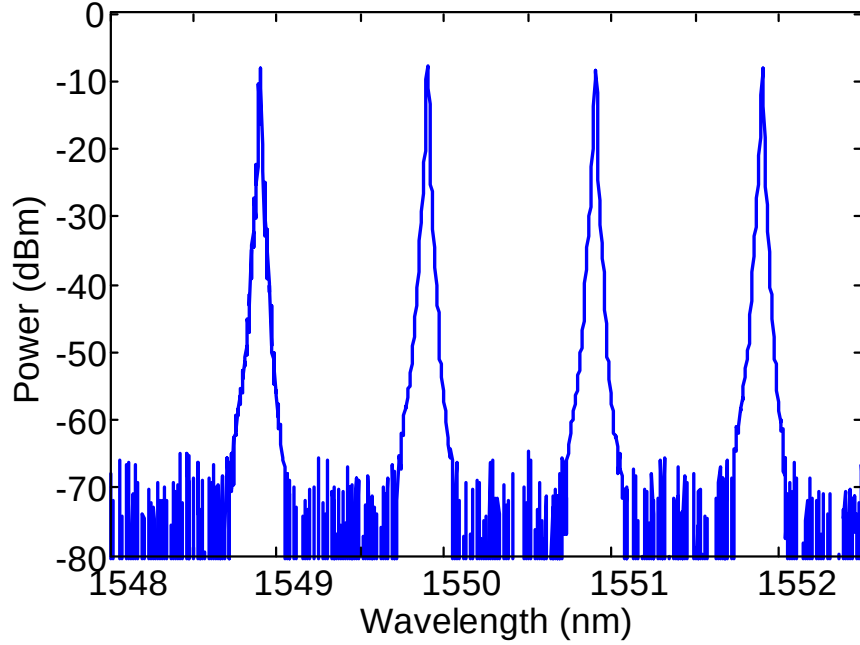


Fig. 8.4 Optical carrier spectrum.

To demonstrate the operation of the system, an electrical pulse with a temporal width of 1 ns is generated by the AWG. The waveforms detected at the outputs of the four PDs following the second WDM are shown in Fig. 8.6(a). Multiple time-delayed replicas of the electrical pulse can be detected at the output of each channel with a repetition time of around 380 ns. In addition, the pulses from the four channels overlap in time only for $N=0$ and an increasing time shifts between the time delayed pulses can be observed for $N>0$, indicating that a true-time delay has been achieved between the channels when the pulse starts to recirculate in the loop. A reduced amplitude can be seen as N increases, as the EDFA cannot be configured to fully compensate for the loop loss to avoid lasing. Nevertheless, the pulses can still recirculate for more than 10 round trips before being imbedded in noise. The reduced amplitude may be compensated by a power amplifier that is commonly used in PAA. The bandwidth limit of the system for $N=10$ is calculated to be 1.58 GHz by (8-4), which is relatively small for broadband applications. More detailed waveforms are shown in Fig. 8.6(b)-(c) for different numbers of round trips of 0, 2 and 4.

The time spacing between two pulses from two adjacent channels are measured to be 5.0 and 10.0 ns for $N=2$ and $N=4$, respectively, which are in perfect agreement with the theoretical values of 2.5 ns per round trip. Due to the relatively large true-time delay, the system can be used for PAAs with large antenna spacing and with a low operation frequency. Fig. 8.7 shows the radiation patterns of a four-element linear PAA, which has a uniform element spacing of 5 m and uses the time delayed signal after different numbers of round trips to feed the antenna elements. A steering angle of 8.6° can be observed when the feed signal recirculates for one additional round trip in the loop, which is in good agreement with (8-3). In four round trips, the beam pointing direction is steered for 36.8° . With a sufficiently large N (over 13), the steering range can fully cover -90° to 90° .

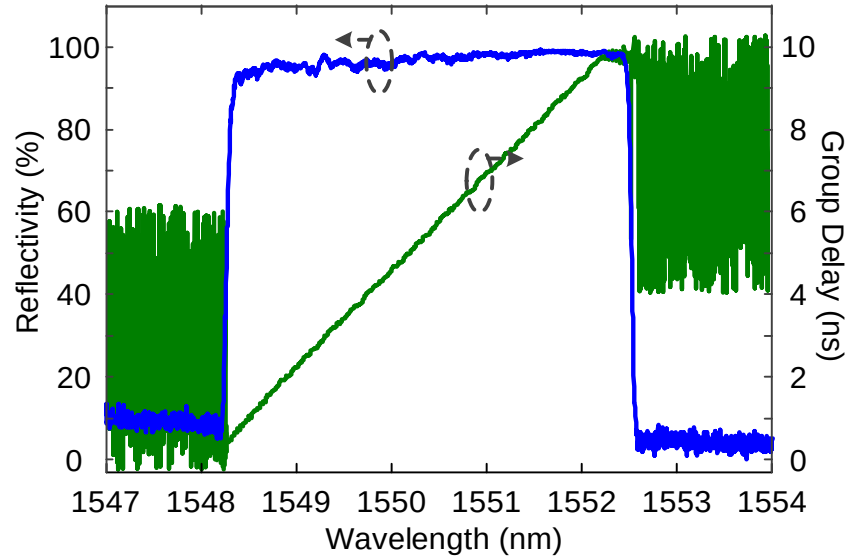


Fig. 8.5 Spectral response of the LCFBG.

8.3 Performance Evaluation

An LCMW with a large bandwidth and a long time duration is commonly used as a feed signal [70]. To test the performance of the system for a signal with long duration, the AWG is

configured to generate an LCMW with a time duration of 300 ns and a chirp rate of 2 MHz/ns starting from DC. Fig. 8.8(a) shows the detected signals. Again, the signal carried by each wavelength has a repetition time of 380 ns. The zoomed-in view of the signals after light recirculating for 0, 2 and 4 round trips are shown in Fig. 8.8(b)-(c). An increasing true-time delay can be observed, which again agrees with the theoretical values. Since the duration of the LCMW is shorter than the repetition time, the time delayed LCMWs do not overlap in time. High signal-to-noise ratio is achieved even when the signal recirculates for more than six round trips in the loop.

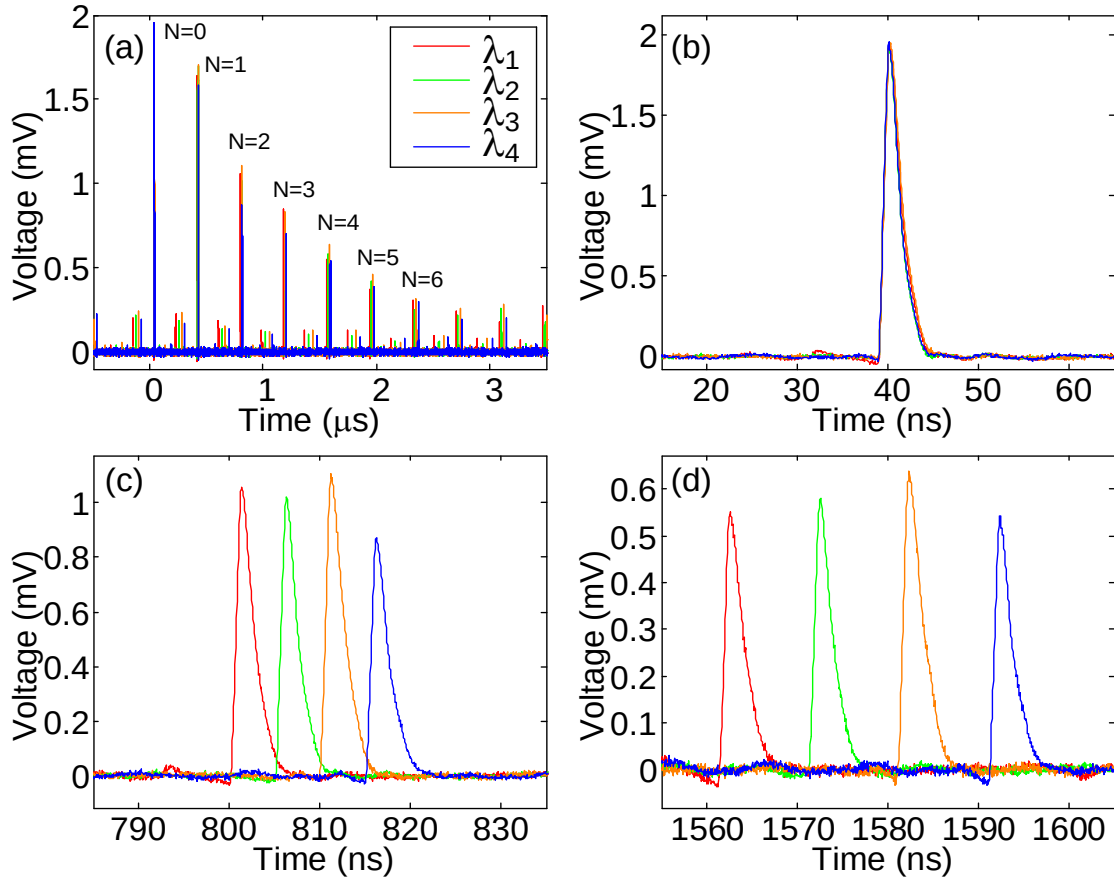


Fig. 8.6 Measured signals at the outputs of the four PDs when an electrical pulse is applied to the MZM. (a) The generated time delayed replicas and the zoom-in view of the signals for (b) $N=0$, (c) $N=2$ and (d) $N=4$.

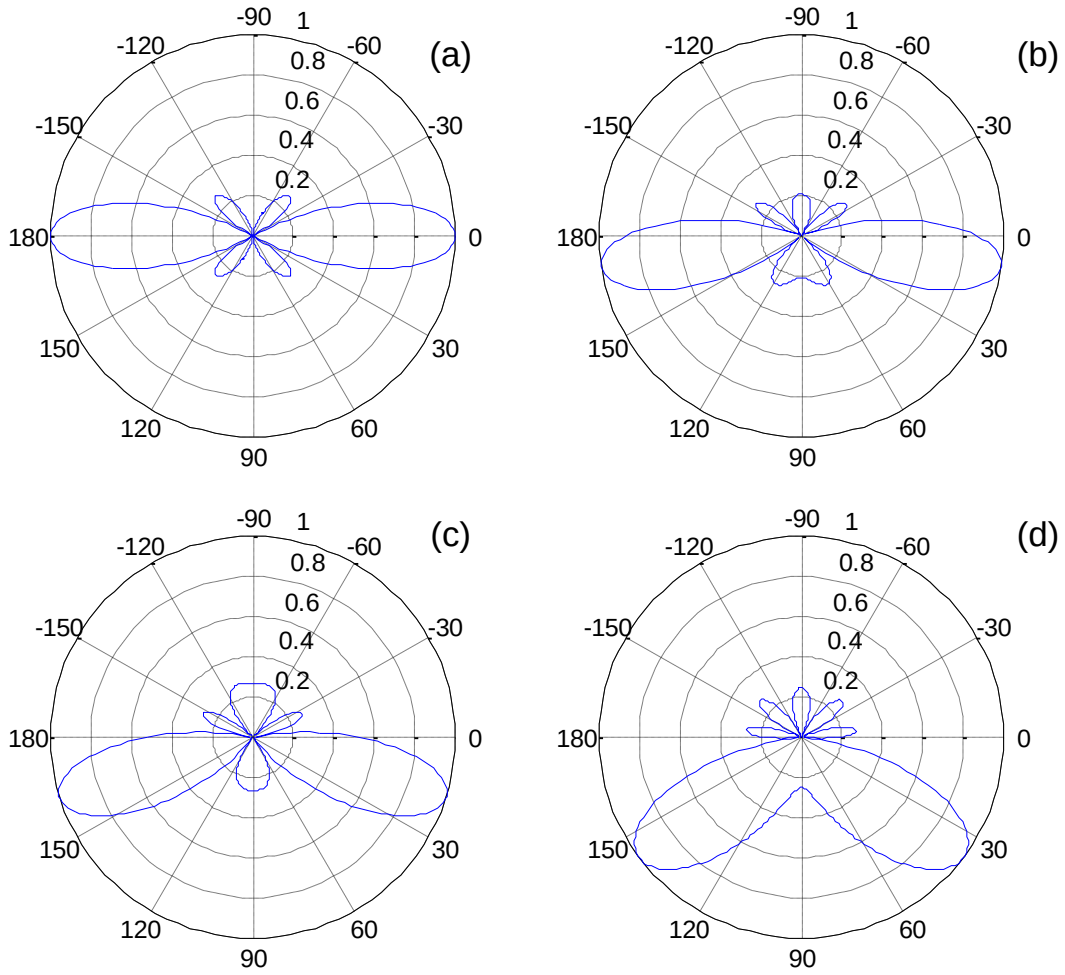


Fig. 8.7 Simulated radiation pattern of a four-element linear PAA with an element spacing of 5 m. The feed signals to the antenna elements experience time delay of 2.5 ns per round trip in our true-time delay system. The PAA initially points at 0°. (a)-(d) correspond to the radiation pattern when the feed signal recirculates for 0, 1, 2, and 4 round trips.

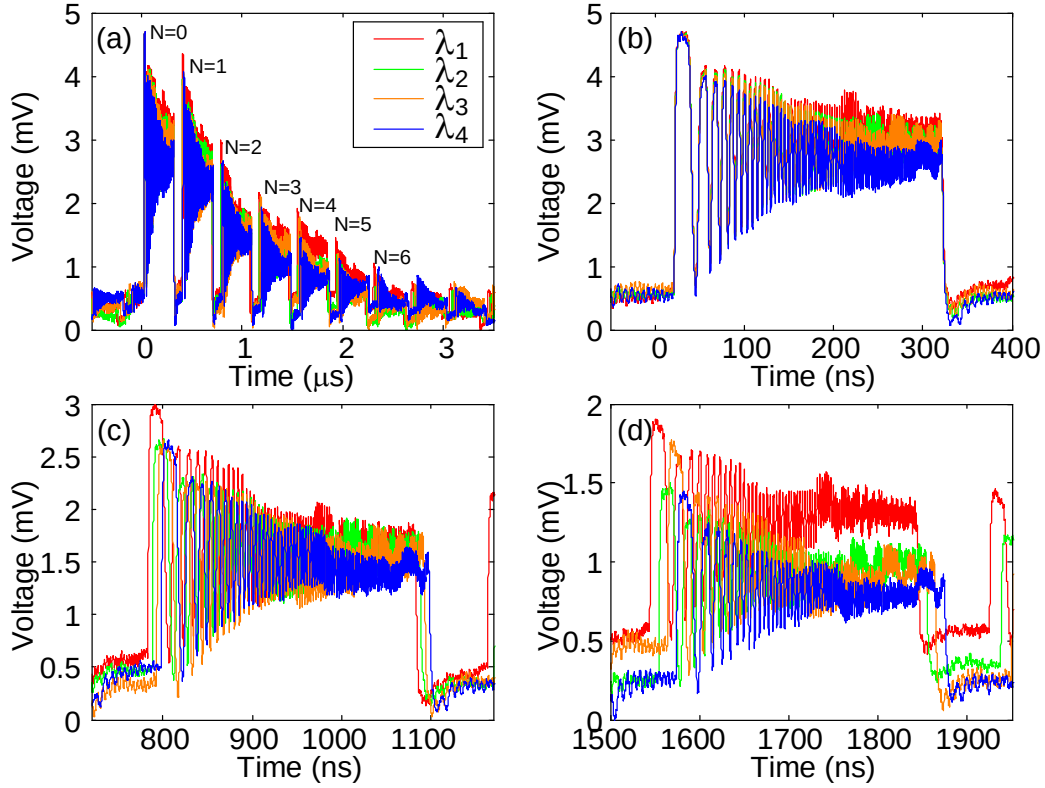


Fig. 8.8 Measured signals at the outputs of the four PDs when an LCMW is applied to the MZM. (a) The generated time delayed replicas and the zoom-in view of the signals for (b) $N=0$, (c) $N=2$ and (d) $N=4$.

For a given beam steering angle, different amount of time delays are required for PAAs with different antenna unit spacing. In our system, the true-time delay per round trip can be easily changed by using LCFBGs with different dispersion coefficients or TLSs with different wavelength spacing, according to (8-2). Here, an LCFBG with a dispersion coefficient of 200 ps/nm is employed, and the wavelength spacing between the TLSs are set to be 0.8 nm. A true-time delay of 160 ps per round trip should be expected. Fig. 8.9 shows the measured time delayed signals when a 1-ns electrical pulse is modulated to the multi-wavelength carrier. The measured true-time delay is 159.2 ps per round trip. Fig. 8.10 shows the radiation pattern of a four-element linear PAA when the time delayed signals are used as the inputs for the antenna elements. Note that in this simulation, an initial time delay was applied so that the beam is

pointing at -28.5° . As the feed signal recirculate in the loop for 0, 1, 2 and 4 round trips, the beam is steered to -13.9° , 0° and 28.5° . The PAA can cover a steering range from -90° to 90° if the dispersive loop allows the optical signal to recirculate for over nine round trips. Assume the signal recirculates in the loop for less than 10 round trips, the bandwidth limit of the system can be calculated to be 5.59 GHz according to (8-4), which is significantly improved compared to the case when a LCFBG with a dispersion coefficient of 2500 ps/nm is used. It can be seen that, in order to get broadband true-time delay system with a large time delay, the best solution is to use a LCFBG with small dispersion and TLSs with large wavelength spacing.

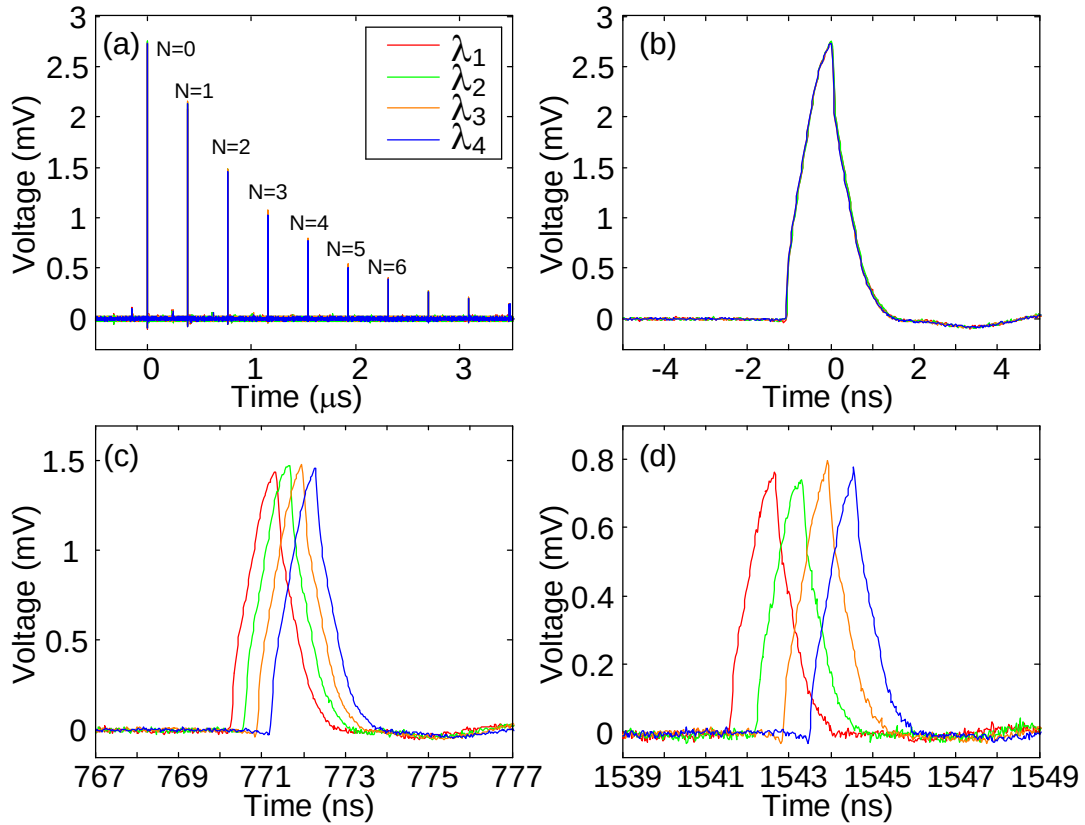


Fig. 8.9 Measured signals at the outputs of the four PDs with a small true-time delay step of 160 ps, and with an electrical pulse as the feed microwave signal. (a) The generated time delayed replicas and the zoom-in view of the signals for (b) $N=0$, (c) $N=2$ and (d) $N=4$.

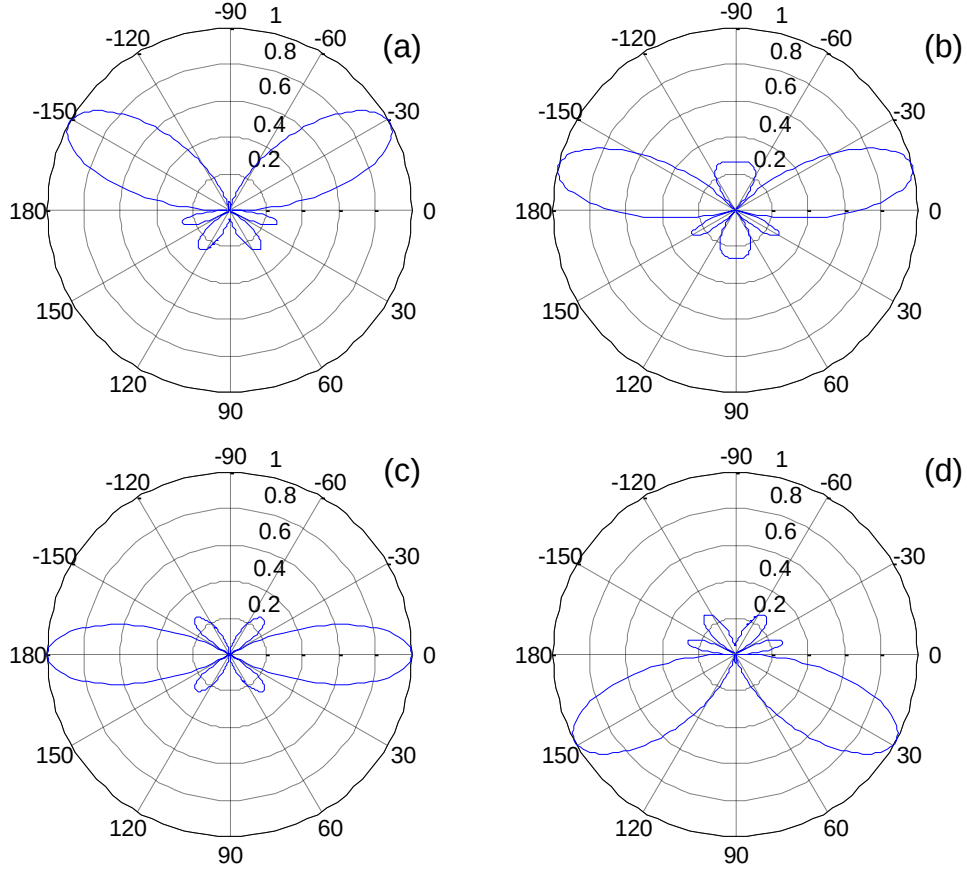


Fig. 8.10 Simulated radiation pattern of a four-element linear PAA with an element spacing of 20 cm. The feed signals to the antenna elements experience time delay of 159.2 ps per round trip in our true-time delay system. The PAA initially points at -28.5° . (a)-(d) correspond to the radiation pattern when the feed signal recirculates for 0, 1, 2, and 4 round trips.

8.4 Discussion and Conclusion

The proposed beamforming system can be used to achieve tunable time delays for a pulsed microwave signal. The pulse duration cannot be longer than the round trip time of the dispersive loop to avoid overlap of two consecutive pulses. For a microwave signal with longer temporal duration, the loop length should be longer. A longer dispersive loop will lead to a slower beam steering speed since the steering speed is limited by the overall round trip time. For our proposed

system, the total time for 10 round trips is 3.8 μs , which is fast enough for most radar applications [148].

Note that the system can generate true time delays for a PAA with non-uniform element spacing, which can be simply done by using a laser array with non-uniform wavelength spacing. It should also be noted that the concept demonstrated here using a recirculating dispersive loop can be applied for many other applications where a large tunable time delay is needed. Such a system is also adaptable to the ripples in the group delay response of the LCFBG used in the dispersive loop. As light recirculate in the loop, the ripples in the group delay response of the LCFBG will start to affect the amount for time delay more significantly. This effect can be mitigate by slightly adjusting the lasing wavelengths of the LDs.

In conclusion, we have proposed and experimentally demonstrated a photonic true-time delay beamforming network based on a switch-controlled wavelength-dependent recirculating dispersive loop. The microwave signal, with a temporal duration less than the round trip time of the dispersive loop, to be radiated to free space was modulated on the optical carriers of a laser array with fixed wavelengths. When the microwave-modulated optical signals recirculate in the dispersive loop, due to the wavelength-dependent time delays in the LCFBG, tunable progressive time delays could be achieved, which would lead to the beam steering when the time-delayed microwave signals are fed to a PAA. It is different from some previously reported techniques where the time delay tuning was realized by tuning the wavelength spacing of a laser array or the dispersion coefficient of an LCFBG, the system here is simpler since a laser array with fixed wavelengths is needed and the progressive time delays are generated and tuned by controlling the number of the round trips of the optical signals recirculating in the dispersive loop. The proposed

true-time delay system was experimentally evaluated. In the experiment, an optical coupler instead of an optical switch was used to study the tunable time delay behavior of the system. Two true-time delay beamforming networks using two dispersive loops incorporating two different LCFBGs with dispersion coefficients of 2500 ps/nm and 200 ps/nm to achieve a time delay difference of 2.5 ns and 160 ps per round trip were demonstrated. The key advantage of the proposed approach is its simplicity for implementation since no TLS and tunable LCFBG are needed.

CHAPTER 9 SUMMARY AND FUTURE WORK

9.1 Summary

The work in this thesis focused on the use of an LCFBG-based DDL for the processing of broadband microwave signals.

1) A broadband and precise microwave time reversal system using an LCFBG-based DDL was proposed and experimentally demonstrated. By using the DDL that has opposite dispersion coefficient when reflecting light at two different ends, a wideband microwave waveform modulated on an optical pulse can be temporally reversed after the optical pulse is reflected by the LCFBG for three times thanks to the opposite dispersion coefficient of the LCFBG when the optical pulse is reflected from the opposite end. The time reversal of three different microwave waveforms with a bandwidth of 4 GHz and a time duration of about 10 ns was demonstrated.

2) A microwave photonic matched filter was implemented for simultaneous generating and compressing an arbitrary microwave waveform. The proposed microwave photonic signal processor was verified by two experiments, in which a 7-bit PCMW with a carrier frequency of 4.08 GHz, and an LCMW with a bandwidth of 7.7 GHz were generated and compressed. Compression ratios of 20.6 and 9.3 were achieved for the two waveforms. The proposed microwave photonic signal processor can find applications in radar systems to generate and compress wideband and high speed microwave signals.

3) Based on the microwave time reversal using an LCFBG-based DDL, a temporal convolution system for the calculation of real time convolution of two wideband microwave

signals was demonstrated, which was realized by three photonic subsystems to perform the time reversal, signal multiplication, and integration. The proposed approach was experimentally evaluated, in which the calculations of three temporal convolutions between two rectangular waveforms, between an inverse sawtooth waveform and a rectangular waveform, and between an arbitrary waveform and a short pulse were experimentally demonstrated.

4) A novel time-stretched sampling system with a large stretching factor has been demonstrated by a repetitive use of an LCFBG in a recirculating dispersive loop. An equivalent GDD as large as 12000 ps/nm with a large stretching factor of 36 was achieved, which is the largest dispersion coefficient ever reported in an optical DDL.

5) An approach to the generation of an LCMW with an extended temporal duration implemented by an FPI for spectral shaping and a dispersive loop for wavelength-to-time mapping was proposed and experimentally demonstrated. Long temporal duration for the generated LCMW was enabled by multi-time use of an LCFBG in a dispersive to perform WTT mapping. Two LCMWs with two temporal widths of 25 and 42 ns were generated at two different frequency bands.

6) A fiber-optic true-time delay beamforming network using a switch-controlled recirculating wavelength-dependent dispersive loop incorporating an LCFBG was demonstrated. The tuning of the time delays are realized by controlling the number of round trips the optical signals recirculate in the loop, which is done by using a 2×2 optical switch to direct the optical signals back to the loop for additional time delays or output of the loop. A four-channel true-time delay beamforming network using two different recirculating dispersive loops with a true-time delay of 2.5 ns and 160 ps per round trip incorporating two LCFBGs with different dispersion

coefficients are demonstrated. The use of the beamforming network to achieve arrayed beamforming that can cover -90° to 90° is demonstrated.

The photonic systems we demonstrated in this thesis have achieved an operation bandwidth from several GHz to up to the THz range, and hence can be used to overcome the bandwidth bottleneck of the digital signal processing techniques that are widely adopted in modern microwave communication and radar systems.

9.2 Future work

In this thesis, we have developed several photonic systems based on optical DDLs that can satisfy the bandwidth requirement of most modern microwave communication and radar systems. However, the performance of some of the systems needs to be further improved. For example, incoherent optical broadband source were used as the optical carrier source for the matched filter. The incoherent source introduces a large amount of noise at the photodetector, which would have negative influence on the overall performance of the system. A possible solution is to use a wideband coherent optical source, such as a laser array, to generate an optical comb. When an optical comb is used as the optical carrier, much lower noise will be generated. The interference signal caused by beating of the optical comb will be at high frequency determined by the spacing between the comb lines and can be removed by a low pass filter.

Another required improvement is with the fiber recirculating loop, which has been used for microwave true-time delay beam forming, time stretched sampling and large TBWP generation. Due to the existence of a gain medium in the loop, lasing occurs when the gain is larger than the round trip loss, which would generate strong interference signal at a photodetector. It is very

important to find a solution to mitigate this effect. In our future work, we will try to use an actively controlled optical switch to allow a pulse to recirculate in the loop for a certain number of round trips while suppressing the lasing at the same time.

Finally, the photonic systems shown in the work are based on fiber optics. In the past years, the development of silicon photonics, group III-V integrated photonics, and the integrated photonics based on silicon nitride platform has enabled the implementation of such systems on a chip. Group III-V integrated photonics will be our first choice as it provides optical source and gain for the systems we demonstrated. The optical gain is especially important when a long waveguide is required for a large time delay. For silicon and silicon nitride platforms, the implementation of such a system requires hybrid integration with the group III-V components as only passive components can be fabricated with silicon or silicon nitride. The advantage of the silicon and silicon nitride platforms is the low waveguide loss, which allows an on-chip time delay as large as tens of nanoseconds. Our future work will be developing integrated microwave photonic systems to function as delay lines for broadband microwave signals and for the realization of various microwave signal processing functions.

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