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Wong, Jia haur

2012

Wong, J. H. (2012). Photonic analog-to-digital conversion and photonic microwave signal generation. Doctoral thesis, Nanyang Technological University, Singapore.

<https://hdl.handle.net/10356/54890>

<https://doi.org/10.32657/10356/54890>

Photonic Analog-to-Digital Conversion and Photonic Microwave Signal Generation

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A thesis submitted to the Nanyang Technological University
in fulfillment of the requirement for the degree of
Doctor of Philosophy

2012

Statement of Originality

I hereby certify that the content of this thesis is the result of original research done by me and has not been submitted for a higher degree to any other university or institute.

Date

Wong Jia Haur

ACKNOWLEDGEMENTS

I would like to express my deepest gratitude to the following people who have supported me during my PhD studies.

First and foremost, I am very fortunate to have Associate Professor Sheel Aditya as my main PhD supervisor. This thesis would not have been possible without his invaluable guidance and encouragement. I would also like to extend my thanks to my PhD co-supervisor, Professor Perry Shum Ping for offering me the opportunity to take up a PhD during my undergraduate studies.

I am also truly grateful to Dr. Lam Quoc Huy, Dr. Kenneth Lee Eng Kian, Dr. Lim Peng Huei, Dr. Zhou Junqiang and Mr. Wu Kan for their helpful input at technical discussions as well as their invaluable suggestions and comments on my work.

My gratitude also goes out to all members in Fibre Technology Laboratory for their support and assistance.

Last but not the least; I would like to thank my dearest family and my lovely partner Michelle Chiang Hui Ling for their understanding, support and encouragement along the way.

ABSTRACT

Driven by the increasing demand for speed and capacity in advanced signal-processing systems such as radar, surveillance, millimeter-wave access networks, and satellite communication, there has been a strong interest in developing analog-to-digital converters (ADC) with high sampling rate and resolution. It is recognized that the jitter-limited accuracy and low sampling rate issues commonly associated with conventional electronic ADCs can be overcome by either optically-sampled ADC or photonic time-stretched ADC (TS-ADC) systems.

Besides ADCs, electronic oscillators generating continuous-wave (CW) or/and arbitrary microwave waveforms also play a critical role in the aforementioned systems. The performance-bottlenecks imposed by the electronic circuits, however, restrict the highest frequency that can be generated and the largest bandwidth that can be achieved by the electronic oscillators. Various photonic techniques that exploit the unique and advantageous features of photonic components that include mode-locked lasers have been developed to circumvent the aforementioned restrictions.

In this thesis, novel schemes for photonic analog-to-digital (A/D) conversion and photonic microwave signal generation have been reported. In the first part of the thesis, we first demonstrate a 2 Giga samples per second (GS/s) optically-sampled ADC system incorporating complementary dual-output modulation and balanced photodetection, subsequently implementing a linearization process on the digitized data for odd-order harmonics suppression. An effective number of bits (ENOB) of ~ 5.34 -bits was obtained for this system. Following this, a photonic TS-ADC system amenable to continuous-time operation based on polarization modulation and balanced photodetection is investigated. Demonstrating with a single channel, we

have obtained a sampling rate of 2 GS/s and ENOB of ~ 3.56 bits for this system. We end the first part of the thesis with a demonstration of a technique which enables the generation of a 10x2 GHz time- and wavelength- interleaved pulse-train (TWIPT) with pulsewidth tunability that is suitable for a 20 GS/s optically-sampled ADC adopting the wavelength-division-demultiplexing (WDD) parallel processing approach for increasing the sampling rate of the electronic ADCs.

The second part of the thesis is devoted to the development of photonic techniques which can generate frequency-tunable CW microwave signals and linearly-chirped microwave signals with wide chirp tuning range. We first demonstrate two techniques which are able to generate frequency-tunable CW microwave signals with large suppression ratio of unwanted frequency components. In both of these demonstrations, we show the generation of stable tunable microwave signals with the frequencies ranging from 4 GHz to 14 GHz. Subsequently, we demonstrate an approach capable of generating microwave waveforms with tunable centre frequencies and chirp rates covering the negative, zero and positive values. In essence, a chirp rate ranging from ~ -126.7 GHz/ns to $\sim +120.8$ is obtained in the experiment.

It should be emphasized that all the techniques demonstrated in this thesis make use of mode-locked lasers, thus rendering it a common component between the two topics studied here. With this common component, there is a good potential to integrate the techniques presented here on a common platform.

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ACRONYMS

A

A/D	Analog-to-Digital
ADC	Analog-to-Digital Converter
ASE	Amplified spontaneous emission
AWG	Arbitrary Waveform Generator

C

CW	Continuous-Wave

D

DAC	Digital-to-Analog Converter
DC	Direct-Current
DCF	Dispersion Compensated Fiber
DFT	Discrete Fourier Transform
DGD	Differential Group Delay
DRA	Distributed Raman Amplification
DST	Direct-Space-to-Time

E

EDF	Erbium Doped Fiber
EDFA	Erbium Doped Fiber Amplifiers
ENOB	Effective Number of Bits
EPS-EFD	Equivalent-Phase-Shift Fiber-Bragg Gratings

F

FBG	Fiber-Bragg Gratings
FD-POP	Fourier-domain programmable optical processor
FFT	Fast-Fourier Transform
FM	Frequency-Modulated
FSR	Free Spectral Range
FWHM	Full-Width-Half-Maximum

G

GDD	Group Delay Dispersion
GPS	Global Positioning System
GVD	Group-Velocity-Dispersion

H

HNLF	Highly Nonlinear Fiber

L

LCFBG	Linearly-Chirped Fiber-Bragg Gratings

M

MZI	Mach-Zehnder Interferometer
MZM	Mach-Zehnder Modulator

O

O/E	Optical-to-Electrical
OSA	Optical Spectrum Analyzer
OTDD	Optical-Time-Division-Demultiplexing
OTDL	Optical-Tunable-Delay-Line

P

PBS	Polarization-Beam-Splitter
PD	Photodetector
PM	Polarization-Maintaining
PMD	Polarization-Mode Dispersion
PolM	Polarization Modulator

R

RF	Radio-Frequency
RIN	Relative Intensity Noise
RMS	Root-Mean-Square

S

SC	Supercontinuum
SCDM	Single-Channel Dispersion Module
SFDR	Spurious-Free Dynamic Range
SINAD	Signal-to-Noise Plus Distortion Ratio
SLM	Spatial Light Modulator
SMF	Single-Mode-Fiber
SNR	Signal-to-Noise Ratio
SOA	Semiconductor Optical Amplifier
SSB	Single Sideband

T

TE	Transverse Electric
THD	Total Harmonic Distortion
TM	Transverse Magnetic
TS-ADC	Time-Stretched Analog-to-Digital Converter
TWIPT	Time- and Wavelength-Interleaved Pulse-Train

V

VOA	Variable Optical Attenuators

W

WDD	Wavelength-Division-Demultiplexing
WDM	Wavelength-Division-Multiplexer

SYMBOLS

$J_n(k)$	n^{th} order Bessel function of the first kind
β_2	2 nd -order GVD of dispersive medium
β_3	3 rd -order GVD of dispersive medium
A_n	Amplitude for the n^{th} channel in MCDM
$\Delta\omega_{sp}$	Angular channel spacing between adjacent channels of MCDM
ω_c	Angular frequency of optical carrier
ω_x	Angular frequency of optical wave ($x = 1, 2, 3, \dots$)
ω_{RF}	Angular frequency of RF signal
$\Delta\omega_{sp}$	Angular frequency spacing
$V_{in}(t)$	Applied input voltage in time domain
B	Bandwidth of PD
$\Delta\omega$	Bandwidth of pulse-train in frequency
$\Delta\lambda$	Bandwidth of pulse-train in wavelength
k_B	Boltzmann's constant
ω_o	Centre angular frequency of the spectrum for a pulse-train
f_{centre}	Centre frequency of microwave waveform
λ_x	Center wavelength of optical pulse/wave ($x = 1, 2, 3, \dots$)
Δf_{sp}	Channel spacing between adjacent channels of MCDM
C	Chirp parameter
ϕ_o	Constant phase shift determined by DC bias voltage
$\otimes$	Convolution
V_{DC}	DC voltage used to bias MZM
$\delta(t)$	Delta function
V_{full}	Detected voltage corresponding to P_o .
L_D	Dispersion length
D_F	Dispersion of LCFBG

D_M	Dispersion of MCDM
D_S	Dispersion of SCDM or SMF
D_λ	Dispersion parameter
ϕ_{DIP}	Dispersion-induced phase
$\sqrt{G_{eff}}$	Effective gain of RF link
R_{eff}	Effective thermal resistance
$E_x(t)$	Electric field of optical pulse/wave in time domain ($x = 1, 2, 3, \dots$)
$E_x(\omega)$	Electric field of optical pulse/wave in frequency domain ($x = 1, 2, 3, \dots$)
$ENOB$	ENOB of a non-ideal ADC
V_b	Externally applied bias voltage to PolM
f_{input}	Frequency of analog input signal
f_{RF}	Frequency of RF signal
Δf_{sp}	Frequency spacing
Δf_{approx}	Frequency spacing approximated to nearest integer multiple of f_{rep}
V_{FS}	Full-scale voltage of ADC
f_o	Fundamental frequency of RF signal
f_{FWHM}	FWHM bandwidth in frequency
τ_{FWHM}	FWHM pulsewidth
GDD	GDD of dispersive medium
GDD_F	GDD of LCFBG
GDD_M	GDD of MCDM
GDD_S	GDD of SCDM or SMF
V_π	Half-wave voltage of MZM
$\Delta\omega_{ch}$	Half-width bandwidth (at 1/e-intensity point) for a channel of MCDM in angular frequency

$f_{RF,inst}$	Instantaneous frequency of microwave waveform
L_x	Length of fiber ($x = 1, 2, 3, \dots$)
$ENOB_{jitter}$	Maximum achievable ENOB due to aperture jitter
$ENOB_{ambiguity}$	Maximum achievable ENOB due to comparator ambiguity
$ENOB_{thermal}$	Maximum achievable ENOB due to thermal noise
$ r(\omega) $	Magnitude response (reflection) of LCFBG
SNR_Q	Maximum signal-to-noise ratio of an ideal ADC
m_{dep}	Modulation depth
ϕ_x	Optical phase of optical pulse/wave ($x = 1, 2, 3, \dots$)
Z_{out}	Output impedance
P_{out}	Output optical power of optical signal
$V_{out}(t)$	Output voltage in time domain
E_o	Peak amplitude of optical pulse/wave
P_o	Peak optical power of optical signal
V_o	Peak voltage of RF signal
$V_{in,pp}$	Peak-to-peak voltage of input RF signal
θ_e	Phase error associated with deviation from MZM's quadrature point
$I(t)$	Photocurrent in time domain
$I_{Env}(t)$	Photocurrent in the absence of modulation
D_p	PMD parameter
ϕ_b	PolM phase bias
τ_o	Pulse half-width related to the FWHM pulsewidth
Q_e	Quantization error of ADC
$V_{out}(t_n)$	Quantized output voltage at time instant t_n
$V_{in}(t_n)$	Recovered input voltage at time instant t_n
f_{rep}	Repetition frequency of pulse-train

T	Repetition period of pulse-train
ΔR	Responsivity imbalance of balanced PD
R	Responsivity of PD
$R_{\pm}$	Responsivity of the individual photodiode of balanced PD
$P_{rf}(\omega_{RF})$	RF power of RF signal in frequency domain
τ_a	RMS aperture jitter
$Q_{e,RMS}$	RMS quantization error of ADC
$f_{sampling}$	Sampling frequency of ADC
$SFDR$	SFDR
$SFDR_{bits}$	SFDR expressed in bits
$E(t)$	Shape of optical pulse in time domain
$SINAD$	SINAD
V_{LSB}	Smallest possible change in the digital output voltage of ADC
SNR	SNR
SNR_{bits}	SNR expressed in bits
ϕ_s	Static bias phase-shift caused by PC on PolM
T_k	Temperature in °K (Kelvin)
Δt_{delay}	Time-delay between adjacent pulses
M	Time-stretched factor
Δt_{tj}	Timing jitter of pulse-train
τ_{jitter}	Total system timing jitter
Δt	Tunable time-delay between arms of MZI
λ_c	Wavelength of optical carrier
$\Delta\lambda_{sp}$	Wavelength spacing

Chapter 1 INTRODUCTION

An active research area for the past two decades, Microwave Photonics is the study of novel photonic techniques which leverages the unique properties of photonic components, such as low loss, broad bandwidth, immunity to electromagnetic interference and high rate-distance product to overcome inherent performance-bottlenecks imposed by conventional electronic methods utilized in diverse and sophisticated radio-frequency (RF) systems [1-3]. In microwave photonics, recently much emphasis has been laid upon developing photonic solutions for applications such as analog-to-digital (A/D) conversion, microwave signal generation, instantaneous microwave frequency measurement, microwave filters, phased array antennas etc. This is evident from the growing number of publications concerning these applications.

Among the abovementioned applications, there has been a particularly strong interest in the topics of photonic A/D conversion and photonic generation of microwave signals due to their significant potential application in systems such as radar (radio detection and ranging), surveillance, wireless access networks, and satellite communication. In the context of radar systems, be it continuous-wave (CW), frequency-modulated-CW (FM-CW) or pulsed radar, their architecture generally comprises a similar set of components and subsystems, which include antennas, duplexer, electronic oscillator, arbitrary waveform generator, transmitter, receiver, and signal processing subsystem etc. It has been shown that optical-sampling using low jitter optical pulses from mode-locked lasers can tackle the well-known jitter-limited resolution issue associated with the electronic analog-to-digital converters (ADCs) which serve as an interface between the receiver and the signal processing

subsystems. In a similar manner, various photonic techniques have also been demonstrated to generate CW, RF-pulsed and chirped-pulsed microwave signals with much larger bandwidth, wider frequency-tunability and better stability as compared to conventional electronic oscillator and arbitrary waveform generator. Interestingly, optical pulses from mode-locked lasers are also utilized in some of the photonic techniques for generation of microwave frequencies.

With photonic techniques for both A/D conversion and microwave frequency generation proving to be viable options to conventional electronic methods, these seemingly different techniques can be implemented together in the same system to improve the system level performance significantly. Possible integration of these photonic techniques through the use of photonic integrated circuits would further benefit advanced systems, such as the aforementioned radar systems, very substantially. One of the key aspects that can prove helpful in the photonic integration is the use of mode-locked lasers in both A/D conversion and microwave frequency generation. These considerations provide the common thread between the two topics of photonic A/D conversion and photonic generation of microwave signals that are considered in this thesis

In this chapter, we first introduce the topic of A/D conversion, describing the basic working principle as well as the main factors limiting the performance of ADCs. This helps to pave the way to understanding the motivation for work on photonic A/D conversion. Next, we briefly review the previous work on photonic ADCs, with an emphasis on optically-sampled ADC and photonic time-stretched ADC (TS-ADC) since these two techniques have been found quite promising.

Following this, we introduce the topic of generation of microwave signals as well as review a number of photonic approaches which can generate frequency-tunable CW microwave signals. Next, we discuss the photonic generation of chirped microwave signals with tunable centre frequency as well as chirp rate.

This chapter ends with a description of the main contributions and the outline of the dissertation.

1.1 Analog-to-digital conversion

Real life signals are analog in nature - varying in time and amplitude in a continuous manner. Characterizing a certain real-world phenomenon, an analog signal has to be translated into a digital representation before it can be manipulated in the digital domain for the purpose of information processing, data transmission, and control systems [4]. Key to this translation process is the electronic A/D conversion, which serves as the interface between the received analog signal and the desired discrete-time, binary-coded form. Performing two basic functions on the analog signal, an ADC first samples and holds a time-varying voltage for a specified time and subsequently quantizes the held voltage into a number of discrete levels.

In general, the sampling process is determined by the ADC's sampling rate f_{sample} , while there are a total of 2^N equally-spaced distinct voltage levels for an ADC with N -bits. Fig. 1.1 illustrates the A/D conversion of a 50 megahertz (MHz) sine wave based on an ideal 3-bit ADC which has a maximum sampling rate of 1 GS/s. As shown in Fig. 1.1, the quantization process introduces an inevitable error between the actual analog input voltage and the value of the digital output. This error is known as the quantization error Q_e and varies between $\pm 0.5V_{\text{LSB}}$, where V_{LSB} (LSB = least

Fig. 1.1 Illustration of the analog-to-digital conversion process obtained by an ideal 3-bit ADC. (a) Sampling and (b) Sampling and quantization.

significant bit) defines the smallest possible change in the digital output voltage.

Specifically, V_{LSB} can be calculated as

$$V_{LSB} = V_{FS} / 2^N \quad (1.1)$$

where V_{FS} is the full-scale peak-to-peak voltage of the ADC.

The quantization error is the solitary noise source in an ideal ADC and its root-mean-square (RMS) value can be calculated to be

$$Q_{e,RMS} = \left[\frac{1}{V_{LSB}} \int_{-0.5V_{LSB}}^{0.5V_{LSB}} (x)^2 dx \right]^{0.5} = \frac{V_{LSB}}{\sqrt{12}} \quad (1.2)$$

where x is the difference between the input and the output voltage. Theoretically, the maximum signal-to-noise ratio of an ideal ADC (SNR_Q) which represents the ratio of the largest RMS input signal amplitude to the $Q_{e,RMS}$ can thus be expressed in dB (decibel) as

$$SNR_Q = 20 \log \left(\frac{V_{pp} / 2\sqrt{2}}{Q_{e,RMS}} \right) = 20 \log \left(\frac{2^N V_{LSB} / 2\sqrt{2}}{Q_{e,RMS}} \right) \quad (1.3)$$

Rewriting in terms of N , SNR_Q is given as

$$\begin{aligned} SNR_Q &= 20N \log(2) + 20 \log \sqrt{12} - 20 \log (2\sqrt{2}) \\ &= 6.02N + 1.76 \end{aligned} \quad (1.4)$$

For a non-ideal ADC, however, additional sources of noise and nonlinear distortions combine and contribute to system performance and these sources limit the effective resolution of the ADC. The ENOB is therefore used to describe the effective resolution of a non-ideal ADC and is defined as

$$ENOB = \frac{(SINAD - 1.76)}{6.02} \quad (1.5)$$

where the signal-to-noise plus distortion ratio (SINAD), measured in dB, defines the ratio of the RMS input signal amplitude to the RMS sum of all noise and distortion, excluding direct-current (DC), within the Nyquist bandwidth of the ADC. For instance, if a 16-bit ADC yields a SINAD of 88 dB, which can be measured by performing Discrete Fourier Transform (DFT) on the digitized data, then the effective resolution of the ADC would be

$$ENOB = \frac{(88 - 1.76)}{6.02} = 14.32 \text{ bits} \quad (1.6)$$

In other words, due to the combined effects of different error sources (including quantization error), the 16-bit ADC would be producing a resolution equivalent to that of a 14-bit ADC. Consequently, the 16-bit ADC can only effectively support up to 14-bits applications.

Besides ENOB and SINAD, three other important metrics which are frequently used to quantify the performance of a non-ideal ADC are signal-to-noise ratio (SNR), spurious-free dynamic range (SFDR), and total harmonic distortion (THD). All the aforementioned performance metrics can be measured by analyzing the Fast-Fourier Transform (FFT) of the digital output of the ADC.

Unlike SNR_Q , SNR is calculated from the FFT data as the ratio of the power at the fundamental frequency f_o to the power of all sources of noise and non-harmonic spurs within the Nyquist bandwidth. On the other hand, SFDR is the ratio of the power at f_o to the power of the largest spurious signal within the Nyquist bandwidth. The SFDR is typically specified in dBc (decibels relative to the carrier) or dBFS (decibels relative to the ADC's full scale). Lastly, THD is the ratio of the power at f_o to the power of all the harmonics ($2f_o, 3f_o, 4f_o \dots$). Similar to how SINAD relates to ENOB, SNR and SFDR can also be expressed in terms of bits, as defined by the following

$$SNR_{bits} = \frac{SNR - 1.76}{6.02} \quad (1.7)$$

$$SFDR_{bits} = \frac{SFDR}{6.02} \quad (1.8)$$

1.1.1 Limits from Walden Chart

Over the last few decades, ADCs have seen and enjoyed rapid progress as attested by the availability of a diverse selection of converters in terms of architectures, resolution and sampling rates. Despite this progress, ADCs seem to be improving slower than the digital electronics which too has experienced a remarkable

growth in terms of speed and storage capacity during the same period. As a result, these indispensable components are found to be key performance-bottlenecks in advanced signal-processing systems, inevitably causing these systems to advance at a slower pace too [5, 6].

Fig. 1.2 illustrates the "Walden Chart", which is a summary of the findings compiled from the comprehensive survey of the ADC field by Robert Walden [5]. Based on various experimental and commercially available ADCs, the "Walden Chart" indicates a distribution of ENOB versus the analog signal bandwidth. Clearly, the chart reveals a trade-off between the ENOB and the signal bandwidth [7]. Aside from quantization noise, Walden pointed out further that the SNR of the surveyed ADCs is limited by three other mechanisms, namely thermal noise, comparator ambiguity and aperture jitter.

Fig. 1.2 Effective number of bits (ENOB) of ADCs versus signal bandwidth. The red circles correspond to the ADCs surveyed by Walden. The lines indicate the limitations imposed by thermal noise, comparator ambiguity, aperture jitter and Heisenberg principle. The figure is reproduced from [5].

The first mechanism, "thermal noise" is considered to include all possible noise sources such as thermal noise, shot noise, $1/f$ noise and input-referred noise. The second mechanism, "comparator ambiguity" results from the finite speed with which the comparator of an ADC responds to small variations in the input signal to produce the correct quantization value. Lastly, "aperture jitter" is due to the sample-to-sample timing variation from the input sample-to-hold switch of an ADC. In other words, the sampling interval is not uniform and it varies randomly. The maximum achievable resolution due to each of the individual mechanisms can be calculated as [5]

Thermal noise

$$ENOB_{thermal} = \log_2 \left(\frac{V_{FS}^2}{6k_B T_k R_{eff} f_{samp}} \right)^{0.5} - 1 \quad (1.9)$$

Comparator ambiguity

$$ENOB_{ambiguity} = \left(\frac{\pi f_{input} T_k}{6.93 f_{samp}} \right) - 1.1 \quad (1.10)$$

Aperture jitter

$$ENOB_{jitter} = \log_2 \left(\frac{2}{\sqrt{3} (\pi f_{samp} \tau_a)} \right) - 1 \quad (1.11)$$

where V_{FS} is the full-scale voltage, k_B is the Boltzmann's constant, T_k is temperature in °K (Kelvin), R_{eff} is the effective thermal resistance, f_{samp} is the sampling frequency, τ_a is the RMS aperture jitter and f_{input} is the frequency of the analog input signal.

The dashed lines in Fig. 1.2 illustrate the limits imposed on conversion accuracy by the aforementioned mechanisms. It can be observed that for the frequencies below ~ 7 MHz, the ENOB appears to be limited by the thermal noise, while the ENOB is limited by the aperture jitter for the frequencies between ~ 7 MHz and ~ 4 GHz. Lastly, the comparator ambiguity is the limiting factor for the ENOB when the frequencies greater than ~ 4 GHz are concerned. More importantly, we observe that ENOB drops by approximately one bit for every doubling of the frequency of the analog input, or -1 bit/octave. The primary contributor to this downward trend is the aperture jitter.

1.1.2 Effects of sampling-clock jitter on ENOB

A sampling-clock, generally derived from an RF oscillator, is required to trigger any state-of-the-art ADC so that the ADC samples the input signal at precisely equal time intervals. In reality, however, sampling takes place at non-uniform intervals as a result of the sampling-clock jitter from the RF oscillators. An artifact of clock variation from cycle-to-cycle, the sampling-clock jitter is often caused by the thermal noise and $1/f$ noise from the active and passive devices which make up the RF oscillators. Similar to the effects of aperture jitter, the effects of sampling-clock jitter will produce exactly the same type of error which will affect the SNR and thus limits the ENOB of an ADC.

As illustrated in Fig. 1.3, a corresponding voltage error will result due to the effects of the aperture and sampling-clock jitter. For any given aperture or sampling-clock jitter, an increase in the frequency of the analog input signal f_{input} , or slew rate, magnifies the associated output error. Since both the aperture τ_a and sampling-clock

Fig. 1.3 Effects of aperture jitter or sampling-clock jitter for analog input with different frequencies.

jitter τ_{clk} have the same effects on the output error and are uncorrelated, the total system timing jitter τ_{jitter} of an ADC is therefore the RMS of these two jitters and can be expressed as

$$\tau_{jitter} = \sqrt{\tau_a^2 + \tau_{clk}^2} \quad (1.12)$$

Considering only the effects of aperture and sampling-clock jitter, the SNR of an ADC is given as

$$SNR = -20 \log(2\pi f_{input} \tau_{jitter}) \quad (1.13)$$

Substituting the above SNR for the SINAD in (1.5), the theoretical ENOB due to the timing jitter can be calculated. Fig. 1.4 illustrates the relationship between ENOB and

Fig. 1.4 Theoretical ENOB of an ADC due to timing jitter versus analog input signal frequency.

the analog input signal frequency for different values of the timing jitter. For a given input frequency, higher ENOB can be achieved when the timing jitter is less. In practice, the ENOB is limited by the timing jitter as most ADCs and RF oscillators exhibit aperture and sampling-clock jitter [6, 7], typically in the order of hundreds of femtoseconds (fs), respectively. This is especially so at higher analog input signal frequencies, as described earlier in Fig. 1.2. Although state-of-the-art RF oscillators with sub-10 fs sampling-clock jitter are available at the present, these are often costly and bulky. Due to these factors, it is difficult to integrate these RF oscillators in chip-scale signal processing applications.

In this context, it has been demonstrated that the timing jitter of mode-locked lasers can be much lower than that for the typical RF oscillators. For instance, passive mode-locked lasers producing ultrashort pulses with 100-200 fs widths have been experimentally verified to have timing jitter as low as a few fs [8-12]. Utilizing such a low timing jitter mode-locked laser as the sampling-clock for an ADC can thus circumvent the jitter-limited ENOB issue imposed by the conventional RF oscillators.

Fig. 1.5 Major classes of photonic ADC.

Additionally, compared to the RF oscillators, the mode-locked lasers tend to be less costly, smaller in footprint and easier to integrate in chip-scale signal processing applications. These advantageous features of the mode-locked lasers, coupled with the unique properties of other photonic devices and components, have led to the development of various photonic approaches with the motivation to improve the performance of the existing ADC systems. This has given rise to investigations on "photonic ADC", and has attracted extensive research interest since the early 1970s [7].

1.2 Photonic analog-to-digital conversion

As discussed in the review work by George C. Valley, photonic ADCs can be categorized into four main classes, consisting of photonic-sampled, photonic-quantized, photonic-sampled and quantized, and photonic-assisted ADCs [7]. These major classes of photonic ADCs are shown in Fig. 1.5 and can be briefly described as follows:

1. Photonic-sampled ADCs sample the analog signal in the optical domain, while the quantization is performed in the electronic domain
2. Photonic-quantized ADCs are the exact opposite of the photonic-sampled ADCs
3. Photonic-sampled and quantized ADCs perform both the sampling and quantization in the optical domain.
4. Photonic-assisted ADCs utilize photonics to improve limiting properties of ADCs, with both the sampling and quantization processes performed in the electronic domain.

A detailed discussion on these four major classes of photonic ADCs can be found in [7], where various photonic techniques which have been developed over the years are reviewed. As the quantization process for the optically-sampled ADCs and the photonic TS-ADCs is accomplished by electronic ADC, we will be emphasizing these two classes of photonic ADCs in this thesis. An electronic ADC with a sampling rate of 2 GS/s and a stated resolution of 10 bits will be used in the experiments.

1.2.1 Optically-sampled ADCs

The concept of using optical pulses for optical sampling was first proposed by Taylor *et al.* [13-15] and this led to demonstrations of many other optically-sampled ADCs. Earlier work has mainly been based on a modulation scheme which utilizes a single-output Mach-Zehnder modulator (MZM) [16-21]. The SFDR of these optically-sampled ADCs tends to be limited by the 3rd-order distortions due to the nonlinear sinusoidal transfer function of the MZM. In later work, phase-encoded sampling scheme which is based on using a complementary dual-output MZM was used in conjunction with a digital linearization scheme in the form of arcsine operation to correct for the 3rd-order distortions generated by the MZM [22-24].

However, in most of these demonstrations, the number of photodetectors (PD) used is equal to the number of outputs which the optically-sampled ADC systems have. This is also the case in multi-channel interleaving optically-sampled ADCs [23, 25, 26].

In order to increase the sampling rate which is limited in electronic ADCs, two multi-channel interleaving schemes have been developed. The time-interleaving approach which is based on optical-time-division-demultiplexing (OTDD) was first recognized by Bell *et al.* [17, 18] and their demonstration provided the foundation for others to work on. Multiple channels with phased-encoded sampling were demonstrated and analyzed in [23, 25, 26], wherein they have shown to have increased the sampling rate of the electronic ADCs, maintaining a high resolution at the same time. However, this approach is often found to be limited by the response speed of the active optical switch which is used to distribute the high-speed optical pulse-train into parallel channels of lower speed. This consequently limits the sampling rate of the entire optically-sampled ADC.

An alternative to the time-interleaving approach, the wavelength-interleaving approach which is based on WDD was first proposed by Frankel *et al.* [21]. Recently, multiple wavelength channels with phased-encoded sampling have also been demonstrated [27]. Key to the wavelength-interleaving approach is the demultiplexing of a high-speed time- and wavelength-interleaved pulse-train (TWIPT) into different wavelength channels of lower speed using only passive components [28]. Thus, as compared to the OTDD approach, the sampling rate of the entire optically-sampled ADC is only limited by the number of wavelengths the TWIPT can have. The early approaches to obtaining a TWIPT are often difficult to implement as they require precise time-interleaving of multi-wavelength pulses together [29, 30]. Later, various other techniques were proposed and these delivered varying results [31-36]. In

particular, it has been shown that through the use of the time-lens compression technique [31, 37], a TWIPT can be generated without the need to manually provide uniform inter-pulse spacing for the multi-wavelength pulses. However, this technique involves the use of both a modulator and a microwave source.

Although optically-sampled ADCs have shown promise in circumventing the performance-bottlenecks imposed by electronic ADCs, there are still many shortcomings in some of the aforementioned techniques. Therefore, it is important for us to investigate new techniques which can help to improve the performance of these photonic ADCs.

In particular, to reduce the required number of PDs as compared to [22, 23, 38], we can implement a scheme which is based on a complementary dual-output MZM and balanced detection for an optically-sampled ADC system. As for the generation of a TWIPT for wavelength-interleaving optically-sampled ADC system, we can utilize a scheme based on dispersion spreading and sectional compression to help avoid the need to manually provide precise time-interleaving of multi-wavelength pulses together as well as the usage of any modulator and microwave source.

1.2.2 Photonic time-stretched ADCs

Proposed first by Jalali and coworkers [39-45], the concept of the photonic TS-ADC was developed with the motivation to enhance the effective input bandwidth and overall sampling rate of the available electronic ADCs. Essentially, a TS-ADC utilizes the properties of dispersion to stretch an input RF signal in time, thus slowing it down prior to digitization by a relatively slow sampling ADC. Early demonstrations were mainly on single-shot TS-ADC systems and involved the use of single-mode-fiber (SMF) [44, 46] as the dispersive medium to obtain a large stretch factor and in

return, a higher sampling rate. Later on, dispersion compensated fiber (DCF) which can provide a larger dispersion than SMF over the same length was utilized frequently in single-shot TS-ADC systems [47-50].

The TS-ADC is also implementable in continuous-time when the virtual time-gating principle is implemented. A proof-of-concept of this principle was first demonstrated by Han and Jalali in a 2 channel continuous-time TS-ADC system [51]. Single sideband (SSB) modulation schemes [52, 53] have also been demonstrated to mitigate the frequency-dependant dispersion penalty issue associated with TS-ADC systems. Recently, it has been demonstrated that the complementary dual-output MZM can also be utilized in a differential manner to reject the dispersion-induced even-order harmonic distortions inherent in TS-ADC systems [54]. The complementary pulses from the MZM outputs are deliberately time delayed by half a repetition period, to ensure that no temporal overlapping of the pulses occurs before combining and time-stretching by a single dispersive channel. This is conceptually simple, however, implementation is difficult due to the stringent requirement of providing precise time-delay between the complementary pulses from the dual-output MZM. Any amplitude and phase imbalances between the pulses will deteriorate the accuracy of the differential operation.

To avoid the practical implementation issues in [54], a polarization modulator (PolM) which supports the complementary pulses on the same transmission fiber can be utilized in a photonic TS-ADC system. The use of a PolM can eliminate the need to provide deliberate time-delay to the complementary pulses as well as avoid the bias-drifting issues which are usually associated with a typical MZM.

1.3 Generation of microwave signals

Microwave refers to electromagnetic waves with frequencies ranging from 300 MHz to 300 GHz, or equivalently, with wavelengths ranging from one meter to one millimeter (mm) in free space [55]. Important developments in microwave engineering occurred during the period of world war II, when it played a critical role in the development of microwave antennas as well as radar systems [56]. Since then, rapid progress has been made in the development of microwave semiconductor devices such as transistors and integrated circuits and this has led to a widespread use of microwave signals in many civilian applications today. These applications include wireless communication systems, global positioning system (GPS), navigation systems, air traffic control radars, broadband wireless access networks, and microwave ovens etc. [55, 56]. Conventionally, microwave signals are generated by electronic oscillators which are based on devices made from either III-V or SiGe semiconductors [57]. Apart from the signal generators which produce repetitive signals, usually in the form of sine or square waves, other forms of signals include the arbitrary waveforms and pulses.

Owing to the performance-bottlenecks imposed by the electronic circuits, these electronic oscillators have limitations in generating high frequency microwave signals. Although current state-of-the art electronic signal generators enable the generation of microwave signals up to frequencies as high as 40 GHz, they are complex to realize as well as costly. Limitations in the digital-to-analog converters (DAC) and timing jitter also cause arbitrary waveform generation to be restricted below 10 GHz [58]. These limitations in conventional electronic methods have spurred interest in finding alternative approaches to generate high frequency

Fig. 1.6 Summary of the various photonic techniques which can generate either CW microwave signals or microwave arbitrary waveforms.

microwave signals. Utilizing the high-speed and broad bandwidth features offered by photonic components, generation of high frequency microwave signals through photonic techniques has shown good promise. In contrast to the large transmission loss in coaxial cable, the low loss nature of optical fibers also supports remote delivery of the microwave signals generated through photonic means [1, 2].

1.4 Photonic generation of microwave signals

Over the last two decades, there has been considerable interest in developing novel photonic techniques to generate microwave signals with attributes such as high frequency, large frequency-tunability and low-phase noise. In general, photonic techniques have been developed to target the generation of CW microwave signals as well as microwave arbitrary waveforms. Various photonic techniques which are associated with the generation of these two types of microwave signals are summarized in Fig. 1.6.

Early photonic techniques to generate high frequency CW microwave signals include optical heterodyning [59-63], external modulation [64-70] and dual-wavelength laser [71-76]. As for arbitrary waveform generation, the popular photonic techniques include direct space-to-time (DST) pulse shaping [77-79], temporal pulse shaping [80-83], optical line-by-line pulse shaping [84-87], and lastly, optical spectral shaping and wavelength-to-time mapping [88-93]. A detailed discussion on the generation of CW microwave signals as well as generation of arbitrary waveform can be found in [1] and [94], respectively. In this dissertation, emphasis will be placed on the generation of CW microwave signals. We will also touch briefly upon the generation of chirped microwave signals, which are a special type of microwave arbitrary waveform.

1.4.1 Photonic generation of CW microwave signals

The optical heterodyning technique involves the beating of two independent tunable CW lasers at a photodetector (PD). Locking the phases of the CW lasers to a microwave reference source either by optical injection locking [59, 60] or by optical phase locked loop [61, 62] enables the generation of low-phase noise CW microwave signals. Frequency-tunability can be achieved by tuning the wavelength separation between the CW lasers. This scheme can easily generate signals up to mm-wave frequencies, but is typically limited by the bandwidth of the PD.

The external modulation scheme is based on the external modulation of the output of a CW laser via either an electro-optic intensity modulator [64] or a phase modulator [66]. Through the use of a low-phase noise electrical drive signal as a modulating signal, low-phase noise microwave signals can be generated. Frequency-tunability can be achieved by changing the frequency of the electrical drive signal, but

is limited by either the bandwidth of the PD or the modulator. It should be emphasized that, despite the simplicity of concept, in both of the above techniques, an external microwave reference source of high quality is essential.

The dual-wavelength laser approach is also based on the beating of two separated wavelengths at the PD, but the two lasing wavelengths coexist in a dual-wavelength laser. The phase correlation between them is better as compared to two free running CW lasers in the optical heterodyning approach. The key to this approach is a fiber-Bragg grating (FBG) which assists in the filtering of two lasing modes in a ring cavity laser [75, 76]. However, the fiber-Bragg gratings are either difficult and complex to fabricate or are expensive, when available commercially.

Besides the abovementioned rather conventional approaches, in recent years, another technique which is widely investigated is the direct photodetection of a passive or active mode-locked laser. This technique is based on the beating of equally-spaced and phase-locked longitudinal modes of the laser in a PD [95-98]. Using appropriate RF bandpass filtering, a single frequency microwave signal can be selected from the spectrum of a mode-locked laser. Since the beating modes in the PD are phase-locked, cancellation of phase fluctuations between the modes takes place, thus enabling the generation of low-phase noise CW microwave signals. Frequency-tunability in this scheme is attainable by incorporating an array of RF bandpass filters and switches for frequency selection.

Compared to all the other techniques, the use of mode-locked lasers to generate frequency-tunable and low-phase noise microwave signals is thus less complex. Therefore, it is important to develop new techniques to generate frequency-tunable CW microwave signals using mode-locked lasers.

1.4.2 Photonic generation of chirped microwave signals

Besides the generation of CW microwave signals, mode-locked lasers can also be used to generate microwave arbitrary waveforms, in particular, chirped microwave signals. Among the four widely-used techniques depicted earlier in Fig. 1.5, the optical spectral shaping and wavelength-to-time mapping technique has shown much flexibility in generating chirped microwave signals whose centre frequency and/or chirp rate can be tuned [90-93]. Moreover, this technique does not require any modulators, microwave sources, microwave filters and it offers higher potential for integration as compared to the other three techniques.

The optical spectral shaping and wavelength-to-time mapping technique is based on propagating an optical pulse-train into an optical spectral-shaper whose magnitude response has the same shape as that of the desired microwave waveform and subsequently propagating it into a dispersive medium where wavelength-to-time mapping is performed. Based on this fundamental concept, various approaches for the generation of chirped microwave signals have been developed over the years.

An early demonstration of the above technique involved the use of two linearly-chirped FBGs (LCFBGs) with different chirp rates in an unbalanced Mach-Zehnder Interferometer (MZI) as an optical spectral-shaper [90]. The major limitation of this approach is that the LCFBGs, once fabricated, are not reconfigurable and, as a result, the chirp rate is not tunable. Although it has been shown in the later and recent demonstrations [91-93] that the chirp rate can be tuned through the use of better optical spectral shapers, none of these demonstrations have been able to generate chirped microwave signals whose chirp rate can be tuned over a wide range as well as configured to yield positive, zero and negative values. Such reconfigurability should

be valuable in systems such as radar, surveillance and satellite communication. Therefore, it is important to develop a technique which will enable us to achieve this feature for the generation of chirped microwave signals.

1.5 Objectives

In this thesis, we have two main objectives.

Firstly, we aim to investigate and develop different techniques that can overcome some of the shortcomings found in the existing schemes for optically-sampled ADC as well as photonic TS-ADC systems. Specifically, we first investigate the performance of an optically-sampled ADC system which is based on complementary dual-output MZM and balanced detection scheme. Linearization in the form of arcsine operation is also described to illustrate the considerations of our proposed scheme. As for photonic TS-ADC, we investigate a technique which is based on polarization modulation and balanced photodetection so as to overcome the practical implementation issues pointed out in [54]. We also investigate a technique which can enable us to generate a TWIPT without the need to manually provide precise time-interleaving of multi-wavelength pulses together and without the usage of modulators and microwave sources. The technique is based on dispersion spreading and sectional compression and we also discuss the feasibility of using the generated TWIPT in a wavelength-interleaving optically-sampled ADC system.

Secondly, we aim to investigate and develop different techniques using mode-locked lasers which can enable us to generate frequency-tunable microwave signals, both CW and chirped. Specifically, we investigate two techniques that can enable us to generate CW microwave signals with large frequency-tunability. The first technique is based on using an array of uniformly-spaced optical combs, while the

second technique is based on using a temporally-stretched and chirped pulse-train. Lastly, we demonstrate the generation of microwave signals with large chirp tuning range by using a technique which involves the use of two temporally-stretched pulse-trains which are time-delayed with respect to each other and having different frequency chirp.

1.6 Major contributions of the thesis

Major contributions of the thesis are briefly described below.

A. Optically-sampled ADC based on complementary dual-output modulation and balanced photodetection:

We demonstrate for the first time a 2 GS/s optically sampled analog-to-digital converter based on a dual-output modulation and balanced photodetection scheme. For a single-tone signal at 200 MHz, we have obtained 5.52 SNR-bits, SFDR of 28 dB as well as ENOB of ~ 4.29 -bits based on our current system. Arcsine operation is digitally implemented on the output of the electronic ADC and is shown to enable significant odd-order harmonic suppression at high modulation depths approaching unity. More specifically, we have demonstrated a significant improvement in the SFDR which is now measured to be ~ 62 dB. Measuring across a range of analog input frequencies and operating the current system in undersampling mode, we measured and obtained an average ENOB of ~ 5.34 -bits after linearization.

B. Photonic TS-ADC based on polarization modulation and balanced photodetection:

We show analytically and experimentally that a PolM which supports transverse electric (TE) and transverse magnetic (TM) modes of opposite phase modulation indices can be utilized to reject dispersion-induced even-order distortions in a TS-ADC. For a single channel, using a 10 bit electronic ADC with a sampling rate of 2 GSs/s, we demonstrate digitization of RF signals up to 950 MHz and obtain ENOB ~ 3.56 bits with a single channel at $\sim 31.6\%$ of the electronic ADC's peak-to-peak full scale voltage.

C. Photonic generation of a TWIPT for optically-sampled ADC systems:

We show analytically as well as experimentally a technique to generate TWIPT with tunable pulsewidth. Based on our proposed scheme which involves dispersion and sectional compression, using a 2 GHz super-continuum (SC) source as an initial pulse-train, we demonstrate the generation of a 10×2 GHz TWIPT with pulsewidth tunability.

D. Photonic generation of frequency-tunable CW microwave signals using an array of uniformly-spaced optical combs:

We show analytically as well as experimentally a photonic technique for generation of frequency-tunable microwave signals. The technique involves propagating an array of uniformly-spaced optical combs with identical comb profile into a length of dispersive SMF prior to photodetection. Based on the proposed scheme, using a 2 GHz SC source as an initial optical comb, we demonstrate the generation of stable microwave signals tunable to 8, 10, and 12 GHz.

E. Photonic generation of frequency-tunable CW microwave signals using a temporally-stretched and chirped pulse-train:

We show analytically and experimentally that direct photodetection of a temporally-stretched and chirped optical pulse-train together with its time-delayed replica can be exploited to realize the generation of tunable CW microwave signals. Based on the proposed scheme, using an initial 2 GHz pulse-train, we demonstrate generation of tunable CW microwave signals from 4 GHz to 14 GHz.

F. Photonic generation of linearly-chirped microwave signals:

We show analytically as well as experimentally an approach to generate microwave waveforms with wide chirp tuning range. The approach is based on the interference of two temporally-stretched pulses which are time-delayed with respect to each other and have different frequency chirp. Based on this approach, we demonstrate the generation of microwave waveforms with different centre frequencies and with the chirp rates ranging from ~ -126.7 GHz/ns to $\sim +120.8$ GHz/ns, including the zero-chirp case.

1.7 Thesis organization

This thesis is organized as follows.

In chapter 1, we introduce the background and motivation for the existing photonic techniques for A/D conversion as well as the generation of microwave signals. The scope of this dissertation is clearly stated, and emphasis is placed on optically-sampled ADC, photonic TS-ADC as well as photonic generation of CW and chirped microwave signals. The main contributions are also presented and this chapter concludes with the organization of the thesis.

In chapter 2, a review of the optically-sampled ADC as well as the photonic TS-ADC is presented. Their working principles and the governing mathematical equations are also presented in this chapter for a thorough understanding.

In chapter 3, a review of the optical heterodyning, external modulation, dual-wavelength laser, which enable the generation of frequency-tunable CW microwave signals is presented. Following that, the working principle of the optical spectral shaping and wavelength-to-time mapping is described.

In chapter 4, we first describe an optically-sampled ADC based on complementary dual-output modulation and balanced detection scheme. Next, we describe a photonic TS-ADC which is based on polarization modulation and balanced detection scheme. Theoretical, simulation, and experimental results are presented.

In chapter 5, we discuss a technique which enables the generation of a TWIPT based on dispersion and sectional compression. This work is relevant to and supports optically-sampled wavelength-interleaved ADCs. Theoretical, simulation and experimental results are presented.

In chapter 6, we investigate a photonic technique which enables the generation of frequency-tunable CW microwave signals based on propagating an array of uniformly-spaced optical combs with identical comb profile into a length of dispersive SMF prior to photodetection. Theoretical, simulation and experimental results are presented.

In chapter 7, we investigate another photonic technique which involves using a temporally-stretched and chirped pulse-train to generate frequency-tunable CW microwave signals. Theoretical, simulation and experimental results are presented.

In chapter 8, we discuss a photonic technique which allows us to generate linearly-chirped microwave signals with wide chirp tuning range based on the principle of optical spectral shaping and wavelength-to-time mapping. Theoretical, simulation and experimental results are presented.

In chapter 9, we summarize our work as well as present the conclusion and suggestions for future work.

Chapter 2 REVIEW OF PHOTONIC ANALOG-TO-DIGITAL CONVERTERS

2.1 Introduction

This chapter describes the basic concepts and working principles of the optically-sampled ADC and the photonic TS-ADC. These two classes of photonic ADCs are emphasized in this thesis as their quantization process is realized by an electronic ADC. An electronic ADC with a sampling rate of 2 GS/s and a stated resolution of 10 bits will be used for the experiments. First, we present the principle of operation of a single-channel optically-sampled ADC based on a typical electro-optic modulation scheme. We then discuss the multi-channel interleaving schemes, namely time-interleaving and wavelength-interleaving, which can be utilized to increase the sampling rate of the electronic ADC. Secondly, we present the principle of operation of a single-shot photonic TS-ADC system which is also based on the electro-optic modulation scheme. Subsequently, we follow up with a discussion of a continuous-time TS-ADC system which is based on the virtual time-gating principle. Along our discussion, we also review some of the previous and recent developments in the aforementioned topics.

2.2 Optically-sampled ADC

As mentioned in Chapter 1, optical sampling using low timing jitter optical pulses from mode-locked lasers can help to overcome the jitter-limited ENOB issue that arises in the electronic ADCs. The schematic of a basic single-channel optically-sampled ADC is illustrated in Fig. 2.1. In general, an optically-sampled ADC system

Fig. 2.1 Schematic of a basic single-channel optically-sampled ADC. eADC: electronic analog-to-digital converter.

comprises an optical pulse source, an optical modulator, a high-speed PD as well as an electronic ADC. Suitable optical pulse sources include passive or active mode-locked lasers which are capable of generating ultrashort pulses with low timing jitter.

Lithium Niobate (LiNbO_3) MZM whose working principle is based on the linear electro-optic effect is typically chosen as the modulator. Optical sampling takes place when the input RF signal to be digitized intensity-modulates the optical pulse-train using the MZM. The pulse-train of samples is converted to the electrical domain by a photodetector and finally digitized by the electronic ADC. As the photodetector is bandwidth-limited, the electrical pulses will be broadened. Since the electronic ADC samples the relatively flat-top region of the broadened electrical pulses, the accuracy of the A/D conversion will not be affected significantly by the aperture jitter originating from the electronic ADC. Analogous to a typical analog photonic link, the performance of an optically-sampled ADC can be analyzed and characterized using standard metrics such as link SNR and SFDR.

2.2.1 Analysis of operation

The electric field of the input optical pulse can be expressed as

$$E_1(t) = E_o E(t) \quad (2.1)$$

where E_o and $E(t)$ represent the peak amplitude and shape of the optical pulse, respectively. This input optical pulse is modulated by the RF signal in the MZM. The electric field at the output of the MZM is given as

$$E_2(t) = E_1(t) \cos\left(\frac{\Delta\phi}{2}\right) \quad (2.2)$$

where $\Delta\phi$ is defined as

$$\Delta\phi = \frac{\pi}{2V_\pi} (V_{DC} + V_{RF}(t)) \quad (2.3)$$

Here, V_{DC} is the DC voltage used to bias the MZM, V_π is the half-wave voltage of the MZM, $V_{in}(t) = V_o \cos \omega_{RF} t$ is the applied input RF signal to the MZM, V_o and ω_{RF} are the amplitude and angular frequency of the RF signal, respectively. The output optical power of $E_2(t)$ can be expressed as

$$\begin{aligned} P_{out} &= |E_2(t)|^2 \\ &= E_o^2 E^2(t) \cos^2(\Delta\phi) \\ &= \frac{1}{2} P_o P(t) [1 + \cos(2\Delta\phi)] \\ &= \frac{1}{2} P_o P(t) \left[1 + \cos\left(\frac{\pi}{V_\pi} (V_{DC} + V_{in}(t))\right) \right] \\ &= \frac{1}{2} P_o P(t) [1 + \cos(\phi_{DC} + k \cos \omega_{RF} t)] \end{aligned} \quad (2.4)$$

where $P_o = E_o^2$, $P(t) = E^2(t)$, $\phi_{DC} = \pi V_{DC} / V_\pi$ and $k = \pi V_o / V_\pi$.

The transfer function of the MZM is known to have the best linearity when biased at the quadrature point. Depending on the polarity of the slope of the MZM's transfer function, its corresponding bias-phase ϕ_{DC} is equated to $\pm \pi / 2$. Biasing the MZM at the quadrature point and assuming the responsivity R of the PD to be flat over the relevant bandwidth, the photocurrent is given as

$$\begin{aligned}
 I(t) &= \frac{1}{2} R P_o P(t) [1 + \cos(\phi_{DC} + k \cos \omega_{RF} t)] \\
 &= \frac{1}{2} R P_o P(t) [1 + \cos(\pi / 2 + \theta_e + k \cos \omega_{RF} t)] \\
 &= \frac{1}{2} R P_o P(t) [1 - \sin(\theta_e + k \cos \omega_{RF} t)] \\
 &= \frac{1}{2} R P_o P(t) [1 - \sin \theta_e \cos(k \cos \omega_{RF} t) - \cos \theta_e \sin(k \cos \omega_{RF} t)]
 \end{aligned} \tag{2.5}$$

Here, we have taken into consideration the DC-bias drift in MZM and included a phase error θ_e which is associated with the deviation from the quadrature point.

Substituting the following Bessel function identities into (2.5),

$$\sin(k \cos \omega_{RF} t) = 2 \sum_{n=1}^{\infty} J_{2n-1}(k) (-1)^{n+1} \cos(2n-1)\omega_{RF} t \tag{2.6}$$

$$\cos(k \cos \omega_{RF} t) = J_0(k) + 2 \sum_{n=1}^{\infty} J_{2n}(k) (-1)^n \cos 2n\omega_{RF} t \tag{2.7}$$

the expression for the photocurrent can be further expanded in a Bessel series

$$I(t) = \frac{1}{2} R P_o P(t) \begin{bmatrix} 1 - J_o(k) \sin \theta_e \\ -2 \cos \theta_e J_1(k) \cos \omega_{RF} t \\ +2 \sin \theta_e J_2(k) \cos 2\omega_{RF} t \\ +2 \cos \theta_e J_3(k) \cos 3\omega_{RF} t \\ -\dots \end{bmatrix} \quad (2.8)$$

where $J_n(k)$ is the n^{th} order Bessel function of the first kind.

In (2.8), the first two terms within the parenthesis are the DC terms and the third term corresponds to the modulating RF signal. As for the remaining higher-order harmonic terms, the odd-order terms are attributed to the nonlinear characteristic of the MZM's transfer function while the even-order terms are caused by θ_e . These even-order harmonics can be eliminated if the MZM is biased exactly at the quadrature point. Taking the Fourier-transform of (2.8) give us its corresponding RF spectrum which is made up of a series of sidebands whose amplitudes are determined by the coefficients of the Bessel functions. The analytical expressions for the 2nd-order distortion and 3rd-order distortion power ratios can then be expressed as:

$$\frac{P_{rf}(2\omega_{RF})}{P_{rf}(\omega_{RF})} = \frac{\sin^2 \theta_e [J_2(k)]^2}{\cos^2 \theta_e [J_1(k)]^2} \quad (2.9)$$

$$\frac{P_{rf}(3\omega_{RF})}{P_{rf}(\omega_{RF})} = \frac{[J_3(k)]^2}{[J_1(k)]^2} \quad (2.10)$$

where, for calculations, we can invoke the property of the Bessel function:

$$J_n(k) \approx k^n / 2^n n!.$$

Figs. 2.2 (a) and (b) illustrate the normalized power of the 2nd and 3rd-order distortions for modulation depth $m_{dep} = 2V_o / V_\pi$ of 0.4 and 0.8, with quadrature phase

Fig. 2.2 Theoretical 2nd-order normalized distortion power (solid lines) and 3rd-order normalized distortion power (dashed lines) for modulation depths of (a) 0.4 and (b) 0.8.

error up to 0.1 radians, respectively. As evident, a larger modulation depth leads to an increase in the power of the 3rd-order distortion, consequently degrading the SFDR. Furthermore, the performance of the photonic ADC system can also be affected and limited by the 2nd-order distortion when θ_e increases beyond 0.01 radians.

Considering only the nonlinearity due to the sinusoidal transfer function of the MZM, the SFDR of an optically-sampled ADC is generally limited by the 3rd-order distortion. In order to obtain a large SNR for the optically-sampled ADC, we can either increase the optical power of the input pulse-train or operate at a larger modulation depth.

For the first approach, however, additional nonlinear distortion would arise whenever the input optical power of the pulse-train exceeds the linear range of the PD's response, thus further degrading the SINAD. Although the SFDR is limited by the 3rd-order distortion when operating the optically-sampled ADC at large

modulation depths, a digital linearization scheme in the form of arcsine operation [16] can suppress the odd-order distortions which are caused by the MZM's nonlinear transfer function. Thus, the second approach proves to be a more attractive choice to improve the SNR.

Recently, phase-encoded optical sampling has also been demonstrated in [22-24], wherein a complementary dual-output MZM is utilized in place of a single-output MZM. Besides suppressing the odd-order distortions due to the dual-output MZM's nonlinear transfer functions, the arcsine operation in this variant of optically-sampled ADC also suppresses the amplitude noise from the pulse source, increases the SNR by 6 dB, as well as removes the DC terms. The only drawback with using the phase-encoded optical sampling scheme is that it requires twice the number of electronic ADCs as the typical single-output intensity modulation scheme. To overcome this drawback, rather than the usual photodetection using separate PDs, balanced detection at the output of the dual-output MZM can be utilized. This is discussed in greater detail in Chapter 4.

2.2.2 Multi-channel interleaving schemes

Till now, we have only discussed the single-channel optically-sampled ADCs in which optical sampling is exploited to improve the jitter-limited ENOB issue associated with the electronic ADCs. However, the sampling rate of these single-channel optically-sampled ADCs is limited by the sampling rate of the back-end electronic ADC. To deal with this particular limitation, multi-channel interleaving schemes have to be implemented. Specifically, there are two possible schemes, namely, the time-interleaving and wavelength-interleaving schemes. The schematics of these schemes are illustrated in Figs. 2.3 (a) and (b), respectively.

Fig. 2.3 Schematic of the multi-channel interleaving schemes. (a) Time-interleaving scheme based on optical time-division-demultiplexing, and (b) wavelength-interleaving scheme based on wavelength-division-demultiplexing. MZM: Mach-Zehnder modulator; SMF: single-mode-fiber.

For the time-interleaving scheme, the sampling is realized by an optical pulse-train whose centre wavelength is fixed and has a repetition rate f_{rep} which is too fast to be captured by a single ADC. The sampled pulse-train can be divided into N parallel channels of pulse-trains with a slower repetition rate of f_{rep}/N through OTDD. After photodetection, these slower electrical pulse-trains are individually

digitized by an electronic ADC. Finally, the data collected from these channels are interleaved together in the digital domain and as a result, the sampling rate of the optically-sampled ADC system increases by N -fold.

In contrast to the time-interleaving scheme, a TWIPT which comprises a repeated sequence of N uniformly-spaced pulses with N different centre wavelengths is used for the optical sampling in the wavelength-interleaving scheme shown in Fig. 2.3 (b). Through the use of WDD, the sampled pulse-train in this scheme is divided into N wavelength channels and slowed down to f_{rep} / N . Similar to the time-interleaving scheme, the slowed down pulse-trains are individually photodetected and digitized by electronic ADCs. Digitally interleaving the data from these channels ensures that this scheme also increases the sampling rate of the optically-sampled ADC system by N -fold.

The time-interleaving approach was first investigated by Bell *et al.* [17, 18] and they demonstrated a 2 Gs/s optically-sampled time-interleaved ADC having 2.8 ENOB. Based on phase-encoded sampling [16, 22], Twichell *et al.* later demonstrated a 208 MS/s optically-sampled time-interleaved ADC using 4 parallel channels of 12-bit 52 MS/s ADCs. Extending this work further, Juodawlkis *et al.* demonstrated a 505 MS/s 8 channel system which has a quantizer-noise limited SNR of 51 dB or 8.2 SNR_{bits} [23]. The analysis and calibration techniques developed in [16, 22, 23] later on provided the foundation for Williamson *et al.* to demonstrate a 505 MS/s 8 channel system which is evaluated to have 9.8 ENOB when sampling a 733 MHz input signal [26].

On the other hand, the wavelength-interleaving approach was first proposed by Frankel *et al.* [21]. Kang *et al.* demonstrated a TWIPT which can enable continuous-

sampling at 160 MS/s when used with a 1-to-4 passive wavelength-division-demultiplexer. Utilizing a generated TWIPT with a 1-to-3 demultiplexer, Coppinger *et al.* later demonstrated a 12 GS/s optically-sampled wavelength-interleaved ADC which has a resolution of 2 ENOB [99]. Based on an 8 wavelength channel TWIPT which can support optical-sampling at 10 GS/s, Clark *et al.* demonstrated non-continuous sampling from a single 1.25 GS/s channel and estimated a resolution of 7 ENOB for the optically-sampled wavelength-interleaved ADC [29]. Recently, Khilo *et al.* demonstrated a 2 wavelength channel TWIPT which has an aggregate sampling rate of 2.1 GS/s. Performing undersampling on a 41 GHz signal, they obtained a resolution of 7 ENOB. Additionally, an integrated silicon chip comprising a modulator, filters and photodetectors was created and incorporated with the generated TWIPT. A 10 GHz signal was sampled and a resolution of 3.5 ENOB was obtained [27].

Although both the time-interleaving and wavelength-interleaving schemes have shown promising results over the years, the need to use a fast active switch for OTDD has deterred many from using the time-interleaving scheme. The reason is that the sampling rate of the optically-sampled ADC system is often found limited by the response speed of the fast active switch which is controlled by complex timing signals to demultiplex a high-speed optical pulse-train into parallel channels of lower speeds. On the other hand, the wavelength-interleaving scheme requires only passive components such as a wavelength-division-demultiplexer to divide the sampled TWIPT into different wavelength channels, thus simplifying its practical implementation. The sampling rate of an optically-sampled ADC utilizing the wavelength-interleaving scheme is instead limited by the number of wavelengths which the TWIPT can support. As a result, much effort has been devoted to

investigate techniques which have the potential to generate a TWIPT suitable for the wavelength-interleaving scheme.

2.2.3 Time- and wavelength interleaved pulse-train (TWIPT)

One of the widely-used approaches to generate a TWIPT is based on providing a uniform time-delay to consecutive pulses of different wavelengths [29, 30]. The schematic of this approach is shown in Fig. 2.4. First, a broadband optical pulse-train of repetition period T is spectrally divided into N wavelength channels using a 1-to- N wavelength-division-demultiplexer. Each of these channels consists of an optical pulse-train which is centered at a different wavelength. Next, these pulse-trains are time-delayed with respect to the adjacent channel by a uniform amount and subsequently coupled back to a single channel using a N -to-1 wavelength-division-multiplexer. As a result, a TWIPT which comprises a repeated sequence of N uniformly-spaced pulses with N different centre wavelengths is generated.

The repetition period of this pulse-train decreases by N times as well. Although conceptually simple, this approach requires precise uniform time-delay between adjacent wavelength channels as well as equal amplitude in all channels. These requirements are realized by optical-tunable-delay-lines (OTDL) and variable optical attenuators (VOA), respectively. However, for a TWIPT which requires many wavelength channels, practical implementation is difficult because of the large number of components that need to be precisely tuned. Over the years, many other approaches have been developed to tackle the abovementioned issue [31-36]. One effective solution is based on the time-lens compression technique [31, 37] which is depicted in Fig. 2.5.

Fig. 2.4 Schematic for the generation of a TWIPT based on uniform time-delay of consecutive pulses with different wavelengths.

Fig. 2.5 Schematic for the generation of a TWIPT based time-lens compression technique. MZM: Mach-Zehnder modulator; PM: Phase modulator; SMF: single-mode-fiber.

In the time-lens compression technique, outputs of N number of CW lasers with uniform wavelength spacing are first multiplexed together and modulated at a frequency $f_{RF} = 1/T$ by a MZM for pulse carving. The N temporally overlaid pulse-trains of different wavelengths generated at the output of the MZM are then fed to the time-lens system, which consists of a phase modulator and a length of SMF. The SMF introduces a quadratic phase shift which compensates the existing phase profile of the temporally-overlaid pulse-trains after phase modulation and as a result, the pulse-trains are temporally compressed and displaced with a uniform time interval. By choosing appropriate length for the SMF in relation to the chosen values of N and

wavelength spacing between the CW lasers, one is able to generate a TWIPT with a uniform inter-pulse spacing of T/N . Recently, Qingwei *et al.* have demonstrated a 40 GS/s optically-sampled ADC system using a TWIPT generated based on the time-lens compression technique and obtained an ENOB of 3.45 bits [37]. So far, this demonstration has one of the higher performances among the many demonstrated optically-sampled ADC system using TWIPT.

Although the time-lens compression technique offers a simpler practical implementation, an array of CW lasers as well as a modulator and a microwave source are required, rendering this technique bulky and expensive. Furthermore, the stability of the TWIPT is largely dependent on the stability of the microwave source

Fig. 2.6 “Walden chart” illustrating ENOB of existing ADCs versus analog input frequency.

Light and dark blue circles correspond to the electronic ADCs surveyed by [5] and [27], respectively. Orange stars are some of the high-performance wideband optically-sampled ADCs demonstrated so far, with the bigger star representing the current state-of-the-art performance [27]. All optically-sampled ADCs in this figure operate in undersampling mode. This figure is reproduced from [27].

used. To overcome these shortcomings as well as the practical issue associated with the previous technique shown in Fig. 2.4, a scheme which is based on dispersion and sectional compression can be implemented to generate a TWIPT. The key to this scheme is a multi-channel dispersion module (MCDM). Further details of this scheme are discussed in Chapter 5.

Fig. 2.6 illustrates the "Walden Chart" as surveyed by [5] and [27]. Several high-performance wideband optically-sampled ADCs (orange stars) demonstrated in undersampling mode recently are shown to have performance which are comparative or better than their electronic counterparts (blue circle). In undersampling, the analog input frequencies are sampled at a rate below the Nyquist rates of the ADC. The current state-of-the-art is a 2.1 GS/s optically-sampled ADC which has an ENOB of 7-bits when undersampling a 41 GHz analog input frequency [27] (depicted by the bigger orange star). Specifically, this demonstration is based on the wavelength-interleaving scheme shown earlier in Fig. 2.3(b) whereby the TWIPT pulse source used is generated by time-interleaving two individual 1.05 GHz mode-locked laser with different wavelengths using the scheme in Fig. 2.4.

2.3 Photonic time-stretched ADC

Unlike optically-sampled ADC which utilizes the optical sampling to circumvent the jitter-limited ENOB issues associated with electronic ADCs, the photonic TS-ADC was proposed and developed by Jalali *et al.* [39-45] with the motivation to enhance the effective input bandwidth and overall sampling rate of the available electronic ADCs. Fig. 2.7 illustrates the schematic of a photonic TS-ADC system which is capable of capturing single-shot input RF signal. An ultrashort broadband optical pulse, typically generated from a supercontinuum (SC) source, is first

Fig. 2.7 Schematic of a single-shot photonic TS-ADC system. eADC: electronic analog-to-digital converter.

propagated through a length of dispersive fiber such that the pulse becomes temporally broadened and chirped.

Using an intensity modulator, the chirped pulse is modulated by an input RF signal and the information carried by this time-domain signal is mapped onto the wavelength domain due to the principle of time-to-wavelength mapping. Next, the modulated chirped pulse propagates through a second length of dispersive fiber. Wavelength-to-time mapping is performed at this fiber, resulting in the signal to be stretched in time as the pulse is further chirped. The initial input RF signal is time-stretched and eventually slowed down, thus relaxing the requirements on the sampling rate of the back-end electronic ADC. Moreover, the effective input bandwidth of the electronic ADC increases as well. As shown in [45], although optical fibers possess nonlinear group-velocity-dispersion (GVD), the nonlinear dispersion in the two fibers of a photonic TS-ADC counteract each other, and as a result, has no effect on the stretch factor. Thus, the stretch factor of a photonic TS-ADC system only relies on the linear GVD.

2.3.1 Analysis of operation

Following the mathematical framework in [45], we can analyze the single-shot photonic TS-ADC system. Assuming that the ultrashort broadband optical pulse is transform-limited and has a Gaussian-shaped profile, its electric field can be expressed in the time-domain as

$$E_1(t) = E_o \exp\left(\frac{-t^2}{2\tau_o^2}\right) \quad (2.11)$$

where E_o is the pulse amplitude and τ_o is the pulse half-width which is related to the full-width-half-maximum (FWHM) pulsewidth $\tau_{FWHM} = 2(\ln 2)^{0.5} \tau_o$. Propagating through the first dispersion module comprising a length L_1 of fiber, the electric field can be expressed in the frequency domain as

$$E_2(\omega) = E_1(\omega) \exp\left(\frac{j\beta_2\omega^2 L_1}{2} + \frac{j\beta_3\omega^3 L_1}{6}\right) \quad (2.12)$$

where $E_1(\omega)$ is the FFT of $E_1(t)$, while β_2 and β_3 are the 2nd-order and 3rd-order GVD of the fiber, respectively.

The dispersed pulse is next modulated with an angular frequency ω_{RF} by a push-pull MZM biased at the quadrature point. The electric field is then given in the time domain as

$$E_3(t) = E_2(t) \cos\left(\frac{\pi}{4} + \frac{k}{2} \cos \omega_{RF} t\right) \quad (2.13)$$

where V_π is the half-wave voltage of the push-pull MZM, $k = \pi V_o / V_\pi$, and V_o is the amplitude of the RF signal.

Propagating the modulated chirped pulse through the second dispersion module comprising a length L_2 of fiber, the electric field can be expressed in the frequency domain as

$$E_4(\omega) = E_3(\omega) \exp\left(\frac{j\beta_2\omega^2 L_2}{2} + \frac{j\beta_3\omega^3 L_2}{6}\right) \quad (2.14)$$

The photocurrent at the output of the PD is then given as

$$I(t) = RE_4(t) \cdot E_4^*(t) \quad (2.15)$$

where R is the responsivity of the PD. Considering only linear GVD, i.e., neglecting the effects of β_3 , as well as using the Bessel function identities as described in sections (2.5) and (2.6), we can expand (2.13) as follows:

$$E_3(t) = \frac{\sqrt{2}}{2} E_2(t) \left[J_0\left(\frac{k}{2}\right) + \sum_{n=1}^{\infty} 2c(n) J_n\left(\frac{k}{2}\right) \cos(n\omega_{RF}t) \right] \quad (2.16)$$

$$\text{where } c(n) = \begin{cases} (-1)^{n/2}, & n = \text{even} \\ (-1)^{(n+1)/2}, & n = \text{odd} \end{cases}.$$

Substituting the FFT of (2.16) into (2.14), the electric field at the input of the PD can be written as

$$E_4(\omega) \approx J_0\left(\frac{k}{2}\right) E_{env}(\omega) + \sum_{n=1}^{\infty} c(n) J_n\left(\frac{k}{2}\right) \exp(jn^2\phi_{DIP}) \left[E_{env}\left(\omega - \frac{n\omega_{RF}}{M}\right) + E_{env}\left(\omega + \frac{n\omega_{RF}}{M}\right) \right] \quad (2.17)$$

$$\text{where } E_{env}(\omega) = \frac{\sqrt{2}}{2} E_o (2\pi\tau_o^2)^{0.5} \exp\left[-\frac{\omega^2\tau_o^2}{2} + \frac{j\beta_2\omega^2(L_1 + L_2)}{2}\right] \text{ and } M = 1 + L_2 / L_1$$

is the stretch factor. The envelope function $E_{env}(\omega)$ corresponds to the electric field

when $k = 0$, while ϕ_{DIP} is the dispersion-induced phase. The details to obtain (2.17) and (2.18) can be found in [45].

Invoking the following properties of Bessel functions, $J_n(k) \approx (1/2^n n!) k^n$ and $J_p(k) J_q(k) \approx (1/2^{p+q} p! q!) k^{p+q}$, we substitute (2.17) into (2.15). Keeping only terms with the lowest $(p+q)$, the photocurrent can be expressed as

$$I(t) = I_{env} \times \left[\begin{aligned} & J_o^2\left(\frac{k}{2}\right) \\ & - 4J_o\left(\frac{k}{2}\right) J_1\left(\frac{k}{2}\right) \cos \phi_{DIP} \cos \frac{\omega_{RF} t}{M} \\ & + \left\{ 2J_1^2\left(\frac{k}{2}\right) - 4J_o\left(\frac{k}{2}\right) J_2\left(\frac{k}{2}\right) \cos 4\phi_{DIP} \right\} \cos \frac{2\omega_{RF} t}{M} \\ & + \left\{ 4J_o\left(\frac{k}{2}\right) J_3\left(\frac{k}{2}\right) \cos 9\phi_{DIP} + 4J_1\left(\frac{k}{2}\right) J_2\left(\frac{k}{2}\right) \cos 3\phi_{DIP} \right\} \cos \frac{3\omega_{RF} t}{M} \\ & + \dots \end{aligned} \right] \quad (2.18)$$

$$\text{where } I_{env} = \frac{1}{2} R E_o^2 \frac{1}{\left[\left(1 + (L_1 + L_2)^2 \right) / L_D^2 \right]^{0.5}} \exp \left(\frac{-t^2}{\tau_o^2 \left[1 + (L_1 + L_2)^2 / L_D^2 \right]} \right) \text{ and}$$

$L_D = \tau_o^2 / |\beta_2|$ is the dispersion length.

Considering the case where k is small, the photocurrent can be simplified as

$$I(t) = I_{env} \times \left[\begin{aligned} & 1 - k \cos \phi_{DIP} \cos \frac{\omega_{RF} t}{M} \\ & + \frac{k^2}{8} (1 - \cos 4\phi_{DIP}) \cos \frac{2\omega_{RF} t}{M} \\ & + \frac{m^3}{96} (\cos 9\phi_{DIP} + 3 \cos 3\phi_{DIP}) \cos \frac{3\omega_{RF} t}{M} + \dots \end{aligned} \right] \quad (2.19)$$

From (2.19), we can obtain a deeper insight into the effects of dispersion penalty and harmonic distortion on the performance of a photonic TS-ADC system. Within

the parenthesis, the first term is the DC term, the second term is the desired time-stretched signal, while the remaining terms are the undesired harmonics. Clearly, the original input RF signal has been slowed down by a factor of M and as a result, this increases the effective sampling rate and input bandwidth of the electronic ADC by a factor of M as well. The stretch factor M in a single-shot TS-ADC system is only limited by the repetition period of the pulse source used. Besides these benefits, however, time-stretching also causes the time-stretched signal to experience frequency-dependant attenuation. This effect is attributed to the beating of carrier with the lower and upper sidebands [41] and is analogous to the undesirable power penalty or RF fading in analog photonic links. However due to time-stretching, the effect is reduced M -fold in a photonic TS-ADC as compared to analog photonic links.

Although the biasing of the MZM at the quadrature point suppresses the 2nd-order harmonic due to the nonlinear transfer function of the MZM, the dispersive fibers which are required for the time-stretching process inevitably introduce the 2nd-order harmonic distortion as indicated by the third term in (2.18). As for the 3rd-order harmonic distortion indicated by the fourth term, it is induced by the combination of the sinusoidal nonlinear transfer function of the MZM as well as the dispersion introduced by the dispersive fibers.

The influence of nonlinear GVD, characterized by β_3 is also investigated in [45]. Analytically and numerically, it has been shown that the wavelength dependence of β_3 introduces a wavelength dependent dispersion penalty on the desired time-stretched signal. Via wavelength-to-time mapping, a time-dependent attenuation is also introduced. In practice, however, the time-dependent attenuation is relatively weak as β_3 is typically a few orders of magnitude lower than β_2 . Thus, this unwanted effect is generally masked by other sources of noise which include amplified

spontaneous emission, pulse-to-pulse variations of the SC source and quantization noise from the electronic ADC.

Early demonstrations of single-shot photonic TS-ADC mainly involved using long lengths of SMF [44, 46] as the dispersive medium to obtain large values of the stretch factor and in turn, a higher sampling rate. To avoid the large loss inherent in long lengths of SMF, DCF which can provide a higher dispersion-to-loss ratio than SMF is utilized to reduce the loss in the photonic TS-ADC system [47-50]. To compensate for the loss introduced either by SMF or DCF, most early demonstrations utilized discrete optical amplifiers such as erbium doped fiber amplifiers (EDFA). In contrast to the discrete amplifiers, the recent demonstrations on distributed Raman amplification (DRA) within DCF has shown the ability to maximize the SNR of the system as it is capable of maintaining a constant signal power throughout the dispersive fiber, thus keeping the signal away from low (lossy) and high (nonlinear) power regimes [47, 50].

With regard to the frequency-dependent attenuation suffered by the desired time-stretched signal, the results from [52] have shown that SSB modulation with zero chirp using a push-pull MZM can mitigate this undesirable effect. Implementing SSB with chirp modulation based on a single-arm MZM biased at quadrature point, the 3-dB bandwidth of the photonic TS-ADC system can be improved by 41.4% as compared to the zero chirp case [53]. As for the 2nd-order harmonic distortion, Gupta *et al.* have shown that by operating the photonic TS-ADC system in a differential manner in conjunction with the digital linearization technique in the form of arcsine operation [54], these dispersion-induced distortions can be cancelled out. Moreover, the 3rd-order harmonic distortion can also be suppressed to a certain extent depending on the degree of amplitude and phase mismatches between the differential channels.

The distortions due to the non-uniform spectral density of the SC and its pulse-to-pulse temporal variations can also be removed through this method.

To realize the differential operation, a complementary dual-output MZM is used and the complementary pulses from the MZM outputs are deliberately time delayed by half a repetition period, to ensure that no temporal overlapping of the pulses occurs before combining and time-stretching by a single dispersive channel. The time-stretched pulses are then required to align in time and are subsequently subtracted in the digital domain using the arcsine operation to obtain the desired differential mode. Practically, the ideal value of time-delay between the complementary pulses is difficult to achieve, and as a result amplitude and phase imbalances between the pulses will arise, affecting the accuracy of the differential operation.

Performance-wise, Bhushan *et al.* demonstrated a 130 GS/s single-shot photonic TS-ADC based on the SMF as the dispersive medium. A stretch factor of 16.2 was obtained on a 11.1 GHz tone which was subsequently digitized by a 8 GS/s single channel real-time oscilloscope, where they obtained approximately 7.16 ENOB [46]. More recently, a 10 Tera samples per second (TS/s) single shot photonic TS-ADC was demonstrated by Chou *et al.* where an unprecedented stretch factor of 250 was attained through the use of distributed Raman optical amplification inside the DCF. An ENOB of 4.5 bits was realized in this scheme [47].

2.3.2 Continuous-time photonic TS-ADC system

The photonic TS-ADC is also implementable in continuous-time when the virtual time-gating principle is implemented. A proof-of-concept of this principle was first demonstrated by Han and Jalali in a 2 channel continuous-time photonic TS-ADC system [51]. Chou *et al.* later demonstrated a 4 channel continuous-time photonic TS-

ADC with a stretch factor of 3. Mating the optical front end with a 4 channel oscilloscope with a sampling rate of 50 GS/s, an effective sampling rate of 150 GS/s was obtained. Digitizing up to 45 GHz input RF signal, they obtained a resolution of 2.5 ENOB [49]. Based on an algorithm which can correct for distortions caused by time-warps and wavelength-dependent MZM bias offsets in photonic TS-ADC system, Gupta *et al.* demonstrated a 2 channel continuous-time ADC which has the highest ENOB at 7 bits for a 10 GHz bandwidth [100].

The concept of a continuous-time photonic TS-ADC system based on virtual time-gating principle can be explained with the help of Fig. 2.8 and Fig. 2.9. An ultrashort broadband pulse-train with a repetition period T [Fig.2.9(a)] is first generated from a SC source prior to the time-stretching process. The pulses are next propagated into the dispersion module 1 (DM1) resulting in the pulses being linearly-chirped and temporally-broadened to the interpulse spacing [Fig. 2.9(b)]. This is an essential criterion towards continuous-time operation. An input RF signal [Fig. 2.9(c)] is next imposed onto the linearly-chirped pulses using an intensity modulator. The modulated chirped pulses [Fig. 2.9(d)] are then temporally-stretched and linearly-chirped for the second time after propagating through the dispersion module 2 (DM2).

As explained earlier for the single-shot photonic TS-ADC system, the stretch factor M is related to the lengths of DM1 and DM2. It is observed at this point that there is temporal overlapping between adjacent pulses [Fig. 2.9(e)]. Although indistinguishable in time, the overlapping sections are actually distinguishable in wavelength. Hence, without the need to use any fast active switches, the overlapping segments can be separated into individual wavelength channels based on wavelength-division-demultiplexing. As depicted in Fig. 2.8, we have used 4 wavelength channels for this illustration.

Fig. 2.8 Functional block diagram for continuous-time photonic TS-ADC system based on virtual time-gating principle. eADC: electronic analog-to-digital converter.

Fig. 2.9 Continuous-time photonic TS-ADC based on virtual time gating principle: (a) Broadband ultrashort pulse-train, (b) pulses stretched after DM1, (c) input RF signal, (d) intensity modulated optical pulses, (e) pulses stretched after DM2, (f1-f4) intensity modulated pulses for channel 1-4, (g) reconstructed signal from digitized data taken from the 4 channels.

Unlike single-shot TS-ADC, the stretch factor M in continuous-time TS-ADC is limited by the number of wavelength channels used. Stretching beyond the number of wavelength channels used will result in crosstalk between similar wavelengths that carried different information in the same time slot before wavelength-demultiplexing. According to linear time-to-wavelength mapping, each individual wavelength channel corresponds to different distinct time segments [Fig.2.9(f1)-(f4)]. A four-channel continuous-time TS-ADC is realized when the signals from all four channels are photodetected and digitized at 'X' GS/s concurrently [[Fig.2.9(f1)-(f4)]. The digitized samples taken from each channel can be used to reconstruct the RF signal, which would increase the effective sampling rate to '4X' GS/s [Fig. 2.9(g)].

Although somewhat similar to the previously discussed multi-channel interleaving optically-sampled ADC system, the continuous-time TS-ADC system slows down the input RF signal through the time-stretching process and consequently the electronic ADCs see a slowed down signal and sample it above the Nyquist rate. This is in contrast to the multi-channel interleaving optically-sampled ADC system where its electronic ADCs see the full bandwidth of the input RF signal and sample at the sub-Nyquist rate. Exploiting this fundamental advantage, it has been shown that the interchannel mismatch errors in the photonic TS-ADC system can be estimated and corrected using the information available in the time-stretched signal [101].

As explained earlier, although the use of a complementary dual-output MZM together with digital linearization can suppress the 2nd-order harmonic distortion in a photonic TS-ADC system, it is however difficult to implement in practice. To avoid the practical implementation issues in [54], a PolM which supports the complementary pulses on the same transmission fiber can be utilized in a photonic TS-ADC system. The use of a PolM can eliminate the need to provide deliberate

time-delay to the complementary pulses as well as avoid the bias-drifting issues which are usually associated with a typical MZM. The details for this alternative approach will be discussed in chapter 4.

Table 2-1 briefly summarizes the advantages and disadvantages of the optically-sampled ADC, photonic TS-ADC and their various variants which have been discussed in this chapter.

Table 2-1 Summary of the optically-sampled ADC and photonic TS-ADC systems.

Photonic ADC	Advantages	Disadvantages
Optically-sampled ADC	• Simple to implement. • Able to perform undersampling.	• Sampling rate limited by backend electronic ADC
Optically-sampled ADC, multi-channel time-interleaving	• Does not require the use of any specially-tailored pulse source. • Overall sampling rate is N times of the individual backend electronic ADC used. N times being the number of channels. • Able to perform undersampling.	• Requires the use of a fast active switch for OTDD. • Overall sampling rate is limited by response speed of switch.
Optically-sampled ADC, multi-channel wavelength-interleaving	• Requires the use of only passive components for WDD. • Overall sampling rate is N times of the individual backend electronic ADC used. N times being the number of channels. • Able to perform undersampling.	• Requires the use of a specially-tailored TWIPT source. • Overall sampling rate is limited by number of wavelengths TWIPT source can support.
Single-shot photonic TS-ADC	• Simple to implement. • Enhance the effective input bandwidth and overall sampling rate of backend electronic factor by M-times. • M-times determined by the dispersion induced in the system and is limited only by the repetition period of pulse source. • Overall sampling rate can reach > 10 TS/s.	• Can only sample waveforms with limited time aperture and not continuous waveform. • Requires use of long length of dispersive fiber and EDFAs for loss compensation, making system bulky.
Continuous-time photonic TS-ADC	• Can sample continuous waveforms. • Enhance the effective input bandwidth and overall sampling rate of backend electronic factor by M-times. • Overall sampling rate can reach > 100GS/s.	• M-times determined by the dispersion induced in the system and is limited by the number of wavelength channels. • Requires use of long length of dispersive fiber and EDFAs for loss compensation, making system bulky. • Requires digital processing in the form of gain and time offset mismatches compensation of the wavelength channels.

2.4 Summary

In this chapter, we have described the operation principle and the method of analysis for a basic single-channel optically-sampled ADC system. In addition, we have also introduced two multi-channel interleaving schemes which can increase the sampling rate of the electronic ADC. The two schemes are the time-interleaving and wavelength-interleaving approaches. We have also discussed the various techniques that enable generation of a TWIPT which is suitable for the wavelength-interleaving approach.

Following the review of the optically-sampled ADCs, we have also described the operation principle and the method of analysis for a basic single-shot photonic TS-ADC system. Realization of a continuous-time photonic TS-ADC system based on the virtual time-gating principle is also presented. Along with this description, we have also reviewed a number of previous and recent developments for both types of ADCs. Lastly, we have also briefly indicated the alternative techniques which can overcome the shortcomings in some of the recently demonstrated techniques for optically-sampled ADC system, generation of TWIPT as well as photonic TS-ADC system. These alternative techniques are described in detail in Chapters 4 and 5.

Chapter 3 REVIEW OF PHOTONIC GENERATION OF MICROWAVE SIGNALS

3.1 Introduction

In this chapter, various techniques developed over the last decade for the photonic generation of microwave signals are reviewed with an emphasis on their principle of operation. Discussed first are a number of widely-used techniques which enable the generation of CW microwave signals. These techniques include optical heterodyning, external modulation, and dual-wavelength laser. As a less complex alternative to these conventional approaches, the technique based on the direct photodetection of a multi-longitudinal mode laser is also described. Lastly, the optical spectral-shaping and wavelength-to-time mapping technique which enables the generation of chirped microwave signals is also discussed.

3.2 Photonic generation of CW microwave signals

Microwave signals are used in diverse applications which include wireless communication, satellite communication, remote sensing, navigation and radar. In most applications, microwave signals are typically generated by electronic sources. Owing to the performance bottlenecks suffered by the electronic circuits, electronic microwave sources have limitations in generating high frequency signals. Additionally, the high frequency electronic microwave sources tend to be expensive and complex in design.

Fig. 3.1 Operation principle for optical heterodyning.

Besides the conventional electronic approach, the photonic generation of microwave signals has also received extensive attention in the recent years and has shown promise in generating microwave signals at very high frequency with high long-term stability and reliability. Furthermore, thanks to the low loss nature of optical fibers, distribution of microwave signals to a remote site over optical fibers can also avoid the large transmission loss which is inherent in coaxial cables.

3.2.1 Optical heterodyning

One simple and effective photonic approach to generate CW microwave signals is through optical heterodyning or simply, the beating of optical signals from two independent CW lasers of different wavelengths at a PD, as illustrated in Fig. 3.1. Assuming that the two optical waves are centered at λ_1 and λ_2 , their electric fields are given as

$$E_1(t) = E_{o,1} \cos(\omega_1 t + \phi_1) \quad (3.1)$$

$$E_2(t) = E_{o,2} \cos(\omega_2 t + \phi_2) \quad (3.2)$$

where $E_{o,1}$ and $E_{o,2}$ represent the amplitude of the optical signals, ω_1 and ω_2 are the corresponding angular frequencies, while ϕ_1 and ϕ_2 are the phase terms of the two optical signals.

Assuming that the two optical signals which are fed to a bandwidth-limited PD have the same polarization, the photocurrent can be expressed as

$$\begin{aligned} I(t) &= R [E_1(t) + E_2(t)] \cdot [E_1(t) + E_2(t)]^* \\ &= R [E_{o,1}^2 + E_{o,2}^2 + 2E_{o,1}E_{o,2} \cos[(\omega_1 - \omega_2)t + (\phi_1 - \phi_2)]] \\ &= R [P_{o,1} + P_{o,2} + 2\sqrt{P_{o,1}P_{o,2}} \cos[2\pi f_{RF}t + (\phi_1 - \phi_2)]] \end{aligned} \quad (3.3)$$

where R is the responsivity of the PD, $P_{o,1}$ and $P_{o,2}$ are the optical powers of the optical signals, and f_{RF} is the frequency of the generated CW microwave signal. As can be seen from (3.3), the frequency of the generated CW microwave signal is equal to the frequency difference of the two optical signals and is related to the wavelength spacing $\Delta\lambda_{sp}$ between them. By using tunable CW lasers, this technique can generate frequencies which can be tuned from microwave to mm-wave signals. However, the generated frequencies are often limited by the bandwidth of the PD.

As the phases of the optical signals are usually not correlated due to the free-running nature of the CW lasers, in this technique the generated microwave or mm-wave signal typically has a high phase noise. To achieve low-phase noise, it is necessary for the two optical signals to have phase coherence between them. This can be realized through either optical injection locking [59, 60] or optical phase-locked loop technique [61, 62], which are variants of the optical heterodyning technique.

Fig. 3.2 Schematic of an optical injection locking system consisting of a master laser and two slave lasers.

Fig. 3.3 Schematic of an optical phase-locked loop system.

Fig. 3.2 shows the schematic of an optical injection locking system which comprises a master laser and two slave lasers. A series of optical sidebands along with an optical carrier λ_c are generated when the master laser is frequency modulated with a frequency of f_{RF} . The modulated signal is then injected into two slave lasers whose free-running wavelengths are selected to be as close as possible to any two separated optical sidebands and hence, the output wavelengths of the slave lasers are phase correlated. In this instance, we have chosen the -2^{nd} -order and $+2^{\text{nd}}$ -order sidebands for the optical injection locking. Beating these two wavelengths at a PD would thus

enable the generation of a low-phase noise CW microwave signal. It should be emphasized that the frequency of the generated microwave signal is always an integer multiple of f_{RF} .

In contrast, phase coherence between two optical signals obtained through the optical phase-locked loop technique is based on locking the phase of one CW laser to another via an electrical feedback loop circuit. The schematic of this approach is illustrated in Fig. 3.3. The phase of the heterodyned signal generated at the output of the PD is first compared with that of a microwave reference at a mixer and subsequently amplified and low-pass filtered. The output voltage at the end of the loop filter is an error voltage which is proportional to the phase difference between the heterodyned signal and the microwave reference. This error voltage is then fed back to control the phase of one of the CW laser diode. By carefully designing the feedback loop gain and response time, the phase of the generated CW microwave signal can be locked to the microwave reference. In both of the above techniques, an external high quality microwave reference source which can operate over a wide frequency range is essential. Furthermore, the phase noise of the generated CW microwave signal is largely dependent on the phase noise of the microwave reference source.

3.2.2 External modulation

The external modulation method is another effective method to generate high quality microwave signals [64-68]. One of the earlier demonstrations [65] is based on the frequency-doubling approach which involves biasing a MZM at the minimum transmission point of its transfer function so that the optical carrier and even-order optical sidebands are suppressed. For a modulation frequency of f_{RF} , by beating the

two 1st-order optical sidebands at the PD, a frequency-doubled CW microwave signal of $2f_{RF}$ is generated.

The schematic of this approach is depicted in Fig. 3.4. Assuming that the $V_{RF}(t) = V_0 \cos \omega_{RF} t$ is the applied input RF signal to the MZM, where V_0 and ω_{RF} are the amplitude and angular frequency of the RF signal, respectively, the electric field at the output of the MZM can be expressed as

$$\begin{aligned} E_1(t) &= E_o \cos\left(\frac{\phi_o}{2} + \frac{k}{2} \cos \omega_{RF} t\right) \exp(j\omega_c t) \\ &= E_o \left[\cos\left(\frac{\phi_o}{2}\right) \cos\left(\frac{k}{2} \cos \omega_{RF} t\right) - \sin\left(\frac{\phi_o}{2}\right) \sin\left(\frac{k}{2} \cos \omega_{RF} t\right) \right] \exp(j\omega_c t) \end{aligned} \quad (3.4)$$

where, $E_o \exp(j\omega_c t)$ is the input optical carrier which has an amplitude of E_o and angular frequency ω_c , $k = \pi V_0 / V_\pi$, V_π is the half-wave voltage of the MZM, and ϕ_o is a constant phase shift determined by the DC bias voltage.

Fig. 3.4 Schematic of the frequency-doubling approach based on biasing the MZM at the minimum transmission point.

The DC bias voltage is tuned to a value $\phi_o = \pi, 3\pi, 5\pi, \dots$ so that the even-order optical sidebands which are associated with the $\cos(0.5k \cos \omega_{RF} t)$ term in (3.4) will vanish. More precisely, this biasing point is in fact the minimum point of the MZM's transfer function. Using the Bessel function identity in (2.6), the electric field can then be expressed as

$$\begin{aligned} E_1(t) &= E_o \sin\left(\frac{k}{2} \cos \omega_{RF} t\right) \exp(j\omega_o t) \\ &= 2E_o \exp(j\omega_o t) \sum_{n=1}^{\infty} J_{2n-1}(k) (-1)^{n+1} \cos(2n-1)\omega_{RF} t \end{aligned} \quad (3.5)$$

where $J_n(k)$ is the n^{th} order Bessel function of the first kind. In order to only generate the frequency-doubled CW microwave signal at the output of the PD, it is required to drive the input RF signal to an appropriate power level such that the amplitude of all the harmonics above the 1st-order harmonic can be considered negligible. The photocurrent can then be approximated as

$$\begin{aligned} I_1(t) &= E_1(t) \cdot E_1^*(t) \\ &\propto \cos^2 \omega_{RF} t \\ &\propto \frac{1}{2} [1 + \cos(2\omega_{RF} t)] \end{aligned} \quad (3.6)$$

Here, we can see that the expression for the photocurrent constitutes a DC term and a frequency-doubled CW microwave signal represented by the 2nd-order harmonics of the applied angular RF frequency ω_{RF} . As demonstrated in [65], a 36 GHz mm-wave signal was generated when the MZM was driven by a 18 GHz input RF signal.

Fig. 3.5 Schematic of the frequency-quadrupling approach based on biasing the MZM at the maximum transmission point.

Later, it was demonstrated that by suppressing the odd-order optical sidebands through biasing the MZM at the maximum transmission point and filtering the optical carrier using a wavelength-fixed notch filter, a frequency-quadrupled microwave signal can be generated at the output of the PD [67]. The schematic of this frequency-quadrupling approach is depicted in Fig. 3.5.

Unlike the previous approach, this approach is based on biasing the MZM at the maximum transmission point on its transfer function so that $\phi_o = 0, 2\pi, 4\pi, \dots$. As a result, the output of the MZM comprises the optical carrier and the even-order optical sidebands while all odd-order optical sidebands which are associated with the $\sin(0.5k \cos \omega_{RF}t)$ term in (3.4) vanish. Using the Bessel function identity in (2.7), the electric field at the output of the MZM for this approach is given by

$$\begin{aligned}
 E_1(t) &= E_o \cos\left(\frac{k}{2} \cos \omega_{RF}t\right) \exp(j\omega_c t) \\
 &= E_o \exp(j\omega_c t) \left[J_o(k) + 2 \sum_{n=1}^{\infty} J_{2n}(k) (-1)^n \cos 2n\omega_{RF}t \right]
 \end{aligned} \tag{3.7}$$

Likewise, it is required to drive the input RF signal to an appropriate power level such that the amplitude of all the harmonics above the 2nd-order harmonic can be considered negligible. After filtering the optical carrier using a notch filter, the photocurrent can be approximated as

$$\begin{aligned} I_1(t) &= E_1(t) \cdot E_1^*(t) \\ &\propto \cos^2 2\omega_{RF}t \\ &\propto \frac{1}{2} [1 + \cos(4\omega_{RF}t)] \end{aligned} \quad (3.8)$$

The photocurrent includes a frequency-quadrupled CW microwave signal. The frequency-quadrupling approach was demonstrated in [67] where a 32 to 50 GHz mm-wave signal was generated when the input RF signal to the MZM was tuned from 8 to 12.5 GHz.

More recently, it has been shown that an architecture which involves cascading two MZMs with a tunable optical phase shifter in between can enable the generation of a frequency-sextupled CW microwave signal [69]. This is realized when an appropriate value is set for the tunable phase shifter as well as biasing one MZM at the minimum and the other at maximum transmission point. An improvement to this scheme, biasing both the MZMs at the maximum transmission point has also been demonstrated to generate a frequency-octupled CW microwave signal [70]. In all of the aforementioned techniques based on external modulation, frequency-tunability is largely related to the frequency of the input RF signal, and the generated frequency is limited by the bandwidth of the MZM or PD. Furthermore, the phase noise of the generated microwave signal is largely dependent on the quality of the input RF signal to the MZM which is generally supplied by an external microwave source.

3.2.3 Dual-wavelength laser

Somewhat similar to the optical heterodyning technique, the dual-wavelength laser approach is also based on the beating of two separated wavelengths at the PD. However, as the two lasing wavelengths coexist in the same gain cavity for a dual-wavelength laser, the phase correlation between them is much better than the two free running CW lasers used in the optical heterodyning approach. Compared to the optical injection locking and optical phase-locked loop techniques though, the phase noise of the generated microwave signal is generally higher since the two wavelengths in a dual-wavelength laser are not tightly phase-locked together.

Typically, a dual-wavelength laser utilizes FBGs in a ring laser cavity setup so as to discriminate two individual longitudinal modes. An example of a dual-wavelength laser is shown in Fig. 3.6. To avoid homogenous line broadening, a semiconductor optical amplifier (SOA) is used in place of an erbium doped fiber amplifier (EDFA). FBG1 is an equivalent-phase-shift FBG (EPS-FBG) which has two ultra-narrow bandwidth passbands that are used to limit one longitudinal mode in each passband.

Fig. 3.6 Schematic of the dual-wavelength laser for generation of CW microwave signal. PC: polarization controller; FBG: fiber-Bragg gratings; SOA: semiconductor optical amplifier.

On the other hand, FBG2 comprises two superimposed regular FBGs having two different center wavelengths. Hence, the ultra-narrow transmission passbands of FBG1 can be filtered out by the two transmission passbands of FBG2 and subsequently the two filtered lasing wavelengths can be heterodyned at a PD. Chen *et al.* demonstrated the generation of a ~ 18.4 GHz microwave signal based on the dual-wavelength laser setup in which the two lasing wavelengths are separated by 0.147 nm [71].

In another variation of a dual-wavelength laser, Yao *et al.* have also demonstrated the generation of a ~ 3.389 GHz microwave signal by incorporating a dual-phase-shift FBG with two ultra-narrow transmission passbands in a simple linear laser cavity [72]. Generation of frequency-tunable CW microwave signals is also demonstrated in [73-75] and it is noted that the frequency-tunability depends largely on the ability to tune the incorporated FBGs to the center wavelengths which create the desired wavelength spacing. Generally, such FBGs can be designed and fabricated in laboratories or purchased commercially. Although modulators and microwave sources are not required for the dual-wavelength laser, the fabrication process for the FBGs is often complex and they are also costly when purchased.

3.2.4 Direct photodetection of a multi-longitudinal mode laser

Besides the previously mentioned approaches, in recent year, it has been shown that direct photodetection of a multi-longitudinal mode laser is also capable of generating frequency-tunable CW microwave signals [95-98]. In particular, mode-locked lasers which consist of equally-spaced and phase-locked longitudinal modes are of great interest in this context. The RF spectrum of a mode-locked laser consists of a fundamental frequency determined by the repetition rate of the mode-locked

laser, as well as its harmonics which are within the bandwidth of the photodetector. By beating these phase-locked modes or harmonics, the phase fluctuations between them can be cancelled, thus enabling the generation of a microwave signal with low-phase noise.

Both passive [96] and active [95] mode-locked lasers have been demonstrated to be suitable choices for this scheme. For microwave generation using passive mode-locked laser, external microwave source and modulators are not required, thus simplifying the implementation based on this approach. In the case of an active mode-locked laser, although external microwave source and modulators are required, they do not need to operate at a very high frequency. By using a regenerative active mode-locked laser, one can completely eliminate the use of an external microwave source.

Since the RF spectrum of a mode-locked laser consists of a series of discrete frequency components, by applying an appropriate microwave bandpass filter centered at the frequency of interest, the desired microwave signal can be extracted. This scheme can be improved further by implementing an array of microwave bandpass filters and switches at the output of the photodetector, which would thus offer a certain degree of frequency-tunability. Although the use of a mode-locked laser to generate frequency-tunable and low-phase noise CW microwave signals is relatively less complex than all the previously mentioned techniques, using an array of microwave bandpass filters and switches to obtain frequency-tunability however does not offer flexibility in the system design and inevitably adds to the cost of the whole system due to additional RF components.

Table 3-1 Summary of the different techniques of photonic generation of microwave signals.

Techniques	Advantages	Disadvantages
Optical heterodyning	• Simple to implement. • Tunable frequency. • Does not require the use of external microwave source.	• High phase noise, due to uncorrelated nature of CW lasers. • Frequency tunability limited by bandwidth of PD.
Optical injection locking	• Tunable frequency. • Low phase noise, quality depends on and limited by external microwave source.	• Requires the use of external microwave source and a master laser. • Frequency is an integer multiple of input RF frequency. • Frequency tunability limited by bandwidth of PD.
Optical phase-locked loop	• Tunable frequency. • Low phase noise, quality depends on and limited by external microwave source.	• Requires the use of external microwave source and RF components for feedback loop control. • Frequency is an integer multiple of input RF frequency. • Frequency tunability limited by bandwidth of PD.
External modulation	• Tunable frequency with frequency M-times of input RF frequency depending on bias point on modulator. • Low phase noise, quality depends on external microwave source.	• Requires the use of external microwave source, modulators and notch filter. • Frequency tunability limited by bandwidth of PD and modulator.
Dual-wavelength laser	• Tunable frequency. • Low phase noise, better than optical heterodyning but no better than optical injection locking & optical phase locked loop. • Does not require external microwave source.	• Requires the use of ultra-narrow bandwidth FBGs, difficult to fabricate or costly. • Frequency tunability limited by ability to tune the FBGs.
Direct photodetection of multi-longitudinal mode laser	• Tunable frequency. • Low phase noise due to beating of phase locked modes. • Does not require external microwave source for passive mode-locked lasers.	• Frequency tunability requires the use of an array of microwave bandpass filter and switches • Requires external microwave source and modulator for active mode-locked lasers.

Therefore, in order to fully utilize the advantageous features of using a mode-locked laser to generate CW microwave signals, it is important to resolve concerns on the frequency-tunability for this approach. In chapter 6, we discuss two techniques which enable us to generate frequency-tunable CW microwave signals without using any microwave bandpass filters, switches as well as modulators and external microwave sources.

Table. 3-1 summarizes the advantages and disadvantages of the different techniques of photonic generation of microwave signals which have been discussed in this chapter so far.

3.3 Photonic generation of chirped microwave signals

Besides CW microwave signals, microwave arbitrary waveforms are also widely-used in sophisticated applications such as radar, wireless communication, satellite communication, remote sensing, and navigation. However, due to the limitations in the DAC and timing jitter, the generation of microwave arbitrary waveforms in the electrical domain is often limited to low frequency and small bandwidth [58]. Fortunately, various photonic techniques developed in recent years have shown the capacity to generate microwave arbitrary waveforms in the GHz and tens of GHz range, thus deeming them as viable solutions to the aforementioned electronic limitations.

For instance, on the basis of converting a spatially distributed pattern directly to a temporally distributed pattern, the direct-space-to-time (DST) pulse shaping technique involves the manipulation of a single pulse to a pulse burst with controllable amplitudes and pulse-to-pulse spacing so as to generate reconfigurable microwave waveforms which exhibit amplitude or phase modulation [77-79]. The key

to this approach is a DST pulse shaper which is typically implemented using free-space optical components that are usually bulky, lossy, and unstable. An alternative method to generate microwave arbitrary waveforms is the temporal pulse shaping technique.

The temporal pulse shaping technique is based on dispersive compression or expansion of highly chirped optical pulses which are amplitude [80, 81] or phase modulated [82, 83] by an input microwave signal. The highly chirped pulses are typically generated by chromatically dispersing optical pulses from a mode-locked laser using a dispersive medium such as SMF. The advantage of this approach is that a high frequency microwave arbitrary waveform can be generated by using a low frequency modulating signal, but at the expense of using modulators and microwave sources.

Another method which can realize microwave arbitrary waveforms is the optical line-by-line pulse shaping technique which involves controlling the amplitude and phase of each frequency comb line in a mode-locked laser [84-87]. The key advantage of this technique is that the generated microwave arbitrary waveforms can span the entire time period between the pulses; in other words, 100% duty cycle can be obtained. In conventional optical pulse shapers, a liquid crystal based spatial light modulator (SLM) is usually tasked to resolve and manipulate the frequency comb lines independently. However, one drawback associated with this technique is that the waveform update rate is limited by the response time of the SLM, which is typically of the order of tens of milliseconds.

Fig. 3.7 Schematic of a microwave arbitrary waveform generator based on the optical spectral shaping and wavelength-to-time mapping technique.

One method which is widely-used to generate microwave arbitrary waveform is the optical spectral shaping and wavelength-to-time mapping technique [88-93]. A straightforward approach, an optical spectral shaper is first used to shape the optical spectrum of a broadband ultrashort optical pulse and the spectrally shaped optical pulse is propagated through a dispersive medium which performs wavelength-to-time mapping on it. Lastly, the spectrally shaped and temporally-stretched optical pulse is fed to a PD for optical-to-electrical (O/E) conversion. Fig. 3.7 illustrates a typical microwave arbitrary waveform generator based on this technique. An important aspect in this technique, the temporal profile of the generated waveform is in fact proportional to the spectrum profile of the spectrally shaped pulse before wavelength-to-time mapping is performed on it. Therefore, by properly designing the spectral response of the optical spectral shaper in the frequency domain, the desired microwave arbitrary waveform can be generated. The details of the analysis for this technique are presented in the following section.

3.3.1 Optical spectral shaping and wavelength-to-time-mapping technique

Defining the optical pulse after spectral filtering to be $E_1(t)$ with a temporal width Δt_o , the electric field $E_2(t)$ at the output of the dispersive device is given as [94]

$$\begin{aligned}
 E_2(t) &= E_1(t) * \exp\left(j \frac{t^2}{2GDD}\right) \\
 &= \int_{-\infty}^{\infty} E_1(\tau) \times \exp\left(j \frac{(t-\tau)^2}{2GDD}\right) d\tau \\
 &= \exp\left(j \frac{t^2}{2GDD}\right) \times \int_{-\infty}^{\infty} E_1(\tau) \times \exp\left(j \frac{\tau^2}{2GDD}\right) \times \exp\left[-j \left(\frac{t}{GDD}\right) \tau\right] d\tau \quad (3.9) \\
 &\approx \exp\left(j \frac{t^2}{2GDD}\right) \times \int_{-\infty}^{\infty} E_1(\tau) \times \exp\left[-j \left(\frac{t}{GDD}\right) \tau\right] d\tau \\
 &= \exp\left(j \frac{t^2}{2GDD}\right) \times E_1(\omega) \Big|_{\omega = \frac{t}{GDD}}
 \end{aligned}$$

where we have considered only the effects of the 2nd-order dispersion β_2 from the dispersive device, GDD is the group delay dispersion of the dispersive device and $E_1(\omega)$ is the Fourier transform of $E_1(t)$. If a length L of SMF is used as the dispersive device, its group delay dispersion $GDD = \beta_2 L$.

As observed from (3.9), the temporal profile of $E_2(t)$ is proportional to the $E_1(\omega)$ and this is only valid when the following condition is satisfied

$$\frac{\tau^2}{2GDD} \leq \frac{\Delta t_o^2}{2GDD} \ll 1 \quad (3.10)$$

Therefore, to generate a desired temporal profile $E_2(t)$, one just needs to design an optical spectral shaper whose spectral response is proportional to $E_2(t)$ as well as

choose a dispersive device with an appropriate GDD that satisfies (3.10). In particular, much effort has been placed on designing suitable optical spectral shaper that can generate chirped microwave signals based on this technique. The chirped microwave signals find use in modern radar systems, wherein pulse compression on these microwave signals using a matched filter can increase the range resolution. As the frequency of a chirped microwave signal increases (up-chirp) or decreases (down-chirp) with respect to time, the magnitude response of the optical spectral shaper should have an increasing or decreasing free spectral range (FSR), respectively.

One of the methods that can be used to design such a magnitude response is to use LCFBGs with different chirp rates in an unbalanced MZI as an optical spectral-shaper [90]. Besides demonstrating the ability to generate linearly-chirped microwave signals, this scheme also allows the centre frequency of the generated microwave signals to be tuned. The only major limitation of this approach is that the chirp rate of the microwave signals is not tunable as the fabricated LCFBGs are not reconfigurable. Much effort has been put into developing optical spectral shapers which can enable the chirp rate to be tuned as well [91-93]. Recently, Li *et al.* have shown that by using an optically pumped LCFBG which is written in an Erbium-Ytterbium co-doped fiber into an unbalanced MZI setup, both centre frequency and chirp rate of the generated chirped microwave signals can be tuned independently [93]. Potentially, this scheme allows the generation of positively chirped, zero, and negatively chirped microwave signals. However, due to the limitations in the LCFBG which they have used, they were unable to show the zero-chirp case as well as achieve a wide chirp tuning range.

In Chapter 7, we propose and demonstrate the generation of linearly-chirped microwave signals whose chirp rate can be tuned to positive, zero and negative values

by incorporating a LCFBG with programmable dispersion into a setup similar to that in [93].

3.4 Summary

In this chapter, we have reviewed and described the working principle of a number of photonic approaches which enable the generation of CW microwave signals. These approaches include optical heterodyning and its two variants - optical injection locking and optical phase-locked loop, external modulation, dual-wavelength laser, and direct photodetection of a multi-longitudinal mode laser. Unlike the other approaches, the direct photodetection of a multi-longitudinal mode laser potentially does not involve the use of any high-speed modulators and microwave reference sources, yet is able to generate frequency-tunable and low-phase noise CW microwave signals. Practical implementation is also simpler as complex components such as FBGs are not required in the setup. However, the frequency-tunability in this approach requires the use of microwave bandpass filters and switches which renders the system bulky and expensive when a large tuning range is required.

In chapter 6, we propose and demonstrate two techniques to generate CW frequency tunable CW microwave signals without using any microwave bandpass filters and switches as well as modulators and external microwave sources.

We have also discussed the generation of chirped microwave signals. Specifically, we have considered in detail the optical spectral shaping and wavelength-to-time mapping technique as it is simple to implement in practice and has delivered promising results. Based on the understanding of its working principle, there have been a number of approaches developed with the goal to design an optical

spectral shaper with chirped FSR. However, none of these demonstrations have been able to generate linearly-chirped microwave signals whose chirp rate can be tuned across a wide range of values that covers the positive, zero and negative region.

In chapter 7, we propose and demonstrate a technique to generate linearly-chirped microwave signals whose chirp rate can be tuned to positive, zero and negative values.

Chapter 4 OPTICALLY SAMPLED ADC AND PHOTONIC TIME-STRETCHED ADC

This chapter describes and discusses the demonstration of two Photonic ADC schemes.

In the first demonstration, we propose and show experimentally a 2 GS/s optically sampled analog-to-digital converter based on dual-output modulation and balanced photodetection. Finally, arcsine operation is digitally implemented on the output of the electronic ADC and is shown to enable significant odd-order harmonic suppression at high modulation depths approaching unity.

In the second demonstration, we propose and show analytically as well as experimentally that a polarization modulator which supports TE and TM modes of opposite phase modulation indices can be utilized to reject dispersion-induced even-order distortions in a photonic time-stretched ADC (TS-ADC). Continuous-time operation based on this scheme is also described using the virtual time gating principle. Based on a time-stretch factor of 4 and an electronic ADC with a sampling rate of 2 GS/s, we demonstrate digitization of RF signals with a single channel. With adequate backend digitizing hardware, a four-channel continuous-time TS-ADC with a sampling rate of 8 Gs/s can be realized to handle RF frequencies as high as 4 GHz.

4.1 Optically-Sampled ADC based on Complementary Dual-Output Modulation and Balanced Photodetection

Unlike conventional coaxial cable links, analog photonic links offer wide bandwidth capability, immunity to electromagnetic interference, low-loss transmission, small system size as well as the potential for highly parallel operation [102, 103]. These favorable attributes of analog photonic links coupled with the availability of low-loss optical fiber as well as modulation devices with high modulation efficiency have resulted in an increase in the optical fiber transmission of RF and microwave signals within the last decade. In particular, such links have found strong application in the field of photonic A/D conversion when implemented with electronic ADCs.

Among the many photonic ADC architectures proposed and demonstrated in recent years, optically-sampled ADCs [16-21] have received extensive attention as these display remarkable potential in terms of enhancing either the resolution or sampling rate of the electronic ADCs. As mentioned in Chapters 1 and 2, the enhancement in the resolution and sampling rate are realized through optical sampling using low timing-jitter pulses which can circumvent the jitter-limited ENOB issues commonly associated with electronic ADCs and the various multi-channel interleaving schemes, respectively.

Typically, intensity modulation is employed in the optically-sampled ADCs wherein an RF signal externally modulates an optical pulse-train using a single-output electro-optic MZM. As the MZM's raised-cosine transfer function is nonlinear, the modulator is generally biased at the quadrature point such that the resulting analog signal waveform can be depleted of any even-order distortions ideally, and may have

only the odd-order distortions. Having a similar transfer function as the MZM, phase and polarization modulation schemes have also been demonstrated, both in the analog photonic links as well as the photonic ADC systems.

As mentioned in Chapter 2, in order to obtain a large SNR, optically-sampled ADC systems often operate at large modulation depth values. However, for sufficiently large modulation depth, 3rd-order distortion rises above the noise floor and severely reduces the SFDR of these systems, consequently deteriorating the accuracy during the A/D conversion. Fortunately, it has been shown that through the use of a digital linearization technique in the form of arcsine operation, the 3rd-order distortion limited SDFR of the optically-sampled ADC can be improved, thus allowing operation at large modulation depth [16]. This digital postprocessing approach offers a much simpler alternative to the implementation of a linearized modulator in the analog photonic link [104]. Implementation of the arcsine operation scheme enables one to obtain an undistorted RF signal provided the distortion is due to the inherent nonlinear characteristic of the modulator. Thus, the 3rd-order distortions which are caused mainly by the raised cosine transfer function of the MZM can be suppressed well below the noise floor, thus enhancing the SFDR in the process.

One of the problems associated with the use of the single-output MZM in optically-sampled ADC schemes is the inability to match the dynamic range of the analog photodetected signal to that of the peak-to-peak full-scale voltage of the backend electronic ADC. However, this problem can be resolved by replacing the single-output MZM with a complementary dual-output MZM. Incorporating the complementary dual-output MZM together with the arcsine operation, phase-encoded optical sampling has been demonstrated in [22-24]. Besides the benefit of suppressing the odd-order distortions due to the nonlinear transfer function of the complementary

dual-output MZM, application of the arcsine operation on the outputs also helps to suppress the amplitude noise from the input pulse-train, increases the SNR by 6 dB, and removes the DC terms.

The phase-encoded optical sampling schemes demonstrated in [22-24] have used twice the number of well-calibrated electronic detection and quantization channels as compared to the conventional optical sampling based on single-output MZM. This is the primary drawback of the phase-encoded optical sampling scheme and to overcome it, in the next section, we propose and demonstrate a phase-encoded optically-sampled ADC which utilizes balanced photodetection on the output of the complementary dual-output MZM.

4.1.1 Principle of operation and mathematical expressions

Fig. 4.1(a) illustrates the schematic of the optically sampled photonic ADC based on dual-output modulation and balanced photodetection scheme. Here, we have used a picosecond (15 ps) pulse-train generated by a home-made 2 GHz regenerative active mode-locked fiber laser as the initial source. The average power of the pulse-train is noted to be approximately 10 dBm. Fig.4.1(b) depicts the SSB phase noise $L(f)$ of the regenerative active mode-locked fiber laser measured with a phase noise analyzer (R&S-FSUP). Its associated RMS timing jitter Δt_j can then be calculated by computing the following integral [24]

$$\Delta t_j = \frac{1}{2\pi f_{rep}} \sqrt{2 \int_{\Delta f_L}^{\Delta f_H} L(f) df} \quad (4.1)$$

Fig. 4.1 (a) Schematic of the optically sampled photonic ADC based on dual-output modulation and balanced photodetection scheme. (b) Comparison of the SSB phase noise between the regenerative active mode-locked laser and a currently available RF signal generator .

where f_{rep} is the repetition rate of the laser and Δf_H and Δf_L are the upper and lower integration range, respectively. Using an integration range of $\Delta f_L = 100$ Hz and $\Delta f_H = 30$ MHz, we obtained a $\Delta t_j \approx 80$ fs for the regenerative active mode-locked laser which is fairly comparable to mid to high-end RF signal generators. For comparison, we have also plotted the $L(f)$ of a RF signal generator operating at 2 GHz in Fig. 4.1(b). Using the same integration range, we obtained a $\Delta t_j \approx 148$ fs for

the RF signal generator. It should be pointed out that, this is the only signal generator [Anritsu 68367C) available to us at this point of time. Next, the pulse-train is modulated by a complementary dual-output MZM and the modulated pulse-train is then fed to a balanced photodetector for O/E conversion. Following that, the electrical output is digitized utilizing a state-of-the-art 2 GS/s electronic ADC which is clocked by the same 2 GHz pulse-train. This electronic ADC is specified to have 10-bit resolution at 2 GS/s; the measured ENOB is approximately 7 at 1 GHz for –1 dB full-scale, where the full-scale peak-to-peak voltage is 500 mV.

Considering that $V_{in}(t) = V_o \cos \omega_{RF} t$ is the applied input RF voltage where ω_{RF} and V_o are its angular frequency and peak voltage, respectively, the output optical power at the complementary arms of the dual-output MZM when biased at the quadrature point, can be expressed as

$$P_{out,+}(t) = \frac{P_o}{2} \left[1 + \sin(k \cos \omega_{RF} t + \theta_e) \right] \quad (4.2)$$

$$P_{out,-}(t) = \frac{P_o}{2} \left[1 - \sin(k \cos \omega_{RF} t + \theta_e) \right] \quad (4.3)$$

where $P_o/2$ is the output optical power of each arm, $k = \pi V_o / V_\pi$, and V_π is the half-wave voltage of the complementary dual-output MZM. Here, we have also included the quadrature phase error θ_e associated with the deviation of bias voltage from the quadrature point.

The photocurrent $I(t)$ at the output of the balanced photodetector is given by

$$\begin{aligned} I(t) &= R_+ P_{out,+}(t) - R_- P_{out,-}(t) \\ &= \frac{\Delta R P_o}{2} + P_o R \sin(k \cos \omega_{RF} t + \theta_e) \end{aligned} \quad (4.4)$$

where R_+ and R_- are the responsivity of the individual photodiode of the balanced photodetector, and ΔR is the imbalance between R_+ and R_- . Here, we observe that the imbalance in the responsivity only results in a DC term in the photocurrent. Through the balanced photodetection process, the common-mode relative intensity noise (RIN) originating from the 2 GHz pulse source can also be suppressed to a certain extent [105-107].

Typically, an RF link comprising components such as RF cable, connector, amplifier, DC block and low-pass filter is placed in between the output of the balanced photodetector and the backend electronic ADC and its purpose is to match the photodetected voltage to the peak-to-peak full-scale voltage of the electronic ADC. Assuming that the RF link has an effective gain of $\sqrt{G_{eff}}$ with a flat response over the Nyquist bandwidth of the electronic ADC and Z_{out} is the output impedance, the output voltage can be expressed as

$$\begin{aligned} V_{out}(t) &= \sqrt{G_{eff}} \cdot I(t) \cdot Z_{out} \\ &= V_{full} \cdot \sin(k \cos \omega_{RF} t + \theta_e) \\ &= V_{full} \cdot \sin\left(\frac{\pi V_{in}(t)}{V_\pi} + \theta_e\right) \end{aligned} \quad (4.5)$$

where V_{full} is the detected voltage corresponding to P_o . Due to the sinusoidal transfer function of the modulator, $V_{out}(t)$ represents a slightly distorted version of $V_{in}(t)$, with the SFDR often limited by the 3rd-order distortion when θ_e is zero.

Following the photodetection, $V_{out}(t)$ is quantized using a 2 GS/s electronic ADC which is clocked by the 2 GHz pulse source. The applied input voltage can then be

determined by implementing the arcsine operation on $V_{out}(t)$. This linearization process can be expressed by

$$V_{in}(t_n) = \frac{V_\pi}{\pi} \left[\sin^{-1} \left(\frac{V_{out}(t_n)}{V_{full}} \right) - \theta_e \right] \quad (4.6)$$

where $V_{out}(t_n)$ and $V_{in}(t_n)$ are the quantized output voltage and the recovered input voltage at time instant t_n , respectively. It is noted that $V_{in}(t_n)$ should reproduce an undistorted discrete-time replica of the applied input voltage $V_{in}(t)$ after linearization. This is only attainable provided that the following criteria are satisfied.

First, the distortions in $V_{out}(t)$ have to be attributable only to the sinusoidal transfer function of the dual-output MZM. This clearly implies that the linearity of the RF link that connects the balanced photodetector and electronic ADC has to be maintained. Secondly, V_{full} has to be measured as accurately as possible; this can be achieved by utilizing a high resolution electronic oscilloscope for measurement. Since $V_{full}/2$ is in fact associated with the peak voltage of an unmodulated pulse-train in one arm of the complementary dual-output MZM, V_{full} is simply calculated as the sum of the measured peak voltages of each arm.

It should be emphasized that any deviation from the aforementioned criteria can significantly limit the accuracy of the linearization process. This can be reflected by the presence of the 3rd-order distortion in the FFT spectrum of $V_{in}(t_n)$. It should also be noted that the phase error θ_e introduces a DC offset term but it does not affect the linearization process as it can be removed by a DC block.

4.1.2 Measured results and discussion

To examine the accuracy of the linearization technique in the proposed scheme, we set the modulation depth $m_{dep} = 2V_o / V_\pi$ to a high value of about 0.75 which would normally result in prominent 3rd-order distortion. Fig. 4.2(a) illustrates the FFT spectrum of a 200 MHz tone with the fundamental and the distortion tones clearly visible. The SNR of this optically sampled ADC measured over 1 GHz bandwidth is ~ 35 dB which works out to be 5.52 SNR-bits using $SNR_{bit} = [SNR - 1.76] / 6.02$. It is noteworthy that the spurs with 20 MHz interval as shown in the FFT spectrum are due

Fig. 4.2 Single-tone at 200 MHz for (a) nonlinearized and (b) linearized FFT data.

Fig. 4.3 Relationship between ENOB and analog input frequency before (blue) and after (after) arcsine operation for undersampling mode.

to the supermode noise of the 2 GHz pulse source for which the fundamental cavity frequency is 20 MHz. The SINAD is noted be ~ 27.56 dB and correspondingly, we obtained an ENOB of 4.29-bits. It is important to point out that this low ENOB is currently limited by both the supermode noise and more importantly, the 3rd-order distortions which are due to the nonlinearity of the dual-output MZM. Examining closer, the SFDR of the current system is limited at ~ 28 dBc by the 3rd-order distortions.

Next, we apply the arcsine operation according to (4.6) on the digitized data. Using an electronic oscilloscope, V_{full} is measured and calculated to be ~ 124 mV. Fig. 4.2(b) illustrates the linearized FFT spectrum for $V_{in}(t_n)$ where we observe the 3rd-order distortion is now suppressed by more than 20 dB to ~ -63 dBc. The SFDR in this case is measured to be ~ 52 dB which is now limited by the spurs caused by the supermode noise. This also led to an improved ENOB of ~ 5.50 -bits, which is almost equal to the 5.52 SNR-bits, thus indicating that the system is now limited by the SNR,

or in other words the supermode noise of the laser. Evaluating the current system further, we sampled a series of input frequencies (20 MHz to 18 GHz) and performed the arcsine operation on them. It should be emphasized that the current system operates in the undersampling mode as most signals are sampled below the Nyquist rate of the ADC used. Fig.4.3 illustrates the relationship between ENOB and analog input frequency before and after arcsine operation for our current system. We clearly observed an improvement in ENOB through the implementation of the linearization process. This led us to obtaining a 2 GS/s optically-sampled ADC system with an average ENOB of ~5.34-bits, operating in undersampling mode.

Fig. 4.4 “Walden chart” illustrating ENOB of existing ADCs versus analog input frequency.

Light and dark blue circles correspond to the electronic ADCs surveyed by [5] and [27], respectively. Orange stars are some of the high-performance wideband optically-sampled ADCs demonstrated so far, with the bigger star representing the current state-of-the-art performance [27]. Red circle corresponds to the performance of our current system. All optically-sampled ADCs in this figure operate in undersampling mode. This figure is reproduced from [27].

For the sake of comparison, our current system (red circle) is also plotted in the "Walden Chart" as shown in Fig. 4.4. This chart has been presented earlier in Fig. 2.6 and clearly depicts the current state-of-the-art optically-sampled ADC which has a sampling rate of 2 GS/s and an ENOB of 7-bits when undersampling a 41 GHz analog input frequency [27] (bigger orange star). Although the ENOB has been improved through the linearization process, the ENOB is still found to be limited as illustrated in Fig. 4.3. Thus, to improve our current system further, suppression of the supermode noise through better fiber laser design, together with appropriate matching of the photodetected signal to the dynamic range of the electronic ADC have to be considered. All of these can effectively increase both the SNR and SINAD, and more importantly improve the ENOB of the optically sampled ADC system.

4.2 Photonic Time-Stretched ADC based on Polarization Modulation and Balanced Photodetection

Different from the optically-sampled ADC, which address the jitter-limitation of the electronic ADCs, photonic TS-ADCs have been proposed and demonstrated in the recent years with the motivation to enhance the effective sampling rate and input bandwidth of the electronic ADC [39-45]. Due to its simplicity and a configuration similar to that for analog photonic links, the photonic TS-ADC technique has attracted significant research interest recently.

As mentioned in Chapter 2, in this technique, an RF signal which is superimposed on a linearly-chirped optical pulse can be stretched in time without significant distortion when the pulse propagates through a dispersive fiber. The time stretching is realized by exploiting the chromatic dispersion in the fiber. The time stretching reduces the bandwidth of the RF signal. Thus, the effective sampling rate

and input bandwidth of the electronic ADC increase in proportion to the stretch factor [45]. The photonic TS-ADC technique has been researched intensively both for single-shot and continuous-time operation [49, 51]. However, like most analog photonic links, dispersion-induced harmonic and intermodulation distortions can arise due to the inherent nonlinear characteristics of the modulators.

In the case of MZM, an additional source of 2nd-order distortion can also arise due to the bias-point drift. These unwanted distortions can severely reduce the dynamic range and consequently deteriorate the accuracy during A/D conversion. To avoid these nonlinear distortions, which are frequency dependent, one cost-effective solution is to operate at sub-octave bandwidths. However, this approach would defeat the objective of a broad bandwidth and high sampling rate.

It has been demonstrated for a TS-ADC that the complementary output pair of a push-pull MZM can be utilized in the differential mode to reject the dispersion-induced even-order distortions, thus permitting multi-octave bandwidth operation [54]. The differential operation approach can be accomplished by digital post-processing using an arcsine operation which also assists in the elimination of nonlinearities caused by the bias-point drift of the MZM. For this scheme, it should be noted that the complementary pulses from the MZM outputs are deliberately time delayed by half a repetition period, to ensure that no temporal overlapping of the pulses occurs before combining and time-stretching by a single dispersive channel. The time-stretched pulses are then required to align in time and are subsequently subtracted in the digital domain to obtain the desired differential mode.

Practically, ideal time-delay between the complementary pulses is difficult to achieve, and as a result amplitude and phase imbalances between the pulses arise,

affecting the accuracy of the differential operation. Further, because of the required time delay, continuous-time operation of the scheme by employing virtual time gating [51] is also difficult; to operate in continuous-time, the pulses from a pulse-train are required to broaden in time to at least an amount equal to the inter-pulse spacing prior to modulation. By stretching the complementary pulses in two separate dispersive fiber channels, the time delay procedure can be avoided. However, this approach will further complicate the hardware and channel mismatch problems will also arise. The use of a PolM in a photonic ADC has been proposed in [108]. This alternative avoids the problem of bias-drift that commonly exists in intensity modulation methods.

In this section, we show analytically and experimentally, that the complementary output pair of the PolM can also be exploited in a differential manner to reject the dispersion-induced even-order distortions for the TS-ADC. The differential operation is achieved by using a polarization beam-splitter (PBS) that generates complementary pulses which are fed to a balanced detector. This detection scheme also helps to mitigate the influence of RIN from the laser source [105-107] as well as to suppress the even-order photodiode-induced nonlinearities [109]. Since no time delay is involved, in contrast to the complementary output MZM scheme, the PolM approach is amenable to continuous-time operation by employing virtual time gating.

Fig. 4.5 Schematic setup of the TS-ADC based on polarization modulation with balanced detection scheme; PC: Polarization controller, PolM: Polarization modulator, PBS: Polarization beam-splitter.

4.2.1 Mathematical expressions and simulation results

Fig. 4.5 illustrates the schematic of the TS-ADC based on a PolM with balanced detection scheme. With the exception of the PolM and detection scheme, the backbone architecture of this TS-ADC is essentially similar to [45]. An optical pulse $E_1(t)$ is first linearly-chirped by propagating through a dispersive fiber of length L_1 . Using a polarization controller (PC1), the polarization state of the chirped optical pulse $E_2(t)$ is then oriented at an angle of 45° with respect to the x-axis of the PolM. Essentially a phase modulator, a PolM however supports both TE and TM modes with complementary phase modulation indices. Thus, the output electric field $E_3(t)$ of the PolM along the x- and y-axis can be expressed using the following Jones matrix formalism [108]:

$$E_3(t) = \begin{bmatrix} E_{3x} \\ E_{3y} \end{bmatrix} = \frac{1}{\sqrt{2}} E_2(t) \begin{bmatrix} \exp\left(j \frac{k}{2} \cos \omega_{RF} t\right) \\ \exp\left(-j \frac{k}{2} \cos \omega_{RF} t\right) \end{bmatrix} \quad (4.7)$$

where $k = \pi V_o / V_\pi$, $m_{dep} = 2V_o / V_\pi$ is defined as the modulation depth, V_π is the half-wave voltage, V_o and ω_{RF} are the voltage amplitude and angular frequency of the RF signal.

The phase modulated chirped pulse is then propagated through another dispersive fiber of length L_2 to attain $E_4(t)$. Using a second polarization controller (PC2), we align the PBS at an angle of 45° with respect to the principal axis of the PolM. Thus the orthogonal outputs of the PBS can be expressed as

$$E_{5+}(t) = \frac{1}{\sqrt{2}} \left[E_{4x}(t) \exp\left(j \frac{\phi_b}{2}\right) + E_{4y}(t) \exp\left(-j \frac{\phi_b}{2}\right) \right] \quad (4.8)$$

$$E_{5-}(t) = \frac{1}{\sqrt{2}} \left[E_{4x}(t) \exp\left(j \frac{\phi_b}{2}\right) - E_{4y}(t) \exp\left(-j \frac{\phi_b}{2}\right) \right] \quad (4.9)$$

where $\phi_b = \phi_s + \pi V_b / V_\pi$ is the PolM phase bias, V_b is the externally applied bias voltage and ϕ_s is the static bias phase-shift caused by PC2. Using the mathematical framework described in [45], the detector currents at the orthogonal outputs of the PBS at $\phi_b = 90^\circ$ can be expressed as

$$\begin{aligned}
 I_{5\pm}(t) = & \frac{1}{2} I_{Env}(t) \\
 & \times \left[\begin{aligned} & J_o^2\left(\frac{k}{2}\right) \mp 4J_o\left(\frac{k}{2}\right)J_1\left(\frac{k}{2}\right)\cos\phi_{DIP}\cos\frac{\omega_{RF}t}{M} \\ & + \begin{pmatrix} 2J_1^2\left(\frac{k}{2}\right) \\ -4J_o\left(\frac{k}{2}\right)J_2\left(\frac{k}{2}\right)\cos 4\phi_{DIP} \end{pmatrix} \cos\frac{2\omega_{RF}t}{M} \\ & \pm \begin{pmatrix} 4J_o\left(\frac{k}{2}\right)J_3\left(\frac{k}{2}\right)\cos 9\phi_{DIP} \\ +4J_1\left(\frac{k}{2}\right)J_2\left(\frac{k}{2}\right)\cos 3\phi_{DIP} \end{pmatrix} \cos\frac{3\omega_{RF}t}{M} \\ & + \dots \end{aligned} \right] \quad (4.10)
 \end{aligned}$$

where $\phi_{DIP} = \beta_2 \omega_{RF}^2 L_2 / 2$ is the dispersion-induced phase and $I_{Env}(t)$ is the photodetector current in the absence of modulation.

We note that the phase modulated signal is converted to an intensity modulated signal and the modulated signal is stretched by a factor of $M = (L_1 + L_2) / L_1$. Similar to the complementary output MZM scheme [54], the proposed scheme shows that the in-phase even-order harmonics can also be effectively cancelled by operating in the differential mode, $I = I_{5+} - I_{5-}$. To achieve differential operation, the complementary pulses are generated and aligned in time by a PBS. These complementary pulses can be received by separate single-ended detectors followed by digital post-processing, or alternatively, received by a balanced detector.

Fig. 4.6 Simulation results for a TS-ADC system. (a) Single-ended output showing 2nd-order and 3rd-order distortions. (b) Differential output based on balanced detection showing rejection of the 2nd-order distortion.

Fig. 4.6 illustrates the simulation results based on (4.10) for a single-tone signal with $m_{dep} = 0.7$, $L_2 = 27$ km, $D_S = 17$ ps/nm/km, $M = 4$ and input RF signal at 12 GHz. Clearly, the 12 GHz input signal has been time-stretched to a 3 GHz output signal and the differential output shows complete cancellation of the 2nd-order distortion as compared to the single-ended output. Under balanced detection, the RIN which couples to the complementary signal pair equally from the laser source can be mitigated [105-107]. The RIN would thus have no significant effect on the output RF signal. Additionally, since photodiodes are not absolutely linear devices, this

detection scheme can also suppress the even-order nonlinearities that may be induced by the photodiodes [109].

Another key significance of the proposed approach is that, unlike MZM which requires an externally applied voltage bias, the bias point of a PolM can be passively defined by just the static bias phase-shift value caused by PC2. Thus we can set $\phi_b = \phi_s$. This eliminates the bias-drift issues commonly associated with MZM, and ideally there should not be any additional source of even-order distortions arising from the PolM. Thus, the SFDR is primarily dominated by the 3rd-order distortions in the case when $\phi_s = 90^\circ$. The 3rd-order distortions can be cancelled substantially if we replace the balanced photodetection with conventional photodetection on each individual arm of the PBS, followed by post-digital linearization processing in the form of arcsine operation. Despite the inability to substantially cancel 3rd-order distortion, the balanced detection offers simplicity in the form of differential operation in the O/E conversion process. .

In practice, there can be misalignment in the state of polarization for the chirped optical pulse at the input of the PolM as well as misalignment in the angle that the PBS makes with the principal axis of the PolM. We can investigate the impact of these misalignments by rewriting Eq. (4.10) into a general form as shown in Eq. (4.11) in the case when no external voltage bias is applied to the PolM for small modulation depth:

$$\begin{aligned}
 I_{5\pm}(t) = & \frac{1}{2} \cos^2(45^\circ - \theta_1) \cdot I_{Env}(t) \\
 & \times \left[\cos^2\left(\frac{\phi_s}{2}\right) (A_\pm^2 + B_\pm^2) - k \cdot \cos \frac{\omega_{RF} t}{M} \left[\cos \phi_{DIP} \left\{ A_\pm^2 \sin\left(\frac{\phi_s}{2}\right) \cos\left(\frac{\phi_s}{2}\right) - B_\pm^2 \cos^2\left(\frac{\phi_s}{2}\right) \right\} \right. \right. \\
 & \left. \left. + A_\pm B_\pm \cdot \sin \phi_{DIP} \left\{ \cos^2\left(\frac{\phi_s}{2}\right) + \sin\left(\frac{\phi_s}{2}\right) \cos\left(\frac{\phi_s}{2}\right) \right\} \right] \right. \\
 & + \frac{k^2}{8} \cdot \cos \frac{2\omega_{RF} t}{M} \left[\begin{aligned} & A_\pm^2 \sin^2\left(\frac{\phi_s}{2}\right) + B_\pm^2 \cos^2\left(\frac{\phi_s}{2}\right) \\ & - \left\{ A_\pm^2 \cos^2\left(\frac{\phi_s}{2}\right) + B_\pm^2 \sin^2\left(\frac{\phi_s}{2}\right) \right\} \cos 4\phi_{DIP} \\ & - A_\pm B_\pm \left\{ \sin\left(\frac{\phi_s}{2}\right) \cos\left(\frac{\phi_s}{2}\right) - \cos^2\left(\frac{\phi_s}{2}\right) \right\} \sin 4\phi_{DIP} \end{aligned} \right] \\
 & + \frac{k^3}{6} \cdot \cos \frac{3\omega_{RF} t}{M} \left[\begin{aligned} & \left\{ A_\pm^2 \sin\left(\frac{\phi_s}{2}\right) \cos\left(\frac{\phi_s}{2}\right) - B_\pm^2 \cos^2\left(\frac{\phi_s}{2}\right) \right\} \cos 9\phi_{DIP} \\ & + A_\pm B_\pm \left\{ \cos^2\left(\frac{\phi_s}{2}\right) + \sin\left(\frac{\phi_s}{2}\right) \cos\left(\frac{\phi_s}{2}\right) \right\} \sin 9\phi_{DIP} \\ & + 3 \left\{ A_\pm^2 \sin\left(\frac{\phi_s}{2}\right) \cos\left(\frac{\phi_s}{2}\right) - B_\pm^2 \sin\left(\frac{\phi_s}{2}\right) \cos\left(\frac{\phi_s}{2}\right) \right\} \cos 3\phi_{DIP} \\ & + 3 A_\pm B_\pm \cdot \sin 3\phi_{DIP} \end{aligned} \right] + \dots \Big] \quad (4.11)
 \end{aligned}$$

Here, θ_1 is the angle at which the polarization state of the chirped optical pulse is aligned to the x-axis of the PolM, $A_\pm = \cos \theta_{2,\pm} + \sin \theta_{2,\pm}$, $B_\pm = \cos \theta_{2,\pm} - \sin \theta_{2,\pm}$, $\theta_{2,+}$ is the angle made by one arm of the PBS with the principal axis of the PolM and $\theta_{2,-} = \theta_{2,+} - 90^\circ$.

As the PolM has an in-built polarizer which is aligned at an angle of 45° with respect to its x-axis, any misalignment in θ_1 from 45° will only result in equal reduction in the amplitude for the orthogonal-polarized signals at the output arms of PBS. As the fundamental tone and distortions weaken at the same rate, the 2nd-order distortions will still be cancelled.

Fig. 4.7 Effects of misalignment in (a) θ_1 and (b) $\theta_{2,+}$ for the proposed differential operation.

Fig. 4.8 Effects of misalignment in static bias phase shift for the proposed differential operation.

Fig. 4.7(a) illustrates this result when θ_1 is misaligned. This is also similar to the case when $\theta_{2,+}$ is misaligned from 45° as illustrated in Fig. 4.7(b). The only

difference is that the 3rd-order distortion now weakens at a different rate from the fundamental tone. On the other hand, when the static bias phase-shift is not adjusted to $\phi_s = 90^\circ + n.360^\circ$, where n is an integer value, incomplete cancellation of 2nd-order distortion can occur as illustrated in Fig. 4.8. It is observed that an offset of more than 2 percent can result in the SFDR to be dominated by the 2nd-order distortion.

4.2.2 Measured results and discussion

Here, we describe a TS-ADC that operates towards continuous-time operation based on the PolM with balanced detection scheme. Fig. 4.9 shows the experimental setup of a four-channel continuous-time TS-ADC which is based on the virtual time gating principle. For clarity, a series of diagrams as presented in Fig. 4.10 is used to illustrate the operation. In the experiment, a 2 GHz regenerative active mode-locked fiber laser generating a pulse-train of 15 ps pulses is used. The pulse-train is amplified before being injected into 2 km of highly nonlinear fiber (HNLF) for SC generation. The SC spectrum is then filtered by a 3.2 nm bandpass filter centered at 1550 nm. This generates a pulse-train consisting of ultra-short broadband pulses [Fig. 4.10(a)] prior to the time-stretching process. These pulses are then linearly-chirped and broadened in time by an amount equal to the interpulse spacing which permits continuous-time operation [Fig. 4.10(b)]. This is achieved by propagating the broadband pulses through a length ($L_1 = 9455$ m) of SMF (SMF1) with total dispersion $D_{S,1}L_1 = -160.735$ ps/nm.

Fig. 4.9 Experimental setup of a TS-ADC towards continuous-time operation based on polarization modulation with balanced detection scheme. SMF: Single-mode-fiber; PC: Polarization controller; PBS: Polarization beam-splitter; EDFA: Erbium-doped fiber amplifier; WDM filter: Wavelength-division-multiplexer filter; eADC: Electronic analog-to-digital converter.

Fig. 4.10 Continuous-time TS-ADC based on virtual time gating principle (a) 2GHz broadband pulse-train, (b) Pulses stretched after SMF1, (c) Input RF signal, (d) Phase modulated optical pulses, (e) Pulses stretched after SMF2, (f1-f4) Intensity modulated pulses for channel 1-4 at one arm of PBS as well as corresponding signal after digitization, (g) Reconstructed signal from digitized data taken from the 4 channels.

The fiber propagation loss is compensated by the optical EDFA positioned after SMF1. The linearly-chirped pulses are next modulated by an RF signal [Fig. 4.10(c)] using the PolM in the manner discussed earlier. The modulated pulses [Fig. 4.10(d)] then propagate through another length ($L_2 = 28174$ m) of SMF (SMF2) with total dispersion $D_{s,2}L_2 = -468.85$ ps/nm and are stretched for a second time [Fig. 4.10(e)]. This yields a stretch factor of $M = (L_1 + L_2) / L_1 \approx 4$.

This technique does not require any polarization-maintaining (PM) fiber in the optical path as the polarization state of the signals is controlled by PC1 and PC2. The exceptions are the pigtails of the PolM and PBS. As modern non-PM fibers are designed to have low polarization-mode dispersion (PMD), with typical values of PMD parameter as low as $D_p = 0.1$ ps/ $\sqrt{\text{km}}$, the ~ 28 km SMF2 will induce a differential group delay (DGD) of 0.53 ps between the orthogonally-polarized signals [112]. This DGD can be ignored as the chirped optical pulses after time-stretch are approximately 2 ns wide; thus the orthogonally-polarized signals can be assumed to propagate in phase.

An EDFA is positioned after SMF2 also for propagation loss compensation. It is observed at this point that there is temporal overlapping between adjacent pulses. Although indistinguishable in time, the overlapping sections are actually distinct in wavelength. Hence, with the use of a passive element such as a wavelength-division-multiplexer (WDM) filter, the overlapping segments can be bandpass filtered into individual wavelength channels. For this purpose, a 4-channel WDM filter with 10 GHz channel spacing is used in the experiment. According to linear time-to-wavelength mapping, each individual wavelength channel corresponds to different distinct time segments.

Using a PBS, the time-segmented phase modulated signal is then converted into an intensity modulated signal at each arm of the PBS [Fig. 4.10(f1-f4)]. In principle, a four-channel continuous-time TS-ADC is realized when the signals from all four channels are photodetected and digitized at ‘X’ GS/s concurrently [Fig. 4.10(f1-f4)]. The digitized samples taken from each channel can be used to reconstruct the RF signal, which would increase the effective sampling rate to ‘4X’ GS/s [Fig. 4.10(g)]. Due to limited back-end digitizing hardware, we are only able to demonstrate digitization of a single channel at a time for our current TS-ADC scheme.

In the experiment, by controlling the PC2 with respect to the PBS pair as what we have discussed in the previous section, we can polarize the incident light into the complementary signal. Desired differential operation is accomplished when the complementary signal pair is photodetected by a balanced photodetector. Since a 2 GS/s electronic ADC is used in the experiment, a balanced photodetector with no less than 2 GHz bandwidth is required in our experiment. However due to availability, we have used a balanced photodetector with 42 GHz bandwidth. The output differential signal is then digitized by a single-input 10 bit electronic ADC with a sampling rate of 2 GS/s and 3 GHz bandwidth, while the full scale peak-to-peak voltage, V_{FS} is 500mV. The 10 bit electronic ADC has approximately 7.2 ENOB measured at 1 GHz for –1 dB full-scale. The electronic ADC is externally synchronized to the repetition frequency of the regenerative active mode-locked fiber laser, thus permitting the evaluation of the TS-ADC performance over a long time aperture.

To exemplify the suppression of dispersion-induced 2nd-order distortion, we set the modulation depth m_{dep} to a high value of about 0.75 to generate prominent 2nd-order distortion. Fig. 4.11 shows a 950 MHz tone digitized at 2 GS/s along with its

Fig. 4.11 Time domain signal for the 950 MHz tone. The points are the digitized data and the curve is sine fit to the data.

sine-fit curve for the single channel centered at 1550.1 nm. Limited by the gain of EDFAs and RF amplifier, the RF power level of this signal prior to digitization is measured to be -12 dBm by an RF spectrum analyzer. This works out to be 158 mV peak-to-peak ($V_{in,pp}$) or 31.6% of the electronic ADC's peak-to-peak full scale voltage. Based on the digitized output, the FFT spectrum of the 950 MHz tone is calculated and illustrated in Fig. 4.12(a) with the fundamental and odd-order distortion tones clearly visible. These distortion tones are in fact the distortion tones resulting from the harmonics of the fundamental tone that aliased back to the Nyquist bandwidth. We also digitize the 950 MHz tone based on single-ended detection with its corresponding FFT spectrum presented in Fig. 4.12(b). Comparing the FFT spectra, we clearly observe the effective suppression of the 2nd-order distortion tone in the balanced detection scheme, whereas the distortion tone is evident in the single-ended case. Quantitatively, the 2nd-order distortion tone is suppressed by 25 dB, down to a level well below the noise floor and the SFDR is limited by the 3rd-order distortion in the balanced detection scheme.

Fig. 4.12 Calculated FFT spectrum of the digitized 950 MHz signal based on (a) balanced detection scheme and (b) single-ended detection scheme.

Similar to the results in [54], the amount of distortion suppression in our technique is limited by the noise floor of the electronic ADC. Significant amplitude and phase imbalance between the two ports of the PBS as well as possible nonlinear distortions introduced by the RF amplifier before digitization of the signal by the electronic ADC can affect the amount of suppression and these imperfections should be minimized. Additionally, an improvement of ~ 6 dB in SNR is observed for the

differential operation using balanced detection scheme. This improvement is attributed to the coherent addition of the complementary signal and the incoherent addition of noise [113]. Based on the FFT spectrum plot, an SNR of 24.25 dB is obtained over the Nyquist bandwidth of 1 GHz, corresponding to an ENOB of 3.56 bits for the differential operation, while the single-ended operation achieves an ENOB of 2.52 bits. The difference of ~ 1 bit in the ENOB is attributed to the ~ 6 dB difference in the SNR.

Although the ENOB for a single channel is relatively low in our photonic TS-ADC demonstration, we are still able to resolve the RF signal as depicted in Figs. 4.12(a) and 4.12(b). This low ENOB is mainly attributed to the low RF power level of the signal prior to digitization. Due to the gain limitation of the EDFAs and the RF amplifier, the RF power level of the signal is capped at -12 dBm or approximately 31.6% of the peak-to-peak full scale voltage. Since the input signal to the electronic ADC is related to ENOB as, $ENOB = \left[SNR - 1.76 - 20 \log_{10} (V_{in,pp} / V_{FS}) \right] / 6.02$, we can estimate the full scale ENOB for a single channel to be ~ 5.40 ENOB with the assumption that the performance is limited by the electronic ADC's quantization noise [45]. Here, we have equated the SNR to 24.25 dB and $V_{in,pp} / V_{FS}$ to 0.316. However, if optical noise becomes an issue, the estimated full scale ENOB will drop.

In order for the RF power level of the signal to match the full scale voltage of the electronic ADC, improvements to the current system have to be made. As the average optical power of the signal entering the PBS is approximately -4 dBm, placing an additional EDFA before the DWDM provides a trivial solution to boost the RF power level of the signal. Specifically, by optically amplifying the signal to about 1 dBm (5 dB gain), the ensuing RF signal can be boosted to ~ -2 dBm (10 dB RF gain) and

matching the full scale voltage of the electronic ADC. Although trivial, the additional EDFA would introduce further optical noise to the signal and possibly degrade the overall ENOB performance. Another alternative is to replace the SMFs in the current system with DCF which can provide a larger dispersion than SMF over the same length (For instance, DCF module from Avanex with its dispersion equivalent to 20 km of SMF having typical dispersion of 17ps/nm). Shorter lengths of DCF would thus introduce less propagation loss to the optical signal and this results in a higher average optical power before the PBS. Without the use of additional EFDA, the RF power of the signal can be easily matched to the full scale voltage of the electronic ADC, however at the expense of using costly DCFs.

The balanced detection scheme offers simplicity in that the 2nd-order distortion and common-mode noise can be optically cancelled without any post-digital linearization processing. However, its inability to cancel the 3rd-order distortion can still limit the ENOB. This is apparent when operating at high modulation depth values. In order to remove the 3rd-order distortion, one can attempt to incorporate the arcsine operation on the digitized signal, which has been discussed earlier in the optically-sampled ADC section. In this context, this linearization process can be expressed by [54]

$$I_{in}(t_n) = \left[\sin^{-1} \left(\frac{I_{5+,out}(t_n) - I_{5-,out}(t_n)}{I_{5+,out}(t_n) + I_{5-,out}(t_n)} \right) \right] \quad (4.12)$$

where $I_{5\pm,out}(t_n)$ and $I_{in}(t_n)$ are corresponding currents for the quantized output voltage and the recovered input voltage at time instant t_n , respectively. Unlike the optically-sampled ADC case, due to the presence of dispersion-induced 3rd-order distortion as indicated in (4.11), the results in [54] have shown that the arcsine

operation is only able to suppress the 3rd-order distortion to a certain extent. Furthermore, due to the use of balanced photodetection and digitization by a single electronic ADC in our scheme, it should be noted that the digitized data for $I_{5+,out}(t_n) - I_{5-,out}(t_n)$ is collected as a single entity while accurate data for $I_{5+,out}(t_n) + I_{5-,out}(t_n)$ cannot be obtained. Although $I_{5+,out}(t_n) + I_{5-,out}(t_n)$ can be estimated as an averaged constant through the use of a low speed, high resolution oscilloscope, the use of an estimation in (4.12) can further limit suppression of the 3rd-order distortion.

Due to imperfect amplitude and phase balance of its photodiodes, the balanced detector can only reject the common-mode noise to a certain extent. By reducing further the common- mode noise sources which are attributed to the RIN of our regenerative mode-locked fiber laser as well as the amplified spontaneous emission (ASE) noise from the EDFAs, the impact of imperfection in the photodiodes can be reduced substantially and thus improve the ENOB.

Although not yet competitive with current electronic digitizers, with adequate digitizing hardware, our proposed approach has the potential to scale up to a sampling rate of 8 GS/s based on four channels, while the ENOB can also be further enhanced provided the aforementioned imperfections are improved. Based on the previously estimated single channel full scale ENOB limited by quantization noise as well as taking into consideration potential stitching and calibration errors, the ENOB for a four channel 8 GS/s photonic TS-ADC can be estimated to be approximated 4-5 based on the current experimental setup.

4.3 Summary

In this chapter, we have proposed and demonstrated two photonic ADC schemes, the first being an optically-sampled ADC scheme, while the second is a photonic TS-ADC scheme.

In the first section on optically-sampled ADC, we have demonstrated a 2 GS/s optically sampled ADC which utilized complementary dual-output modulation and balanced photodetection scheme. For a single-tone signal at 123 MHz, we have obtained 5.36 SNR-bits, SFDR of 28 dB as well as an ENOB 3.76-bits based on our current hardware. By utilizing a conventional electronic oscilloscope, we were able to measure V_{full} which we incorporated into the arcsine operation that was applied on the digitized data. This linearization technique has shown the ability to suppress the 3rd-order distortions introduced by the nonlinearity of the dual-output MZM. More specifically, we have demonstrated a significant improvement in the SFDR which is now measured to be ~ 43 dB and is limited by the supermode noise of the laser. Furthermore, an improved ENOB of ~ 5.15 -bits has been obtained. Measuring across a range of analog input frequencies and operating the current system in undersampling mode, we measured and obtained an average ENOB of ~ 5 -bits after linearization.

In the second section on photonic TS-ADC, we have shown analytically and experimentally that the complementary output of a polarization modulator can be exploited differentially to suppress the dispersion-induced 2nd-order distortion in a TS-ADC. The differential operation on the complementary output is accomplished by a balanced detection scheme. Using the polarization modulator, we have demonstrated a TS-ADC based on virtual time gating principle towards continuous-time operation. We showed in a single channel that, the 2nd-order distortion can be suppressed using

our proposed differential approach. Compared to single-ended operation, an improvement of 6 dB in SNR is also observed. We digitized an RF signal of 950 MHz at a sampling rate of 2 GS/s and obtained an ENOB of ~ 3.56 bits with a single channel at 31.6% of the electronic ADC's peak-to-peak full-scale voltage. With adequate backend digitizing hardware, a four-channel continuous-time TS-ADC with a sampling rate of 8 GS/s can be realized to handle RF frequencies as high as 4 GHz.

Chapter 5 PHOTONIC GENERATION OF TWIPT FOR OPTICALLY-SAMPLED ADC SYSTEM

5.1 Introduction

In the previous chapter, we have demonstrated a 2 GS/s single-channel optically-sampled ADC system whose sampling rate is fundamentally limited by the 2 GS/s backend electronic ADC. As mentioned in Chapter 2, in order to obtain high-speed sampling rates for an optically-sampled ADC system, multi-interleaving channel schemes in the form of either time-interleaving or wavelength-interleaving can be implemented to circumvent the performance bottleneck imposed by electronic ADCs with low-sampling rates [7, 23, 28, 29, 46].

The time-interleaving approach is based on OTDD parallel processing which involves the use of an electronic switch which is controlled by complex timing signals to demultiplex a high-speed optical pulse-train into parallel channels of lower speed [23]. However, this approach is often found to be limited by the response speed of the switch, which consequently limits the speed of the entire optical sampling system. On the other hand, the wavelength-interleaving approach which is based on WDD parallel processing involves demultiplexing a TWIPT into different wavelength channels by using only passive components such as a 1-to- N wavelength-division-demultiplexer [28]. As the practical implementation of this approach is much simpler than the previous one, much effort has been devoted to investigate schemes to generate a TWIPT [28-37].

Early approaches to obtaining a TWIPT focused on utilizing a broadband pulse-train which is spectrally-sliced into multi-wavelength channels, followed by spectral-

flattening with the aim of amplitude equalization and finally time-interleaving these multi-wavelength pulses using variable time-delay [29, 30]. Conceptually simple, practical implementation based on this approach is, however, difficult. This is especially so, when one is required to fulfill the stringent requirement of precise uniform inter-pulse spacing for a large number of wavelength channels. Since then, various other techniques to generate a TWIPT have been developed and demonstrated with promising results. In particular, it has been shown that through the use of the time-lens compression technique [31], a TWIPT can be generated without the need to provide uniform inter-pulse spacing for the multi-wavelength pulses.

In general, the time-lens compression technique involves phase modulating an array of temporally overlaid pulse-trains whose wavelengths differ by a uniform wavelength spacing, and subsequently propagating them through a length of SMF. The SMF introduces a quadratic phase shift which compensates the existing phase profile of the time-overlaid pulse-trains and as a result, the pulse-trains are temporally compressed and displaced with a uniform spacing in time. In other words, by choosing appropriate parameter values in this technique, one is able to generate a TWIPT with a precisely uniform inter-pulse spacing in time. It should be emphasized that although the time-lens technique offers simple practical implementation, this approach does require an array of CW lasers as well as modulators and microwave sources.

More recently, a much simpler technique to generate a TWIPT, by first dispersing a mode-locked pulse-train and subsequently sectional-compressing the dispersed pulse-train, has been presented [114]. The dispersion-spreading and sectional-compression of the pulse-train are realized by a SMF with standard dispersion parameters and a tunable multi-channel dispersion compensation module,

respectively. The tunable multi-channel dispersion compensation module is based on two superimposed FBGs connected to an optical circulator. The advantage of this approach is that, unlike [31], it does not require any modulators and microwave sources. However, only experimental results have been presented in [114]; also tunability of the pulse width in the generated TWIPT has not been considered. In this chapter, we present a scheme which generates a 10×2 GHz TWIPT with tunable pulsewidth together with a mathematical analysis and supporting simulations.

5.2 Principle of operation, mathematical expressions and simulation results

Fig. 5.1(a) shows the schematic of the proposed scheme. We begin with a broadband pulse-train with a bandwidth of $\Delta\lambda = \lambda_H - \lambda_L$ as well as a repetition period T . This pulse-train is fed to a single-channel dispersion module (SCDM) which has a group-delay characteristic as depicted in Fig. 5.1(b). Given a dispersion of $T / \Delta\lambda$ ps/nm, the pulsewidth of the pulse-train is temporally-broadened to T by the SCDM. Moreover, the pulses at the output of the SCDM are also linearly-chirped across $\Delta\lambda$. The output from the SCDM is propagated into a MCDM which has an overall group-delay characteristic as shown in Fig. 5.1(c). Unlike the SCDM, the MCDM comprises a total of N channels within $\Delta\lambda$, with the channels uniformly-spaced by $\Delta\lambda / N$ and centered at λ_1 to λ_N , respectively. Due to the uniform-spacing between adjacent channels, each channel is group-delayed by the same amount with respect to any of its adjacent channels.

Fig. 5.1 (a) Schematic diagram of the proposed scheme for generating time- and wavelength interleaved pulse-train with tunable pulsewidth, (b) Group delay of the single-channel dispersion module, and (c) group delay of the multi-channel dispersion module.

Since all the channels share a similar group-delay characteristic, setting the dispersion of the MCDM to $-T / \Delta\lambda$ can compensate for the dispersion previously imposed by the SCDM within each channel. As a result, each of the chirped pulses that propagate through the MCDM is segmented and compressed into N pulses which are uniformly-spaced temporally and are centered at wavelengths λ_1 to λ_N . In other words, a TWIPT with a repetition period of T/N and N wavelengths can be generated. To realize the scheme, standard SMF or a LCFBG can be used as the

SCDM, while the MCDM can be realized by a tunable multi-channel dispersion compensation module which is commercially available.

Mathematically, the spectrum for the initial broadband pulse-train with a repetition period of T is expressed as

$$E_1(\omega) = E(\omega - \omega_o) \cdot \sum_{k=-\infty}^{\infty} \delta\left[(\omega - \omega_o) - k \frac{2\pi}{T}\right] \quad (5.1)$$

where $E(\omega - \omega_o)$ represents the shape of the broadband spectrum which has a center angular frequency at ω_o and a bandwidth of $\Delta\lambda$. Assuming that the SCDM is realized by a length L of a SMF with typical 2nd-order dispersion β_2 , the output at the SCDM is given by

$$\begin{aligned} E_2(\omega) &= E_1(\omega) \cdot e^{\frac{j\beta_2 L (\omega - \omega_o)^2}{2}} \\ &= E_1(\omega) \cdot e^{\frac{jGDD_S (\omega - \omega_o)^2}{2}} \end{aligned} \quad (5.2)$$

where $GDD_S = \beta_2 L$ is the group delay dispersion of the SCDM.

The output from the SCDM is next fed to the MCDM. Here, we simplify the mathematics by assuming the channels to have a Gaussian-profile with a half-width bandwidth (at 1/e-intensity point) of $\Delta\omega_{ch}$. Depending on the parity of N , the output of the MCDM is expressed by either (5.3) or (5.4) in the case where there are N uniformly-spaced channels within $\Delta\lambda$:

$$\begin{aligned} &E_3(\omega) \Big|_{N=\text{odd}} \\ &= E_2(\omega) \cdot \sum_{n=-N/2}^{N/2} e^{\frac{-[(\omega - \omega_o) - n\Delta\omega_{sp}]^2}{2\Delta\omega_{ch}^2}} \cdot e^{\frac{jGDD_M [(\omega - \omega_o) - n\Delta\omega_{sp}]^2}{2}} \end{aligned} \quad (5.3)$$

$$\begin{aligned}
 & E_3(\omega) \Big|_{N=\text{even}} \\
 &= E_2(\omega) \cdot \sum_{n=-N/2}^{N/2-1} e^{\frac{-[(\omega-\omega_o)-(n+1/2)\Delta\omega_{sp}]^2}{2\Delta\omega_{ch}^2}} \cdot e^{\frac{jGDD_M[(\omega-\omega_o)-(n+1/2)\Delta\omega_{sp}]^2}{2}} \quad (5.4)
 \end{aligned}$$

where GDD_M is the group delay dispersion and $\Delta\omega_{sp}$ is the angular channel spacing between adjacent channels of the MCDM, respectively. It should be noted that GDD_M is a tunable parameter which can provide positive or negative dispersion and can be tuned within the limits imposed by the MCDM. Equations (5.3) and (5.4) can be approximated and rewritten as (5.5) and (5.6), respectively, where A_n is the amplitude for the n^{th} channel:

$$\begin{aligned}
 & E_3(\omega) \Big|_{N=\text{odd}} \\
 &= \sum_{n=-N/2}^{N/2} E(\omega - \omega_o) \cdot \left[e^{\frac{jGDD_S(\omega-\omega_o)^2}{2}} \cdot e^{\frac{-[(\omega-\omega_o)-n\Delta\omega_{sp}]^2}{2\Delta\omega_{ch}^2}} \right] \cdot \sum_{k=-\infty}^{\infty} \delta\left[(\omega - \omega_o) - k \frac{2\pi}{T}\right] \\
 &\approx \sum_{n=-N/2}^{N/2} A_n \cdot \left[e^{\frac{-[(\omega-\omega_o)-n\Delta\omega_{sp}]^2 \cdot [1/\Delta\omega_{ch}^2 - j(GDD_S + GDD_M)]}{2}} \right] \cdot \sum_{k=-\infty}^{\infty} \delta\left[(\omega - \omega_o) - k \frac{2\pi}{T}\right] \quad (5.5)
 \end{aligned}$$

$$\begin{aligned}
 & E_3(\omega) \Big|_{N=\text{even}} \\
 &= \sum_{n=-N/2}^{N/2-1} E(\omega - \omega_o) \cdot \left[e^{\frac{jGDD_S(\omega-\omega_o)^2}{2}} \cdot e^{\frac{-[(\omega-\omega_o)-(n+1/2)\Delta\omega_{sp}]^2}{2\Delta\omega_{ch}^2}} \right] \cdot \sum_{k=-\infty}^{\infty} \delta\left[(\omega - \omega_o) - k \frac{2\pi}{T}\right] \\
 &\approx \sum_{n=-N/2}^{N/2-1} A_n \cdot \left[e^{\frac{-[(\omega-\omega_o)-(n+1/2)\Delta\omega_{sp}]^2 \cdot [1/\Delta\omega_{ch}^2 - j(GDD_S + GDD_M)]}{2}} \right] \cdot \sum_{k=-\infty}^{\infty} \delta\left[(\omega - \omega_o) - k \frac{2\pi}{T}\right] \quad (5.6)
 \end{aligned}$$

Taking the inverse Fourier-Transform of (5.5), the total optical field at the output of the MCDM can be expressed in the time-domain as (5.7), where $\otimes$ represents convolution. Likewise, the time-domain expression for (5.6) is expressed as (5.8).

$$E_3(t)|_{N=\text{odd}} = F^{-1} \left[E_3(\omega)|_{N=\text{odd}} \right] \\ \propto \sum_{n=-N/2}^{N/2} A_n \cdot \left[e^{\frac{-(t+nGDD_S\Delta\omega_{sp})^2}{2[1/\Delta\omega_{ch}^2 - j(GDD_S + GDD_M)]}} \cdot e^{j(\omega_0 + n\Delta\omega_{sp})t} \cdot e^{\frac{jGDD_S(n\Delta\omega_{sp})^2}{2}} \right] \otimes \sum_{m=-\infty}^{\infty} \delta(t-mT) \quad (5.7)$$

$$E_3(t)|_{N=\text{even}} = F^{-1} \left[E_3(\omega)|_{N=\text{even}} \right] \\ \propto \sum_{n=-N/2}^{N/2-1} A_n \cdot \left[e^{\frac{-(t+(n+1/2)GDD_S\Delta\omega_{sp})^2}{2[1/\Delta\omega_{ch}^2 - j(GDD_S + GDD_M)]}} \cdot e^{j[\omega_0 + (n+1/2)\Delta\omega_{sp}]t} \cdot e^{\frac{jGDD_S[(n+1/2)\Delta\omega_{sp}]^2}{2}} \right] \otimes \sum_{m=-\infty}^{\infty} \delta(t-mT) \quad (5.8)$$

Examining (5.7) and (5.8) closely, we observe that $E_3(t)|_{N=\text{odd(even)}}$ consists of N uniformly-spaced pulses which have different center wavelengths and amplitudes, repeated for every repetition period T . The difference in the center wavelength between adjacent pulses is determined by $\Delta\omega_{sp}$, while the time-delay between adjacent pulses Δt_{delay} can be calculated as

$$\Delta t_{\text{delay}} = GDD_S \Delta\omega_{sp} \quad (5.9)$$

where Δt_{delay} is seen to be directly proportional to both GDD_S and $\Delta\omega_{sp}$. Thus, in order to generate a TWIPT which has a repeated sequence of N uniformly-spaced pulses with N different wavelengths, the following condition has to be satisfied

$$\Delta t_{delay} = \frac{T}{N} = GDD_s \Delta \omega_{sp} \quad (5.10)$$

Assuming both T and $\Delta \omega_{sp}$ are fixed parameters, the condition in (5.10) can be easily satisfied by choosing an appropriate value of GDD_s for N channels within a fixed $\Delta \lambda$. Since GDD_s is the group delay dispersion of the SMF, we can calculate the required L to satisfy (5.10) as follows

$$L = \frac{T}{N \beta_2 \Delta \omega_{sp}} \quad (5.11)$$

Furthermore, by tuning the dispersion of the MCDM in such a way that $GDD_M = -GDD_s$, the pulses can be temporally compressed to their transform-limited pulsewidths which are dependent on $\Delta \omega_{ch}$. Tuning GDD_M to other values would therefore permit the generation of pulses with tunable pulsewidths.

Using (5.6), we perform simulations based on an initial pulse-train with a repetition frequency $f_{rep} = 1/T$ of 2 GHz. The MCDM is used to provide uniform channel spacing of $\Delta f_{sp} = \Delta \omega_{sp} / 2\pi = 100$ GHz. Each of the channels has a Gaussian-shaped profile and a FWHM bandwidth of $f_{FWHM} = 2(\ln 2)^{0.5} \omega_o / 2\pi = 50$ GHz. A 9.3466 km long SMF with standard dispersion parameter $D_\lambda = (2\pi c / \lambda^2) \beta_2 = 16.7$ ps/nm/km is utilized as the SCDM. This specific length of SMF is calculated from (5.11) where we have assumed that there are $N=4$ uniformly-spaced channels within the $\Delta \lambda$ of the initial pulse-train.

In order to generate transform-limited pulses for all the channels, the total dispersion introduced by the SCDM has to be compensated and this is obtainable when the channel dispersion of the MCDM is tuned to

$D_M = -D_S = -D_\lambda L \approx -156.1 \text{ ps/nm}$. The total dispersion of the SCDM and MCDM are related to the corresponding group delay dispersions by the relation $D_{S(M)} = (2\pi c / \lambda^2) GDD_{S(M)}$. It should be emphasized that these parameter values are chosen with the target to generate a $4 \times 2 \text{ GHz}$ TWIPT which has a sequence of $N=4$ uniformly-spaced transform-limited pulses with $N=4$ wavelengths.

Assuming that $A_1 = A_2 = \dots = A_N = A_o$, which is fairly accurate when an initial broadband pulse-train is used, we multiply (5.8) with its complex-conjugate and obtain the total optical power of the TWIPT as illustrated in Fig. 5.2. Fig. 5.2(a) clearly depicts a $4 \times 2 \text{ GHz}$ TWIPT with a sequence of 4 pulses uniformly-spaced at $\Delta t_{\text{delay}} = 125 \text{ ps}$, with the sequence being repeated at 500 ps . The 4 pulses are also seen to have different center wavelengths (λ_1 to λ_4). Fig. 5.2(b) shows a zoomed-in trace of a pulse with wavelength λ_1 . It is observed that the pulse has a Gaussian profile and has a transform-limited pulsewidth of 8.83 ps . Likewise, the pulses with other wavelengths are also noted to have similar profile and transform-limited pulsewidths. It is important to point out that the transform-limited pulsewidth is limited by the value of f_{FWHM} imposed by the MCDM. Pulsewidth tunability is investigated next. Fig. 5.2(c) shows the relationship between the FWHM pulsewidth and the channel dispersion of the MCDM. It is observed that the FWHM pulsewidth is at its minimum when channel dispersion is set at $D_M \approx -156.1 \text{ ps/nm}$, while it broadens for channel dispersion deviating from this value.

Fig. 5.2 Simulation results for a generated 4×2 GHz TWIPT; (a) time-domain trace of the TWIPT, (b) zoom-in trace of a pulse with wavelength at λ_1 , and (c) channel dispersion dependence of the pulsewidth.

Practically, the tunable multi-channel dispersion compensation module has a much wider channel dispersion tuning range than what we have shown in Fig. 5.2(c); therefore a wider pulsewidth-tunability range can be achieved easily. Finally, we repeat the simulations with the target to generate a 10×2 GHz TWIPT. We use the same values for f_{rep} , Δf_{sp} , f_{FWHM} and D_λ , and assume that the $\Delta\lambda$ of the initial pulse-

Fig. 5.3 Simulation results for a generated 10×2 GHz TWIPT; (a) time-domain trace of the TWIPT, and (b) channel dispersion dependence of the pulsewidth.

train can now support $N = 10$ uniformly-spaced channels. Based on (5.11), the required length of the SMF is $L = 3.7386 \text{ km}$. To compensate for the total dispersion introduced by the SCDM, the channel dispersion of the MCDM is now tuned to $D_M = -D_S = -D_\lambda L \approx -62.4 \text{ ps/nm}$ so as to generate transform-limited pulses.

Fig. 5.3(a) illustrates the time-domain trace of the generated 10×2 GHz TWIPT which comprises a sequence of 10 pulses uniformly-spaced at $\Delta t_{\text{delay}} = 50 \text{ ps}$, the sequence being repeated with a period of 500 ps. Each of these 10 pulses is seen to have a different center wavelength (λ_1 to λ_{10}). The relationship between the FWHM pulsewidth and the channel dispersion of MCDM is depicted in Fig. 5.3(b). In this case, a smaller value of $D_M \approx -62.4 \text{ ps/nm}$ is enough to compensate for the dispersion

introduced by the SCDM for the generation of transform-limited pulsewidth of 8.83 ps. By extension, to achieve any given pulsewidth, a smaller channel dispersion of the MCDM is required as compared to the previous case.

Through the simulation results, we have shown that in the context of an experimental setup with fixed f_{rep} , D_λ , Δf_{sp} and f_{FWHM} , the proposed scheme can generate a $N \times f_{rep}$ TWIPT which has a sequence of N uniformly-spaced pulses with N wavelengths. This is made possible only when there are N channels within the $\Delta\lambda$ of the initial pulse-train and one chooses an appropriate value of I based on (5.11). Furthermore, by tuning the channel dispersion of the MCDM to $D_M = -D_S$, one is able to generate transform-limited pulses. Pulsewidth-tunability is also attainable when D_M is tuned to other values.

It is pertinent to mention here that our analysis has so far been based on the assumption that the channel profile of the MCDM is Gaussian-shaped which helps to simplify the mathematics. However, in practice, the channel profile of the multi-channel dispersion compensation module that is used as the MCDM is much closer to a super-Gaussian shape which has steeper trailing and leading edges. More extensive numerical simulations have to be carried out for such profiles. Furthermore, as assumed in our analysis, it is highly desirable in practice that the input pulse-train be broadband so that its bandwidth is able to accommodate a large number of wavelength channels which have approximately the same amplitude. For the pulses in the generated TWIPT to have high extinction ratio, the extinction ratio of the input pulse-train should be at least 15 dB and this is generally obtainable using a mode-locked laser.

Fig. 5.4 Experimental setup for the proposed scheme for generating time- and wavelength interleaved pulse-train with tunable pulsewidth. EDFA: erbium-doped fiber amplifier; HNLF: highly nonlinear fiber; PD: photodetector and SMF: single-mode-fiber.

5.3 Measured results and discussion

Fig. 5.4 shows the experimental setup for the proposed scheme. The setup comprises a SC source (initial pulse-train), a length of SMF which forms the SCDM and a commercially available tunable multi-channel dispersion compensation module to serve as the MCDM. An optical spectrum analyzer and sampling oscilloscope are used to measure the output from the MCDM. First, a 2 GHz SC pulse-train is generated by injecting ultrashort pulses from a home-made 2 GHz regenerative active mode-locked fiber laser into a length of HNLF. The 2 GHz pulse-train generated from the mode-locked fiber laser has a FWHM pulsewidth of 14 ps and a 3-dB bandwidth of 0.32 nm centered at 1550 nm. The pulses are amplified by a high power EDFA before being injected into a 2 km spool of HNLF which has a small normal dispersion value of -1 ps/km/nm and a nonlinearity coefficient of 11.2/(W-km).

Fig. 5.5(a) illustrates the spectrum of the generated 2 GHz SC source which has a ~ 14 nm bandwidth with less than 0.5 dB fluctuation. This bandwidth is generated

when the output power of the high power EDFA is ~ 26 dBm. Further increase in the output power can allow the generation of a SC with wider bandwidth. Compared to a femtosecond pulse-train with a Gaussian-profile spectrum, utilizing such a flat spectrum in the setup can ensure that the amplitudes for different channels at the output of the tunable multi-channel dispersion compensation are approximately the same, without the use of any amplitude compensation scheme. The inset in Fig. 5.5(a) shows the corresponding pulsewidth of ~ 0.3 ps, which is close to being transform-limited. The corresponding oscilloscope trace of the 2 GHz SC source is shown in Fig. 5.5(b) wherein the inter-pulse spacing is 500 ps.

To generate a 10×2 GHz TWIPT, we then carve a 8 nm-wide spectrum from that of the SC source using an optical bandpass filter as we are using a tunable multi-channel dispersion compensation module which has a uniform channel spacing of $\Delta f_{sp} = 100$ GHz (or 0.8 nm). This amount of bandwidth is occupied by $N = 10$ uniformly-spaced channels as depicted in Fig. 5.5(c). The first channel is centered at 1546.12 nm and all channels have a profile close to a 6th-order super-Gaussian shape which has a 1-dB bandwidth of 0.52 nm. Based on our experimental parameters, from (5.11), a ~ 3.73 km of SMF with a dispersion of $D_s \approx 62.4$ ps/nm is required to provide uniform interpulse spacing to the 10 channels.

Fig. 5.5 Measured results for (a) optical spectrum & transform-limited pulsewidth for initial 2 GHz SC source; (b) corresponding oscilloscope trace of the 2 GHz SC source; (c) optical spectrum at the output of the tunable multi-channel dispersion compensation module; (d) oscilloscope trace of the generated 10x2 GHz TWIPT.

Fig. 5.6 Measured results for oscilloscope of the demultiplexed 10x2 GHz TWIPT.

We then set $D_{MCDM} = -63$ ps/nm (limited by the 1 ps/nm resolution of the tunable multi-channel dispersion compensation module) so as to generate nearly transform-limited pulses for the TWIPT. The average power of the 10x2 GHz TWIPT before photodetection is noted to be ~ 6 dBm. The oscilloscope trace of the generated 10x2 GHz TWIPT is illustrated in Fig. 5.5(d). Clearly, there are 10 uniformly-spaced pulses within 500 ps. The amplitude variation in the pulses can be attributed to the ripple in the spectrum of the initial 2 GHz SC source as well as the unequal magnitude response of the tunable multi-channel dispersion compensation module. The maximum amplitude difference is noted to be less than $\sim 9\%$.

To verify our scheme further, the output of the 10x2 GHz TWIPT is fed to a passive wavelength-division-demultiplexer for demultiplexing and Fig. 5.6 illustrates 10 individual 2 GHz pulse-trains which are of different wavelengths and time offsets. The maximum amplitude difference in the pulses for each wavelength channel is less

Fig. 5.7 Experimental (dots and error bars) and simulation results (dashed line) for channel dispersion dependence of the pulsewidth.

than 0.3%. Using an autocorrelator, the pulsewidths for all channels are measured to be ~ 10.2 ps and the pulses for each channel exhibit a high extinction ratio of ~ 20 dB. It should be noted that, the measured extinction ratio is currently limited by the autocorrelator which we have used for the measurement.

Next, pulsewidth tunability is verified by setting different values of D_M . Fig. 5.7 depicts the measured FWHM pulsewidth for five different values of D_M (0, -30 , -63 , -90 and -120 ps/nm). Clearly, we observe that when D_{MCDM} deviates from -63 ps/nm, the corresponding FWHM pulsewidth broadens. We have also numerically simulated the channel dispersion dependence of the pulsewidth, assuming that the channels have a 6th-order super-Gaussian shape and a 1-dB bandwidth of 0.5 nm. Evidently, the measured values are in good agreement with the simulated ones. The FWHM pulsewidth is nearly transform-limited when $D_M = -63$ ps/nm. Although D_M can be tuned from -900 ps/nm to 900 ps/nm, the useful tuning range of the generated 10×2 GHz TWIPT should be from ~ 10.2 ps to 50 ps so as to avoid overlap between adjacent pulses. As mentioned earlier, utilization of a tunable multi-channel

Fig. 5.8 Simulation result for time delay between adjacent pulses due to (blue) sole effect of β_2 and (red) combined effects of β_2 and β_3 .

dispersion compensation module which has a wider channel bandwidth can facilitate the generation of transform-limited pulses shorter than the current 10.2 ps.

So far, our earlier analysis has been based on the assumption that the dispersion of the SCDM (realized by a length of SMF) is governed only by β_2 . However, the effects of higher-order dispersion terms of the SMF, in particular the third-order dispersion (TOD) term β_3 can affect the performance of the TWIPT and is therefore investigated here. Since the higher-order dispersive effects are governed by the dispersion slope S , the TOD can be calculated based on the following expression [115]

$$\beta_3 = \frac{S - (4\pi c / \lambda^3) \beta_2}{(2\pi c / \lambda^2)^2} \quad (5.12)$$

As $\beta_2 = 0$ at $\lambda = \lambda_{ZD}$ (zero dispersion wavelength), using values of $\beta_2 \approx -21.2852 \text{ ps}^2/\text{km}$, $S = 0.09 \text{ ps}/(\text{nm}^2 \cdot \text{km})$ and $\lambda_{ZD} = 1300 \text{ nm}$ for the SMF which we have used (Corning SMF-28), the TOD is calculated as $\beta_3 \approx 0.0723 \text{ ps}^3/\text{km}$. Including TOD, the group delay τ_g introduced by the SMF is thus given as

$$\begin{aligned} \tau_g &= \frac{\delta(\beta_2 L \omega^2 / 2 + \beta_3 L \omega^3 / 6)}{\delta\omega} \\ &= \beta_2 L \omega + \beta_3 L \omega^2 / 2 \end{aligned} \quad (5.13)$$

Subsequently, the time-delay between adjacent pulses $\Delta t_{\text{delay}, \beta_2 \& \beta_3}$ due to the combined effects of β_2 and β_3 can be calculated using the following expression

$$\begin{aligned} \Delta t_{\text{delay}, \beta_2 \& \beta_3} &= \tau_{g, i+1} - \tau_{g, i} \\ &= \beta_2 L \Delta \omega_{sp} + 0.5 \beta_3 L (\omega_{i+1}^2 - \omega_i^2) \\ &= \Delta t_{\text{delay}} + 0.5 \beta_3 L (\omega_{i+1}^2 - \omega_i^2) \end{aligned} \quad (5.14)$$

where ω_i is the center angular frequency for the i^{th} channel of the MDCM and Δt_{delay} has been previously derived in (5.9). Fig. 5.8 depicts a simulation result which compares $\Delta t_{\text{delay}, \beta_2 \& \beta_3}$ and Δt_{delay} based on the experimental parameters used as well as the calculated β_3 . Clearly, the discrepancy between $\Delta t_{\text{delay}, \beta_2 \& \beta_3}$ and Δt_{delay} can be attributed to the effect of β_3 and it is also noted that the maximum discrepancy is kept within 0.5 ps. Such a discrepancy would cause a slightly non-uniform spacing (<1%) between the adjacent pulses of the generated TWIPT. One possible solution to cancel the β_3 effect arising from the SMF is through the use of a LCOS (liquid-crystal-on-silicon) based SLM which allows the introduction of a complementary β_3 effect and even higher-order dispersion effects on the initial pulse-train. This would render β_2

as the sole dispersion effect and thus ensure the generation of a highly uniformly-spaced TWIPT.

Additionally, we have also investigated how the performance of our system can be affected by an initial frequency chirp as well as chirp fluctuation in the broadband pulse source used. As it is not trivial to describe the evolution of a chirped broadband pulse through a length of SMF mathematically, therefore for simplicity, the aforementioned effects will be investigated based on the assumption of a chirped Gaussian pulse. Ignoring the uncorrelated effect of β_3 , the spectrum of a chirped Gaussian pulse after propagating through a length of SMF can be expressed as [112]

$$\begin{aligned}
 E(\omega) &= \left(\frac{2\pi\tau_o^2}{1+j(C+\delta C)} \right) \cdot e^{-\frac{\omega^2}{2(1+j(C+\delta C))/\tau_o^2}} \cdot e^{\frac{j\beta_2 L \omega^2}{2}} \\
 &= \left(\frac{2\pi\tau_o^2}{1+j(C+\delta C)} \right) \cdot e^{-\frac{\omega^2}{2(1+(C+\delta C)^2)/\tau_o^2}} \cdot e^{j \left(\frac{\beta_2 L \omega^2}{2} + \frac{(C+\delta C)\tau_o^2 \omega^2}{2(1+(C+\delta C)^2)} \right)}
 \end{aligned} \tag{5.15}$$

where τ_o is the half-width at $1/e$ of the maximum of the pulse which is related to the pulsewidth at full-width half maximum τ_{FWHM} as $\tau_{FWHM} = 2\sqrt{\ln 2}\tau_o$, C is the chirp parameter defined as $C = (\Delta\omega^2\tau_o^2)^{0.5}$ and δC is the chirp fluctuation.

According to (5.15), the group delay introduced to the pulse by its initial chirp and SMF is given as

$$\begin{aligned}
 \tau_g &= \frac{\delta \left(\beta_2 L \omega^2 / 2 + (C + \delta C) \tau_o^2 \omega^2 / 2 (1 + (C + \delta C)^2) \right)}{\delta \omega} \\
 &= \beta_2 L \omega + \frac{(C + \delta C) \tau_o^2 \omega}{1 + (C + \delta C)^2}
 \end{aligned} \tag{5.16}$$

As evident in (5.16), an initial frequency chirp can introduce a similar amount of group delay to the pulse using a shortened length of SMF as compared to the unchirped case. However, any chirp fluctuation can also prove to be detrimental as the group delay and more importantly, the time-delay between adjacent pulses fluctuates around the desired value. Additionally, chirp fluctuation also affects the amplitude as well as the bandwidth of the spectrum as indicated by the first two terms in (5.15), respectively. Thus, fluctuation in the amplitude and pulsewidth of the generated TWIPT's pulses becomes inevitable in the presence of any chirp fluctuation.

The generated 10×2 TWIPT based on the proposed scheme can be applied in a high-speed optically-sampled ADC system based on WDD parallel processing, as shown earlier in Fig. 2.3(b). For example, after sampling an input RF frequency, the modulated 10×2 GHz TWIPT can be demultiplexed into 10 channels at a lower sampling rate (2 GHz) for WDD parallel processing and captured by 10 individual 2 GS/s electronic ADCs. Effectively, a 20 GS/s photonic ADC can be achieved based on the incorporation of the generated TWIPT in a WDD parallel processing setup. Since pulsewidth and peak power are related, the pulsewidth tunability offered by our proposed scheme can be used to lower the peak power of the TWIPT so as to avoid generating additional nonlinear distortions that may arise from the nonlinear response of the PD during photodetection. Hence, the pulsewidth tunability can prevent the SINAD of the optical sampling ADC system from further degradation. However, due to inadequate amount of hardware (electronic ADCs and low noise amplifiers) at the point of experimentation, it has not been possible to demonstrate such a photonic ADC based on our generated TWIPT.

Since the bandwidth of the SC can be widened by increasing the power of the EDFA, this scheme can be easily scaled up to many more channels. For instance, to

generate a 20×2 TWIPT, the SC source is required to be broadened to > 16 nm and the optical bandpass filter bandwidth to be also tuned to 16 nm. The required length of SMF would be 1.8693 km, corresponding to $D_S \approx 31.2$ ps/nm. Nearly transformed-limited pulses can thus be generated when $D_M = -31$ ps/nm. Finally, the practical implementation of the proposed scheme can be further simplified if the SC source is replaced by a broadband source such as a femtosecond mode-locked fiber laser and by replacing the optical bandpass filter with a LCOS based SLM which can ensure channel equalization [116]. The SMF can also be replaced with a more compact component such as a LCFBG.

5.4 Summary

In this chapter, we have demonstrated analytically and experimentally a technique which allows to generate a TWIPT by dispersing a broadband pulse using SMF followed by sectional-compression of the dispersed pulse using a tunable multi-channel dispersion compensation module. For an initial broadband pulse-train with a fixed f_{rep} as well as a tunable multi-channel dispersion compensation module with a fixed channel spacing Δf_{sp} , the analysis and simulation results presented here have shown that for N channels within the bandwidth of the initial pulse-train, choosing appropriate values for the length of SMF as well as the channel dispersion D_M can allow us to generate a $N \times f_{rep}$ TWIPT and also achieve pulsewidth tunability.

We have experimentally demonstrated the generation of a 10×2 GHz TWIPT based on an initial 2 GHz SC source. Nearly transform-limited pulsewidths of 10.2 ps are measured for all channels when we set $D_M \approx -D_S$. Pulsewidth tunability has also

been experimentally verified by tuning the value of D_M . The advantages of the proposed scheme are its simplicity in practical implementation and robustness. Furthermore, no microwave components are required. A potential application of the generated 10×2 TWIPT is for high-speed optically-sampled ADC systems based on WDD parallel processing.

Chapter 6   PHOTONIC   GENERATION   OF

FREQUENCY-TUNABLE CW MICROWAVE SIGNALS

6.1   Introduction

Photonic generation of microwave signals with attributes such as low-phase noise, high frequency, frequency-tunability and long term stability, has been an active area of research over the last decade since such microwave signals are of great importance in modern systems and applications which include radio-over-fiber networks, broadband wireless access, radar, satellite communication, navigation and spectroscopy [117-120].

Early photonic techniques to generate CW microwave signals with the aforementioned attributes include optical heterodyning [59-63], external modulation of a laser [64-70], and heterodyning lasing wavelengths from a dual-wavelength laser [71-76]. The principles of operation, governing mathematics, recent developments, and limitations of each of these techniques have been described in Chapter 3. In particular, in the first two techniques, an external microwave reference source of high quality is essential. In the third technique, the FBGs which are incorporated in the cavity of the dual-wavelength laser can be complicated to fabricate or expensive to purchase.

A recent technique for generation of CW microwave signals is the direct photodetection of the output from a mode-locked laser [95-98]. By beating the phase-locked longitudinal modes of a mode-locked laser, the phase fluctuations between them can be cancelled, thus enabling the generation of microwave signals with low-

phase noise. Some of the schemes in this technique require microwave filters and switches, however, such an approach does not offer flexibility in the system design and inevitably adds to the cost of the whole system.

In this chapter, we discuss two photonic techniques based on mode-locked lasers that allow us to overcome the need to use microwave bandpass filters and switches, but can generate frequency-tunable CW microwave signals. In the following section, we discuss the first technique which utilizes an array of uniformly-spaced optical combs generated with the help of a mode-locked laser.

6.2 Photonic Generation of Frequency-tunable CW Microwave Signals Using an Array of Uniformly-Spaced Optical Combs

It has been shown that by passing an optical comb through two narrowband optical filters, two comb lines can be selected for beating in a PD such that a microwave signal can be generated [117, 121, 122]. In such a scheme, the wavelength separation between the two selected comb lines determines the frequency of the generated microwave signal, while frequency-tunability can be achieved by tuning the center wavelength of the narrowband filters to appropriate comb lines. Specifically, the frequency of the generated microwave signal can be tuned to any integer multiple of the comb spacing (frequency spacing between the adjacent comb lines).

It should be emphasized that the range of frequency-tunability is only limited by the bandwidth of the PD as well as the operating wavelength range of the narrowband filters. Such narrowband filters can be realized by commercial arrayed waveguide gratings which typically have a passband bandwidth of 10 to 20 GHz. To ensure that

only one comb line is selected by the filter, the optical comb usually has a comb spacing which is larger than the passband bandwidth of the filter. In a typical optical comb generation set-up, which is based on injection of ultrashort pulses from mode-locked laser into a nonlinear medium, to generate an optical comb with 10 to 20 GHz comb spacing, an active mode-locked laser is used. This means that an external microwave reference source operating at a high frequency is required in a typical optical comb generation setup.

On the other hand, Qasymeh *et al.* [120] have shown recently that by time-multiplexing a series of optical combs, frequency-tunable microwave signals can be generated without the use of any optical or microwave filters. The principle of this approach has also been adopted and demonstrated in an arbitrary waveform generator [123]. For the experimental results reported in [120], the optical combs are generated by modulating a number of tunable CW laser sources with different optical carriers using a cascade of an intensity modulator and a phase modulator. However, this approach of generating optical combs is inconvenient when a large number of optical combs are required. Furthermore, it is difficult to control the number of comb lines in each optical comb as well as the amplitude of each comb line.

Besides the comb generation approach proposed in [120], optical combs can also be generated by injecting ultra-short optical pulses into a highly nonlinear medium for SC generation, followed by optical filtering. Based on this new approach, we show in this section, analytically and experimentally, generation of frequency-tunable CW microwave signals. Unlike [120], we generate an array of uniformly-spaced optical combs with identical comb profile based on spectral-filtering a SC source using a Fourier-domain programmable optical processor (FD-POP). The SC source used here

is generated by injecting ultrashort optical pulses from a regenerative active mode-locked fiber laser into a length of HNLF.

It should be emphasized that our approach does not require any external microwave reference source and modulators. In the case of a regenerative active mode-locked fiber laser, the mismatch between the instantaneous driving frequency of the modulator and the cavity length is known to be reduced due to the regenerative feedback loop. Therefore, the dominating contributions to the phase noise of its generated optical comb are intrinsic and arise from the cavity detuning and quantum noise [24].

On the other hand, for architectures that utilize a microwave source to generate optical combs, the most important contributions to the phase noise come from the microwave source; thus the phase noise of its comb lines is generally much higher than what can be achieved for a regenerative active mode-locked fiber laser. Additionally, unlike some of the aforementioned schemes in which narrowband optical filters are used to select two particular comb lines for beating in a PD, our approach does not require the comb spacing to be larger than the minimum passband bandwidth of the optical filter. By using an optical comb with smaller comb spacing, the frequency-tunability can also have a smaller frequency step. The principle of operation, governing mathematics and simulation results are presented next and following that, we present measured results supporting the theoretical results. Using a 2 GHz SC source as the initial optical comb, we demonstrate the photonic generation of microwave signals tunable to 8, 10, and 12 GHz.

Fig. 6.1 (a) Schematic diagram of the proposed scheme for generating frequency-tunable microwave signals. PD: photodetector; SMF: single-mode-fiber; (b) spectral-filtering of the initial optical comb into an array of uniformly-spaced optical combs; f_{rep} : repetition rate of optical comb; Δf : frequency spacing between adjacent optical carriers; color lines are optical carriers.

6.2.1 Principle of operation, mathematical expressions and simulation results

Fig. 6.1(a) shows the schematic of the proposed scheme. An initial optical comb with a wide spectrum is first passed through a FD-POP to spectrally-slice the initial optical comb into an array of uniformly-spaced optical combs, each with an identical comb profile. Fig. 6.1(b) depicts a simple illustration of this process. Subsequently, the array of optical combs is propagated through a length of standard SMF. The dispersion in the SMF causes a wavelength-dependent time-delay for all comb lines. Finally, the array of time-delayed optical combs impinges on a PD for O/E conversion. The electrical output is monitored using an electronic spectrum analyzer.

For an initial optical comb which is generated by injection of ultrashort mode-locked pulses into an HNLF, the total optical field can be expressed as [112]

$$E_1(t) = \sum_{n=-M}^M A_n e^{j(\omega_c + n\omega_{rep})t} \cdot e^{j\phi_n} \quad (6.1)$$

where A_n , ϕ_n and $\omega_c + n\omega_{rep}$ are the amplitude, phase and angular frequency of a specific mode among $2M+1$ comb lines, ω_c is the angular frequency of the optical carrier and ω_{rep} is the angular frequency spacing between the adjacent comb lines or the angular repetition frequency of the mode-locked laser. Due to the nature of mode-locking, the phase difference between adjacent comb lines is locked to a constant value of ϕ , which implies that $\phi_n = \phi_o + n\phi$ where ϕ_o is the initial phase.

The initial optical comb $E_1(t)$ is passed through the FD-POP which leads to an array of optical combs, each with a different optical carrier frequency. Assuming that the total number of optical combs is P and that each optical comb has $2N+1$ comb lines of the same amplitude, the total optical field $E_2(t)$ at the output of the FD-POP is given as

$$\begin{aligned} E_2(t) = & A_1 \sum_{n=-N}^N e^{j(\omega_1 + n\omega_{rep})t} \cdot e^{j\phi_{n,1}} \\ & + A_2 \sum_{n=-N}^N e^{j(\omega_2 + n\omega_{rep})t} \cdot e^{j\phi_{n,2}} \\ & + \dots \\ & + A_P \sum_{n=-N}^N e^{j(\omega_P + n\omega_{rep})t} \cdot e^{j\phi_{n,P}} \end{aligned} \quad (6.2)$$

where A_p and ω_p are the amplitude and angular frequency of the p^{th} optical carrier, respectively. The phase of a particular comb line belonging to an optical comb having

the p^{th} optical carrier is given as $\phi_{n,p} = \phi_p + n\phi$, where ϕ_p is its initial phase.

This array of optical combs with different optical carrier frequencies is chromatically dispersed by a length L of SMF which causes a wavelength-dependent time-delay to all comb lines. Assuming only 2nd-order dispersion β_2 , the output of the SMF can be expressed as

$$\begin{aligned}
 E_3(t) = & A_1 \cdot e^{j\omega_1 t} \sum_{n=-N}^N e^{jn\omega_{rep}t} \cdot e^{j\frac{1}{2}\beta_2 L(\omega_1 + n\omega_{rep})^2} \cdot e^{j\phi_{n,1}} \\
 & + A_2 \cdot e^{j\omega_2 t} \sum_{n=-N}^N e^{jn\omega_{rep}t} \cdot e^{j\frac{1}{2}\beta_2 L(\omega_2 + n\omega_{rep})^2} \cdot e^{j\phi_{n,2}} \\
 & + \dots \\
 & + A_p \cdot e^{j\omega_p t} \sum_{n=-N}^N e^{jn\omega_{rep}t} \cdot e^{j\frac{1}{2}\beta_2 L(\omega_p + n\omega_{rep})^2} \cdot e^{j\phi_{n,p}}
 \end{aligned} \tag{6.3}$$

As the PD is a square-law device, the resultant photocurrent is expressed as

$$I(t) = R |E_3(t)|^2 \tag{6.4}$$

where R is the responsivity of the PD.

As shown in [120], beating between comb lines from optical combs of different optical carriers will not be detected by the PD when the following condition is satisfied

$$(\omega_x - N\omega_{rep}) - (\omega_{x-1} + N\omega_{rep}) > B \tag{6.5}$$

where $x \in \{2, 3, \dots, P\}$ and B is the bandwidth of the PD.

Thus, the photocurrent in (6.4) can be described by the following expression

$$\begin{aligned}
 I(t) = & R |A_1|^2 \sum_{n=-N}^N e^{jn\omega_{rep}t} \cdot e^{j\frac{1}{2}\beta_2 L (\omega_1 + n\omega_{rep})^2} \cdot e^{j\phi_{n,1}} \\
 & \cdot \sum_{n=-N}^N e^{-jn\omega_{rep}t} \cdot e^{-j\frac{1}{2}\beta_2 L (\omega_1 + n\omega_{rep})^2} \cdot e^{-j\phi_{n,1}} \\
 & + R |A_2|^2 \sum_{n=-N}^N e^{jn\omega_{rep}t} \cdot e^{j\frac{1}{2}\beta_2 L (\omega_2 + n\omega_{rep})^2} \cdot e^{j\phi_{n,2}} \\
 & \cdot \sum_{n=-N}^N e^{-jn\omega_{rep}t} \cdot e^{-j\frac{1}{2}\beta_2 L (\omega_2 + n\omega_{rep})^2} \cdot e^{-j\phi_{n,2}} \\
 & + \dots \\
 & + R |A_p|^2 \sum_{n=-N}^N e^{jn\omega_{rep}t} \cdot e^{j\frac{1}{2}\beta_2 L (\omega_p + n\omega_{rep})^2} \cdot e^{j\phi_{n,p}} \\
 & \cdot \sum_{n=-N}^N e^{-jn\omega_{rep}t} \cdot e^{-j\frac{1}{2}\beta_2 L (\omega_p + n\omega_{rep})^2} \cdot e^{-j\phi_{n,p}}
 \end{aligned} \tag{6.6}$$

As indicated by (6.6), the photocurrent comprises a number of series of microwave components with different amplitudes, frequencies and phases. Aggregating the microwave components which are of the same frequency, either constructive or destructive interference between them can take place depending on their relative phase shifts. For a fixed length of SMF, the relative phase shift between any two microwave components at the same frequency can be easily adjusted by tuning the frequency spacing of the associated optical carriers using the FD-POP.

For a clearer insight, assuming $A_1, A_2, A_3, \dots, A_p = A_0$ and denoting the angular frequency separation between the adjacent optical carriers as $\Delta\omega_{sp} = \omega_p - \omega_{p-1}$ and ω as the angular frequency of the first optical carrier, after some mathematical processing (6.6) can be rewritten as

$$I(t) = R|A_0|^2 \left\{ (2N+1) + \sum_{n=1}^{2N} \frac{2\sin[nP\phi_A]}{\sin[n\phi_A]} \cdot \frac{\sin[\{2N-(n-1)\}n\phi_B]}{\sin[n\phi_B]} \cdot \cos\left[n(\omega_{rep}t + \phi + \beta_2 L \omega_{rep}) + n(P-1)\phi_A\right] \right\} \quad (6.7)$$

In (6.7),

$$\phi_A = \frac{\beta_2 L \omega_{rep} \Delta\omega_{sp}}{2} \quad (6.8)$$

$$\phi_B = \frac{\beta_2 L \omega_{rep}^2}{2} \quad (6.9)$$

Further, in (6.7), the first term within the parentheses is a DC term, while the second term represents a summation of all the microwave components in the photocurrent. There are a total of $2N$ microwave components. The amplitude of the n^{th} microwave component V_n is given as

$$V_n(\omega_{rep}) \propto \frac{2\sin[nP\phi_A]}{\sin[n\phi_A]} \cdot \frac{\sin[\{2N-(n-1)\}n\phi_B]}{\sin[n\phi_B]} \quad (6.10)$$

Here, it is observed that the resultant amplitude of the n^{th} microwave component is largely dependent on the values of the parameters, ω_{rep} , β_2 , L , P , N and $\Delta\omega_{sp}$. Thus, using an initial optical comb with a fixed ω_{rep} as well as a fixed length L of SMF with typical values for β_2 , the amplitude of the microwave components can be varied simply by choosing appropriate values for P , N and $\Delta\omega_{sp}$.

Examining (6.10) closer, we can generate a dominant microwave signal at the J^{th} harmonic ($n = J$) with all the other harmonics fully suppressed provided the first term

on the right hand side of (6.10) is a maximum when $n = J$ and a null when $n \neq J$.

Specifically, the following criteria should be satisfied

$$\begin{aligned} \sin[nP\phi_A] &= 0 & \text{for } 1 \leq n \leq 2N \\ \sin[n\phi_A] &= 0 & \text{for } n = J, \quad 1 \leq n \leq 2N \\ \sin[n\phi_A] &\neq 0 & \text{for } n \neq J, \quad 1 \leq n \leq 2N \end{aligned} \quad (6.11)$$

This can be achieved when ϕ_A is properly designed such that $\phi_A = k\pi/J$, where k is any integer. Assuming $k = 1$, P would therefore have to be chosen as $P = J$ so as to ensure that the numerator $\sin[nP\phi_A]$ will always be a null for all n . Since ϕ_A is generally much larger than ϕ_B and N is typically of the order of tens, the effect of the second term on the right hand side in (6.10) is small and therefore can be ignored. Combining the above value of ϕ_A and (6.8), to generate a microwave signal at the J^{th} harmonic of the repetition frequency, the value of $\Delta\omega_{sp}$ can be calculated as

$$\Delta\omega_{sp} = \frac{2k\pi}{\beta_2 L \omega_{rep} J} = \frac{2\pi}{\beta_2 L \omega_{rep} J} \quad (6.12)$$

Here, we have used $k = 1$ so as to minimize the required value of $\Delta\omega_{sp}$. Clearly, we observe that $\Delta\omega_{sp}$ is inversely proportional to J .

Using (6.7), we first perform simulations using an initial optical comb with a repetition rate f_{rep} of 2 GHz. A 2 km long SMF with $\beta_2 = 20 \text{ ps}^2/\text{km}$ is used. We assume a total of $P=5$ optical combs, with each of these optical combs having 13 ($N=6$) comb lines of the same amplitude. The first optical comb has its optical carrier frequency ($\omega/2\pi$) set at $f_1 = 192.65 \text{ THz}$, assuming one of the comb line lies on this frequency. Based on the condition $P=J$, $P=5$ optical combs will enable us to

generate dominant microwave signals with frequencies at integer multiple of the 5th harmonic ($J=5$). Using (6.12), we can then calculate the required frequency spacing between the adjacent optical carriers ($\Delta f_{sp} = \Delta \omega_{sp} / 2\pi$) which will allow us to fully suppress all the other harmonics. This gives $\Delta f_{sp} = 397.8874$ GHz. However, since the optical combs comprise only discrete frequencies, we have to approximate Δf_{sp} to the nearest integer multiple of f_{rep} , which in this case is $\Delta f_{approx} = 398$ GHz. This approximation will limit the amount of suppression of the unwanted harmonics.

Taking the magnitude squared of the Fourier transform of (6.7), we obtain the corresponding RF spectrum of the photocurrent, which is illustrated in Fig. 6.2(a). A total of 12 microwave frequency components are shown to be generated in the RF spectrum. However, it is evident that the power for both the 10 GHz (5th harmonic) and 20 GHz (10th harmonic) signals are significantly larger than the other harmonics. With the exception of the 10th harmonic, it can be seen that there is greater than 50 dB suppression relative to the 5th harmonic for all the other harmonics. The suppression ratio can be better but for the approximation which we have made for Δf_{sp} .

We next examine the amplitude of all the 12 harmonics based on (6.10). As shown in Figs. 6.2(b-d), with the exception of the 5th and 10th harmonics exhibiting a maximum, notches are evident for all the other harmonics when the repetition frequency is 2 GHz. This result clearly supports the RF spectrum presented in Fig. 6.2(a). In addition, it is noted from Figs. 6.2(b-d) that when the repetition frequency is either 4 or 6 GHz, a similar result can be obtained.

Fig. 6.2 Simulation results for a generated microwave signal at 10 GHz. (a) RF spectrum for the 10 GHz signal; (b) amplitude for the 1st to 4th harmonic, (c) amplitude for the 5th to 8th harmonic, and (d) amplitude for the 9th to 12th harmonic.

For the generation of a single-tone microwave signal, we keep all the other parameters the same and reduce the number of comb lines in each optical comb to 9 ($N = 4$) and repeat the simulations. Figs. 6.3(a) and 6.3(b)-(c) depict the RF spectrum of the photocurrent as well as the amplitude of all the harmonics, respectively. It is observed clearly that by reducing the number of comb lines, a dominant single-tone microwave signal at 10 GHz can be generated. Again, there is greater than 50 dB suppression relative to the 5th harmonic for all the unwanted harmonics.

Since $2N$ microwave components are generated when each optical comb has $2N+1$ comb lines, to generate a single-tone dominant microwave signal at the J^{th} harmonic, the value of $2N$ must be equal to or greater than the value of J , but at the same time smaller than $2J$ so as to avoid generating the $2J^{\text{th}}$ harmonic. Thus we can generalize that in order to generate a dominant single-tone microwave signal at the J^{th} harmonic, the number of comb lines in each optical comb should fulfill the following criteria

$$J \leq 2N < 2J \quad (6.12)$$

Furthermore, the amplitude of the desired microwave signal increases in strength with N as long as (6.13) is satisfied. Since the number of beating signals, whose frequencies are the same as the desired microwave signal, increases when N increases, summation of these beating signals results in the desired microwave signal having larger amplitude. However, the tradeoff would be an increase in the number of unwanted harmonics.

Fig. 6.3 Simulation results for a generated microwave signal at 10 GHz. (a) RF spectrum for the 10 GHz signal; (b) amplitude for the 1st to 4th harmonic, and (c) amplitude for the 5th to 8th harmonic.

Frequency-tunability of our scheme is verified next. Again, we use an initial 2 GHz optical comb and a 2 km length of SMF, $P=7$ optical combs with each comb having a total of 13 ($N=6$) comb lines of equal amplitude. Based on (6.12), Δf_{sp} is calculated to be $\Delta f_{sp} = 284.2053$ GHz and is approximated to $\Delta f_{approx} = 284$ GHz. As shown in Fig. 6.4(a), we can obtain a dominant single-tone microwave signal at 14

Fig. 6.4 Simulation results for a generated microwave signal at (a) 14 GHz and (b) 28GHz.

GHz (7th harmonic), with greater than 42 dB suppression for the remaining harmonics. Also, as shown in Fig. 6.4(b), a microwave signal at 28 GHz (14th harmonic) is attainable when we use $P=14$ optical combs, with each comb having a total of 19 ($N=9$) comb lines of equal amplitude; Δf_{sp} is calculated to be $\Delta f_{sp} = 142.1026$ GHz and is approximated to $\Delta f_{approx} = 142$ GHz. A greater than 40 dB suppression relative to the 14th harmonic for the other harmonics is observed.

It should be emphasized that the difference in the suppression ratio seen in Figs. 6.3 and 6.4 is due to the difference in $|\Delta f_{sp} - \Delta f_{approx}|$. A smaller difference in $|\Delta f_{sp} - \Delta f_{approx}|$ as seen in Fig. 6.3 results in a greater suppression of all the unwanted harmonics. One potential method to minimize $|\Delta f_{sp} - \Delta f_{approx}|$ is to utilize a tunable

dispersion-compensation module to fine tune the value of $\beta_2 L$ in (6.12), such that the calculated Δf_{sp} is exactly at an integer multiple of f_{rep} . However, in the proof-of-concept experiments here, this method is not implemented.

Through the simulation results, we have shown that in the context of an experimental setup with fixed f_{rep} , β_2 and L , the proposed scheme has the potential to generate a single-tone microwave signal of high purity at the J^{th} harmonic of the repetition frequency by first setting $P = J$ and N satisfying (6.13). Using (6.12) we can then calculate the value for Δf_{sp} and approximate it to the nearest integer multiple of f_{rep} such that all the unwanted harmonics are suppressed with respect to the desired harmonic. Lastly, the J^{th} harmonic can be any harmonic of the repetition frequency of the optical comb which lies well within the bandwidth of the PD used.

6.2.2 Measured results and discussion

Fig. 6.5 shows the experimental setup for the proposed scheme. Here, the initial optical comb used in the experiment is based on a 2 GHz SC source which is generated by injection of ultrashort pulses from a typical 2 GHz regenerative active mode-locked fiber laser into a length of 2 km HNLF. It should be emphasized that no external microwave reference source is used for the generation of the initial optical comb. The RF spectrum of the mode-locked pulse-train as well as the optical spectrum of the generated SC are shown in Fig. 6.6(a) and Fig. 6.6(b), respectively. The generated SC has ~ 14 nm of bandwidth at the flat region that can support spectral-filtering of a large number of optical combs with different optical carriers.

Fig. 6.5 Experimental setup for the proposed scheme for generating tunable microwave signals. EDFA: erbium-doped fiber amplifier; HNLF: highly nonlinear fiber; PD: photodetector; SMF: single-mode-fiber.

The SC signal is fed to FD-POP which filters the initial optical comb into an array of uniformly-spaced optical combs. We have used a LCOS-based (Liquid-crystal on silicon) wavelength selective switch [116, 124] (Finisar WaveShaper 4000S) as the FD-POP. According to its specification sheet, the minimum passband bandwidth of the WaveShaper is approximately 10 GHz, while the minimum frequency step size is 1 GHz. The array of optical combs is then propagated into a readily-available spool of standard SMF of 2.9159 km length. The output of the SMF is photodetected by a 12 GHz amplified photodiode, with the RF spectrum of the corresponding photocurrent measured using an RF spectrum analyzer (20 KHz - 26.5 GHz). Next, the WaveShaper is programmed to filter the 2 GHz SC into an array of $P = 5$ optical combs, with each optical comb having a total of 9 ($N=4$) comb lines which is equivalent to a bandwidth of 18 GHz for each optical comb. The optical combs are uniformly-spaced by $\Delta f_{approx} = 272$ GHz, whereas $\Delta f_{sp} = 272.9088$ GHz when calculated using (6.12) based on the parameters in the experimental setup. These experimental values are chosen based on (6.7), with the target to generate a 10 GHz

Fig. 6.6 (a) RF spectrum of the pulse-train generated by the 2 GHz regenerative active mode-locked fiber laser, and (b) optical spectrum of the generated 2 GHz SC.

signal. Fig. 6.7(a) illustrates the optical spectrum of the post-filter optical combs, while Fig. 6.7(b) depicts a zoomed-in view of an optical comb.

As the separation between adjacent comb lines (0.016 nm) is very close to the minimum resolution (0.015 nm) of the optical spectrum analyzer (OSA) used by us, the individual comb lines in each optical comb cannot be distinguished in the optical spectrum. This makes the equalization of the amplitude of all comb lines using the WaveShaper difficult. As shown in Fig. 6.7(b), the profile of the optical comb suggests unequal amplitudes for the comb lines and this can be attributed to the amplitude response of the WaveShaper when acting as a bandpass filter.

Fig. 6.7 (a) Optical spectrum of 5 optical combs with a frequency spacing of $\Delta f_{approx} = 272$ GHz ; (b) zoomed-in view of an optical comb; (c) RF spectrum of the generated microwave signal at 10 GHz with a frequency span of 26.5GHz, and (d) RF spectrum of the generated microwave signal at 10 GHz with a frequency span of 200 KHz and RBW of 2 KHz.

On the other hand, if one uses an optical comb with a higher repetition rate, for instance 5 GHz, the amplitude information for each comb line would be visible on the OSA, making it possible to equalize the amplitudes using the WaveShaper. Due to the RF components used for our regenerative active mode-locked fiber laser, we are only able to realize a 2 GHz repetition frequency at the moment. Also, a higher repetition frequency would reduce the flexibility of choosing the frequency of the generated microwave signal.

In the current experiment, for simplicity, each optical comb is treated as a single comb line and attenuated to have approximately the same amplitude. After propagation of the array of optical combs through the SMF and photodetection, we obtain the microwave signal, with its RF spectrum illustrated in Fig. 6.7(c). We can observe the desired 10 GHz signal and that all the other harmonics within the photodiode's bandwidth are suppressed by ~ 30 dB. Substituting the parameters of the experimental setup in (6.7), we obtain more than 33 dB suppression relative to the 5th harmonic for the other harmonics. Therefore, the experimental result is in good agreement with the calculated result.

The small difference in the amount of measured and calculated suppression values can be attributed to the unequal amplitude of the comb lines as mentioned earlier. Furthermore, unlike the simplifying assumption in the analysis, the comb lines near the comb edges can suffer additional undesired phase shift due to the phase response of the FD-POP. This effect is similar to the effect of typical sharp-cutoff filters, where strong phase variations occur for frequencies near the cut-off. This phase shift can also cause distortions in the form of lower suppression of the unwanted harmonics. Still, this level of harmonic suppression, as shown in Fig. 6.7(c), appears better than that reported in [120].

Additionally, with the frequency spacing set much higher than the bandwidth of the PD used, the PD is unable to detect the beating signals between any two optical combs and this is indeed confirmed by Fig. 6.7(c). Fig. 6.7(d) shows the details of the RF spectrum of the generated microwave signal with a smaller frequency span of 200 KHz. The high noise floor in these figures can be attributed to the use of an amplified photodetector as well as the low average optical power at its input. Besides the use of the FD-POP which has an inherent fiber to free space coupling loss of 5-6 dB, the low average optical power is also due to the suppression of substantial amount of spectrum in the initial optical comb in an effort to generate the desired array of optical combs. Implementing an EDFA prior to photodetection in the setup as well as using a high gain low-noise RF amplifier after photodetection can help to improve the noise floor of the generated microwave signal. The stability of the generated microwave signal in terms of power fluctuation is also investigated. Over a period of a few hours, we observe that there is less than 0.1 dB fluctuation in the generated microwave signal.

Frequency-tunability is also verified in the experiment. For the generation of a microwave signal at 12 GHz, we use the WaveShaper to yield $P = 6$ optical combs uniformly-spaced by $\Delta f_{approx} = 228$ GHz ($\Delta f_{sp} = 227.424$ GHz) with each optical comb having a total of 11 ($N = 5$) comb lines. The optical spectrum of the generated optical combs and the RF spectrum of the detected microwave signal are illustrated in Figs. 6.8(a) and Fig. 6.8(b), respectively. It is observed that with respect to the 12 GHz microwave signal, the other harmonics are suppressed by ~ 30 dB.

Fig. 6.8 (a) Optical spectrum of 6 optical combs with a frequency spacing of $\Delta f_{approx} = 228$ GHz, and (b) the corresponding RF spectrum of generated microwave signal at 12 GHz.

Lastly, we generate a microwave signal at 8 GHz, where we have $P=4$ optical combs uniformly-spaced by $\Delta f_{approx} = 342$ GHz ($\Delta f_{sp} = 341.136$ GHz) with each optical comb having a total of 7 ($N=3$) comb lines. The optical spectrum of the generated optical combs and the RF spectrum of the detected microwave signal are illustrated in Fig. 6.9(a) and Fig. 6.9(b), respectively. It is noted that the other harmonics are now suppressed by about 28 dB relative to the 8 GHz microwave signal.

It is pertinent to mention here again that the proposed scheme has the capability to generate microwave signals at higher frequencies, provided a PD with a higher cut-off frequency is used. Since the response speed of the WaveShaper is typically tens to hundreds of milliseconds depending on the material of its liquid crystal, this puts a limit on how fast one can tune the frequency of the generated microwave signal in the

Fig. 6.9 (a) Optical spectrum of 4 optical combs with a frequency spacing of $\Delta f_{approx} = 342$ GHz , and (b) corresponding RF spectrum of generated microwave signal at 8 GHz.

proposed scheme. Furthermore, our scheme is not limited only to using a SC source generated by regenerative active mode-locked fiber lasers. Broadband femtosecond pulse sources, generated by passive mode-locked fiber lasers which do not require any microwave source as well as have low-phase noise, can also be used as the initial optical comb source.

6.3 Photonic Generation of Frequency-Tunable CW Microwave Signals Using a Temporally-Stretched and Chirped Pulse-Train

In the previous section, we have proposed and demonstrated a new frequency-tunable microwave generation technique which is based on generating an array of uniformly-spaced optical combs with identical comb profiles and propagating the

array of combs through a length of standard SMF, followed by photodetection. Although this technique offers a simple approach to generate CW microwave signals whose frequency can be tuned, it requires the use of a rather bulky and expensive component in the form of the Fourier-domain optical processor (FD-POP). Furthermore, due to the limitations in the FD-POP, the unwanted harmonics are only suppressed by about 30 to 35 dB relative to the desired CW microwave signal.

In this section, we present a second technique which only utilizes a mode-locked laser and other passive optical components, but still allows us to generate CW microwave signals with the same frequency-tunability as the previous technique and at the same time, achieves more than 50 dB of suppression. This technique involves the direct photodetection of a temporally-stretched and chirped pulse-train together with its time-delayed replica. This technique is somewhat similar to the technique demonstrated in [89, 125-127], in which widely tunable pulsed signals with frequencies in the GHz to THz range can be generated when a highly chirped pulse together with its variable time-delayed replica are heterodyned at a PD. However, it should be emphasized that in our approach, the pulse-train has to be temporally-stretched in such a way that the pulses overlap with each other, thus permitting generation of CW signals.

Fig. 6.10 Schematic diagram of the proposed scheme for generating tunable microwave signals using a temporally-stretched and chirped pulse-train. OC: optical coupler, OTDL: optical-tunable-delay-line. OSC: oscilloscope.

The principle of operation and the governing mathematics, together with numerical simulations are presented next. To verify the theoretical results, we also present measured results based on the proposed approach. We demonstrate the photonic generation of tunable CW microwave signals ranging from 4 GHz to 14 GHz.

6.3.1 Principle of operation, mathematical expressions, and simulation results

Fig. 6.10 shows the schematic of the proposed scheme. A train of ultrashort optical pulses is first chromatically dispersed in a dispersive element (e.g. SMF) such that the pulses are temporally-stretched and chirped. The temporally-stretched pulse-train is then sent to a typical MZI. An optical time delay line (OTDL) is included in one of the arms of the MZI so as to realize tunable path length difference between the two arms. Consequently, the output at the MZI comprises a superposition of the temporally-stretched pulse-train and its tunable time-delayed replica. The output from the MZI is converted into the electrical domain using a photodetector. The electrical output is monitored using an electronic sampling oscilloscope.

Assuming a Gaussian pulse shape and taking into consideration any initial chirp, the electric field of an ultrashort pulse-train, $E_1(t)$ can be expressed as

$$E_1(t) = E_o \exp\left(\frac{-(1+iC)t^2}{2\tau_o^2}\right) \otimes \sum_{n=-\infty}^{\infty} \delta(t-nT) \quad (6.14)$$

where E_o is the amplitude of each pulse, τ_o is the half-width at $1/e$ of the maximum of the pulse which is related to the pulsewidth at full-width half maximum τ_{FWHM} as $\tau_{FWHM} = 2\sqrt{\ln 2}\tau_o$, C is the chirp parameter defined as $C = (\Delta\omega^2\tau_o^2)^{0.5}$, $\Delta\omega$ is the bandwidth of the pulse-train, T is the repetition rate of the pulse-train and $\otimes$ represents convolution. The repetition frequency of the pulse-train is $f_{rep} = 1/T$.

The pulse-train is chromatically dispersed by a length of SMF so that the pulses are temporally-stretched and chirped; at the output of the SMF the pulse-train $E_2(t)$ can be expressed as

$$E_2(t) = \frac{E_o\tau_o}{\sqrt{1+iC}} \cdot \frac{1}{\sqrt{h}} \exp\left(\frac{-t^2}{2h}\right) \otimes \sum_{n=-\infty}^{\infty} \delta(t-nT) \quad (6.15)$$

where $h = \tau_o^2 / (1+iC) - i\beta_2 L$, β_2 is the 2nd-order dispersion parameter and L is the length of the SMF. Here, we have assumed linear dispersion so that the 3rd-order dispersion effects are considered negligible. This assumption is fairly accurate when $\tau_o |\beta_2 / \beta_3| \geq 1$, where β_3 is the 3rd-order dispersion parameter of the SMF [112].

Based on the typical values of β_2 and β_3 of the SMF, this inequality can be easily satisfied for pulse widths of tens of ps. The analysis and experiment presented in this section are based on this condition. The temporally-stretched and chirped

pulse-train is sent to a MZI which first splits the pulse-train into two paths of different path length. As the path length difference is tunable by the OTDL, this result in the pulse-train in one arm having a tunable time-delay Δt relative to the pulse-train in the other arm. After recombination at the output of the MZI, the electric field $E_2(t)$ can be expressed as

$$E_3(t) = \frac{1}{2} \left\{ E_2 \left(t + \frac{\Delta t}{2} \right) + E_2 \left(t - \frac{\Delta t}{2} \right) \right\} \quad (6.16)$$

Since the photodetector is a square-law device, the resultant photocurrent can be expressed as

$$\begin{aligned} I(t) &= R |E_3(t)|^2 \\ &= \frac{R E_o^2 \tau_o}{4 \sqrt{\tau_1}} \{ x(t) + y(t) + z(t) \} \end{aligned} \quad (6.17)$$

where R is the responsivity of the photodetector and the three time-varying terms on the right-hand-side of (6.17) are respectively expressed as

$$\begin{aligned} x(t) &= \sum_{n=-\infty}^{\infty} \exp \left(\frac{-\left\{ (t - nT) + \Delta t / 2 \right\}^2}{\tau_1^2} \right) \\ &+ 2 \sum_{m < n} \exp \left(\frac{-\left\{ t_{m,n+}^2 + [(n - m)T / 2]^2 \right\}}{\tau_1^2} \right) \cos \left((n - m)T \frac{C_1}{\tau_1} t_{m,n+} \right) \end{aligned} \quad (6.18)$$

$$\begin{aligned}
 y(t) = & \sum_{n=-\infty}^{\infty} \exp \left(\frac{-\{(t-nT)-\Delta t/2\}^2}{\tau_1^2} \right) \\
 & + 2 \sum_{m < n} \exp \left(\frac{-\{t_{m,n}^2 + [(n-m)T/2]^2\}}{\tau_1^2} \right) \cos \left((n-m)T \frac{C_1}{\tau_1} t_{m,n_-} \right)
 \end{aligned} \tag{6.19}$$

$$\begin{aligned}
 z(t) = & 2 \sum_{n=-\infty}^{\infty} \exp \left(\frac{-\{(t-nT)^2 + \Delta t^2/4\}}{\tau_1^2} \right) \cos \left(\frac{C_1}{\tau_1} \Delta t (t-nT) \right) \\
 & + 2 \sum_{m < n} \exp \left(\frac{-\{t_{m,n}^2 + [\Delta t + (n-m)T]^2/4\}}{\tau_1^2} \right) \cos \left(\frac{C_1}{\tau_1} (\Delta t + (n-m)T) t_{m,n} \right) \\
 & + 2 \sum_{m < n} \exp \left(\frac{-\{t_{m,n}^2 + [\Delta t - (n-m)T]^2/4\}}{\tau_1^2} \right) \cos \left(\frac{C_1}{\tau_1} (\Delta t - (n-m)T) t_{m,n} \right)
 \end{aligned} \tag{6.20}$$

where

$$t_{m,n} = t - (n+m)T/2 \tag{6.21}$$

$$t_{m,n_+} = t - [(n+m)T - \Delta t]/2 \tag{6.22}$$

$$t_{m,n_-} = t - [(n+m)T + \Delta t]/2 \tag{6.23}$$

$$\tau_1^2 = \tau_o^2 + 2\beta_2 LC + (\beta_2 L / \tau_o)^2 (1 + C^2) \tag{6.24}$$

$$C_1 = \beta_2 L (1 + C^2) / \tau_o^2 + C \tag{6.25}$$

The half-width at $1/e$ maximum of the pulse-train in the electrical domain is τ and its chirp parameter changes from C to C_1 . The terms in (6.18) and (6.19) appear due to the individual squaring of the relative time-delayed temporally-stretched pulse-

trains $E_2(t + \Delta t / 2)$ and $E_2(t - \Delta t / 2)$ at the photodetector, respectively. On the other hand, terms in (6.20) result from the mixing of $E_2(t + \Delta t / 2)$ and $E_2(t - \Delta t / 2)$ at the photodetector.

For a clearer insight, with some mathematical processing, (6.18), (6.19) and (6.20) are rewritten as follows

$$x(t) = \left[\begin{aligned} &\exp\left(\frac{-\{t + \Delta t / 2\}^2}{\tau_1^2}\right) \\ &+ 2 \sum_{n=m=1}^{\infty} \exp\left(\frac{-[(n-m)T / 2]^2}{\tau_1^2}\right) \cdot \exp\left(\frac{-\{t - [(n-m)T - \Delta t] / 2\}^2}{\tau_1^2}\right) \\ &\cdot \cos\left((n-m)\omega_H \{t - [(n-m)T - \Delta t] / 2\}\right) \end{aligned} \right] \otimes \sum_{n=-\infty}^{\infty} \delta(t - nT) \quad (6.26)$$

$$y(t) = \left[\begin{aligned} &\exp\left(\frac{-\{t - \Delta t / 2\}^2}{\tau_1^2}\right) \\ &+ 2 \sum_{n=m=1}^{\infty} \exp\left(\frac{-[(n-m)T / 2]^2}{\tau_1^2}\right) \cdot \exp\left(\frac{-\{t - [(n-m)T + \Delta t] / 2\}^2}{\tau_1^2}\right) \\ &\cdot \cos\left((n-m)\omega_H \{t - [(n-m)T + \Delta t] / 2\}\right) \end{aligned} \right] \otimes \sum_{n=-\infty}^{\infty} \delta(t - nT) \quad (6.27)$$

$$z(t) = 2 \left[\begin{aligned} & \exp\left(\frac{-\{t^2 - \Delta t^2 / 4\}}{\tau_1^2}\right) \cos(\omega_c t) \\ & + \sum_{n-m=1}^{\infty} \exp\left(\frac{-\{[(n-m)T + \Delta t] / 2\}^2}{\tau_1^2}\right) \cdot \exp\left(\frac{-\{t - (n-m)T / 2\}^2}{\tau_1^2}\right) \\ & \cdot \cos\left(\{(n-m)\omega_H + \omega_c\} \cdot \{t - (n-m)T / 2\}\right) \\ & + \sum_{n-m=1}^{\infty} \exp\left(\frac{-\{[(n-m)T - \Delta t] / 2\}^2}{\tau_1^2}\right) \cdot \exp\left(\frac{-\{t - (n-m)T / 2\}^2}{\tau_1^2}\right) \\ & \cdot \cos\left(\{(n-m)\omega_H - \omega_c\} \cdot \{t - (n-m)T / 2\}\right) \end{aligned} \right] \otimes \sum_{n=-\infty}^{\infty} \delta(t - nT) \quad (6.28)$$

where

$$f_c = \frac{\omega_c}{2\pi} = \frac{C_1 \Delta t}{2\pi \tau_1^2} \quad (6.29)$$

$$f_H = \frac{\omega_H}{2\pi} = \frac{C_1 T}{2\pi \tau_1^2} \quad (6.30)$$

It is clear from (6.29) and (6.30) that both f_c and f_H are a function of the broadened pulsewidth τ_1 and the corresponding chirp C_1 . The relationship in (6.29) is similar to what has been derived in [127]. However, the stretching and overlapping of pulses in our approach results in an additional relationship as given by (6.30). For a fixed amount of dispersion, the frequency f_H is fixed by the repetition rate of the pulse-train while the frequency f_c can be varied by tuning the relative time-delay.

The magnitude squared of the Fourier transform of (6.17) gives the detected RF spectrum. The first term in (6.26) and (6.27) represents a Gaussian spread of discrete low-frequency components centered at DC, while the second term represents a

Gaussian spread of discrete high-frequency components centered at $N.f_H$, where N is an integer. Similarly for (6.28), the first term represents a Gaussian spread of discrete high-frequency components which are centered at a tunable frequency f_c ; the second and third term represent a Gaussian spread of discrete high-frequency components centered at $N.f_H + f_c$ and $N.f_H - f_c$, respectively. It should be emphasized that each individual frequency component is separated from its adjacent neighbor by f_{rep} .

In the case when $\Delta \ll T$, the frequency f_H is much greater than f_c and can possibly exceed the bandwidth ($< 40\text{GHz}$) of a conventional photodetector. Owing to the low-pass filtering effect of such a photodetector, the higher frequency components centered at $N.f_H$, $N.f_H + f_c$ and $N.f_H - f_c$ can be suppressed substantially. Consequently, the detected microwave signal will only consist of the stronger frequency components centered at DC and f_c . For such a case, ignoring the higher frequency terms, the Fourier-transform of the first term in (6.26), (6.27) and (6.28) is shown in (6.31), (6.32) and (6.33), respectively:

$$\begin{aligned}
 & F \left\{ \exp \left(\frac{-\{t + \Delta t / 2\}^2}{\tau_1^2} \right) \otimes \sum_{n=-\infty}^{\infty} \delta(t - nT) \right\} \\
 &= \frac{\sqrt{2\pi}\tau_1}{T} \exp \left(\frac{-\tau_1^2 \omega^2}{4} \right) \cdot \exp \left(\frac{j\omega\Delta t}{2} \right) \cdot \sum_{k=-\infty}^{\infty} \delta \left(\omega - k \frac{2\pi}{T} \right)
 \end{aligned} \tag{6.31}$$

$$\begin{aligned}
 & F \left\{ \exp \left(\frac{-\{t - \Delta t / 2\}^2}{\tau_1^2} \right) \otimes \sum_{n=-\infty}^{\infty} \delta(t - nT) \right\} \\
 &= \frac{\sqrt{2\pi}\tau_1}{T} \exp \left(\frac{-\tau_1^2 \omega^2}{4} \right) \cdot \exp \left(\frac{-j\omega\Delta t}{2} \right) \cdot \sum_{k=-\infty}^{\infty} \delta \left(\omega - k \frac{2\pi}{T} \right)
 \end{aligned} \tag{6.32}$$

$$\begin{aligned}
 & F \left\{ 2 \exp \left(\frac{-\{t^2 - \Delta t^2 / 4\}}{\tau_1^2} \right) \cos(\omega_c t) \otimes \sum_{n=-\infty}^{\infty} \delta(t - nT) \right\} \\
 &= \frac{\sqrt{2}\pi\tau_1}{T} \exp \left(\frac{-\Delta t^2}{4\tau_1^2} \right) \cdot \left[\begin{aligned} & \exp \left(\frac{-\tau_1^2 (\omega + \omega_c)^2}{4} \right) \\ & + \exp \left(\frac{-\tau_1^2 (\omega - \omega_c)^2}{4} \right) \end{aligned} \right] \cdot \sum_{k=-\infty}^{\infty} \delta \left(\omega - k \frac{2\pi}{T} \right) \quad (6.33)
 \end{aligned}$$

Clearly, all the three equations represent a Gaussian spread of discrete frequency components centered either at DC or f_c , with a separation of f_{rep} between adjacent frequency components. In addition, the extent of the Gaussian spread becomes smaller as τ_1 increases. Furthermore, as seen in (6.33), a particular frequency component at $N.f_{rep}$ can attain the maximum possible amplitude when f_c coincides with $N.f_{rep}$; the amplitude of that frequency component reduces when f_c deviates from $N.f_{rep}$. More importantly, in the case when f_c coincides with $N.f_{rep}$, it is possible to generate a microwave signal with a single frequency f_c provided the frequency components adjacent to f_c and DC are suppressed substantially (> 50 dB) relative to f_c .

By setting a target of RF power difference between f_c and its adjacent frequency component to be greater than 50 dB, for instance, the following inequality can be obtained

$$20 \log_{10} \left\{ \exp \left(\frac{-\tau_1^2 \left[(\omega_c \pm \omega_{rep}) - \omega_c \right]^2}{4} \right) \right\} \leq -50 \text{ dB} \quad (6.34)$$

Solution of (6.34) enables us to obtain the minimum value of τ_i for meeting the above target. It should be emphasized that, this value of τ_i will also enable the suppression of frequency components adjacent to DC by > 50 dB, thus allowing the generation of a microwave signal with a single frequency f_c . Based on this minimum value of τ_i and by using (6.24), we can work out the minimum amount of dispersion or length of SMF required in the system which broadens the initial τ_o to τ_i .

Using (6.17), we first perform simulations using an initial optical pulse-train with a repetition rate f_{rep} of 2 GHz, a τ_{FWHM} of 10 ps, and a C of 10. Based on these parameters as well as (6.24) and (6.25), the minimum length of SMF required to generate a microwave signal with a single frequency is calculated as 10.695 km. We first investigate the case when the length of SMF used in the system is shorter than the required minimum length. A 9 km long SMF is used to disperse the pulses such that they are stretched and overlapped in time. With τ_{FWHM} for the temporally-stretched pulses increasing to ~ 533.49 ps, the full pulse-width becomes approximately 1.5 times the inter-pulse spacing. The relative time-delay is set to $\Delta t = 24.09$ ps so that, from (6.29), $f_c = 20$ GHz. On the other hand, we obtain $f_H = 419.64$ GHz from (6.30). For completeness, in the following results, we have also included the combined frequency response of the photodetector, RF cables and connectors as a simple Butterworth low-pass filter of 3rd-order with a cut-off frequency of 40 GHz.

Fig. 6.11 Simulation results for a generated microwave signal at $f_c = 20$ GHz. (a-c) RF spectrum for different frequency spans, and (d) the time-domain waveform.

The RF spectrum of (6.17) is shown in Fig. 6.11(a) and it is observed that several frequencies exist across the frequency span. A closer examination in Figs. 6.11(b) and (c) reveals that each peak in Fig. 6.11(a) is in fact a Gaussian spread of discrete frequency components centered at DC, f_c , or $N.f_H \pm f_c$. Due to the assumed low-pass filtering effect, only the frequencies centered at DC and f_c are strong since the higher frequency components are suppressed significantly. The corresponding time-domain waveform is shown in Fig. 6.11(d). The fast oscillation comprises all the frequencies centered at $f_c = 20$ GHz, while the slow oscillation is due to the residual 2 GHz.

Fig. 6.12 Simulation results for a generated microwave signal with $f_c = 20$ GHz . (a) RF spectrum, and (b) the time-domain waveform.

Next, for the same initial pulse-train, we now increase the amount of total dispersion in the system by using a 15 km long SMF, much longer than the previously calculated minimum length required for generation of microwave signal of single frequency. Compared to the previous case, with τ_{FWHM} for the temporally-stretched pulses increasing to ~ 895.78 ps, the full pulses now broaden to approximately 2 times the inter-pulse spacing. For a valid comparison, we set $\Delta t = 40.40$ ps so that f_c remains at 20 GHz. The value of f_H also shifts to a lower frequency of 247.54 GHz.

Figs. 6.12(a) and (b) present the RF spectrum and the time-domain waveform of the generated microwave signal for this case. In contrast to the previous case, the inset of Fig. 6.12(a) shows that the discrete frequencies adjacent to DC as well as f_c have been suppressed, leaving the 20 GHz signal as the only dominant signal. This is due to the narrowing of the Gaussian spread in the spectrum corresponding to the increased pulse width. Thus, by proper management of dispersion in the system, one can generate a microwave signal of a single frequency. Moreover, by ensuring that f_c coincides with one of the harmonics of f_{rep} , the amplitude of the generated frequency can be maximized.

Fig. 6.13 Simulation results for generated microwave signals with $f_c = 10$ GHz and $f_c = 4$ GHz . (a) RF spectrum, and (c) the time-domain waveform for $f_c = 10$ GHz , (b) RF spectrum, and (d) the time-domain waveform for $f_c = 4$ GHz .

Finally, frequency-tunability based on the proposed scheme is demonstrated in Figs. 6.13(a)-(d) as we set $\Delta t = 20.20$ ps and $\Delta t = 8.08$ ps so as to obtain f_c of 10 GHz and 4 GHz, respectively, for a fiber length of 15 km. Proper tuning of Δt therefore allows us to change the frequency of the generated microwave signal to any harmonic of f_{rep} that lies within the bandwidth of the photodetector. Inclusion of the timing jitter effect, Δt_{ij} in the initial pulse-train gives us the following

$$f_c = \frac{\omega_c}{2\pi} = \frac{C_1 \cdot (\Delta t + \Delta t_{ij})}{2\pi \cdot \tau_1^2} \quad (6.35)$$

It clearly reveals that f_c will have a random shift due to the timing jitter. This will lead to random amplitude fluctuation of the generated microwave signal as f_c deviates randomly around the desired frequency at $N \cdot f_{rep}$. However, this random amplitude fluctuation can be considered negligible when using a pulse-train with a Δt_{ij} in the fs range for cases with Δt in the ps range.

It should also be pointed out that, by using an initial pulse-train with a greater value of C corresponding to a wider bandwidth, the required minimum length of SMF to stretch and overlap the pulses in time can be greatly reduced. This will reduce the corresponding loss and will enable the generation of microwave signals with larger amplitude.

6.3.2 Measured results and discussion

Fig. 6.14 shows the experimental setup for the proposed scheme. In the experiments, a 2 GHz pulse-train, with a τ_{FWHM} of 10 ps, a C of 14 and an output power of 9.25 mW is used. Figs. 6.15(a) and 6.16(a) show the RF spectrum and the

Fig. 6.14 Experimental setup for generating tunable microwave signals using temporally-stretched and chirped pulse-train. OC: optical coupler, OTDL: optical-tunable delay-line, OSC: oscilloscope, VOA: variable optical attenuator, PC: polarization controller.

time-domain waveform of the initial pulse-train, respectively. This optical source is obtained by first amplifying the optical pulses from a 2 GHz regenerative active mode-locked fiber laser using a high power EDFA and subsequently injecting them into a HNLF for SC generation and lastly bandpass filtering a ~ 5 nm band in the flat region of the SC. It should be pointed out that, based on the parameters of our initial pulse-train, the required minimum length of SMF to fulfill the criteria set by (6.34) is approximately 7.66 km. However, we have used a readily-available spool of standard SMF of 27 km to chromatically disperse the pulse-train which results in the pulses being temporally stretched as well as linearly-chirped. Of course, due to the longer fiber length, the generated microwave signals have a relatively low strength.

At this point, there occurs temporal overlapping of the pulses. Subsequently, the temporally stretched pulse-train is sent to an MZI which has a relative time-delay Δt between the two arms, controlled by an OTDL. A VOA is placed in each arm of the MZI so as to balance the optical power of each arm before recombining, while the

polarization controllers are used to align the polarizations of the optical signals. The optical power at the output of the MZI is measured to be 0.4 mW. The output of the MZI is photodetected by a 40 GHz amplified photodiode and the time-domain waveform of the generated microwave signal is measured using a 40 GS/s oscilloscope. The corresponding RF spectrum is measured using an RF spectrum analyzer (20 kHz - 26.5GHz).

Based on the parameters used in the setup, we can calculate from (6.29) the value of Δt that is needed to generate a microwave signal of a specific frequency. We first generate a 4 GHz microwave signal by setting the Δt of the MZI at 14.64 ps. The RF spectrum of the 4 GHz microwave signal and its time-domain waveform are shown in Figs. 6.15(b) and 6.16(b), respectively. Fig. 6.15(b) clearly shows the presence of only the desired 4 GHz microwave signal as the other harmonics of f_{rep} within the bandwidth of the photodetector have been suppressed by more than 60 dB.

It is important to note that, since the pulses overlap with each other due to dispersion, the generated microwave signal as shown in Fig. 6.16(b) is CW in nature. The inset in Fig. 6.15(b) shows a zoom-in view of the 4 GHz microwave signal. Besides the desired signal, some unwanted frequencies of relatively weak amplitude are still present; these are attributed to the supermode noise of the regenerative active mode-locked fiber laser. These unwanted frequencies, which first appear as RIN on the initial pulse-train, lead to the slight distortion in the time-domain waveform of the generated microwave signal. Furthermore, we have observed that the amplitude of the microwave signal is maximum when f_c coincides with $N \cdot f_{rep}$.

By changing Δt to 21.96 ps, 36.60 ps and 51.24 ps, the generated microwave signal is tuned to 6 GHz, 10 GHz and 14 GHz, respectively. These results are presented in Figs. 6.15(c)-(e) and 6.16(c)-(e). Again, the undesired frequency components are suppressed by more than 60 dB demonstrating that this approach can suppress well the undesired frequency components. Significantly, since the microwave signal is generated from a particular mode of the mode-locked pulse source, we observe a good long-term stability of the generated microwave signals. Over a period of more than 2 hours, there is less than 0.05 dB fluctuation in the power of the generated microwave signals, as seen on the RF spectrum analyzer.

The time-domain waveforms of the generated microwave signals are captured using a sampling oscilloscope; therefore signal peaks may not be captured faithfully, particularly for the higher frequencies. It is pertinent to mention here that the slow-varying amplitude envelope as observed in Fig. 6.16 is likely to be caused by the sampling jitter of the sampling oscilloscope used for the measurements. This is supported by the excellent noise suppression seen in the RF spectra in Fig. 6.15. Due to the limited bandwidth of the real-time oscilloscope, we are only able to demonstrate a narrow range of frequency-tunability. However, the proposed scheme is able to generate microwave signals as high as the cut-off frequency of the amplified photodiode at 40 GHz or 20 times f_{rep} . Using a photodetector with a higher cut-off frequency will enable the generation of microwave signals at even higher frequencies.

Fig. 6.15 RF spectrum of (a) 2 GHz initial pulse-train, (b) 4 GHz generated microwave signal, (c) 6 GHz generated microwave signal, (d) 10 GHz generated microwave signal, and (e) 14 GHz generated microwave signal.

Fig. 6.16 Time-domain waveforms corresponding to Fig 6.15. (a) 2 GHz initial pulse-train, (b) 4 GHz generated microwave signal, (c) 6 GHz generated microwave signal, (d) 10 GHz generated microwave signal, and (e) 14 GHz generated microwave signal.

Fig. 6.17 Comparison of SSB phase noise spectrum between (blue) 10 GHz microwave signal bandpass filtered directly from 2GHz regenerative active mode-locked fiber laser, (red) 10 GHz microwave signal generated based on our scheme and (black) 10 GHz microwave signal generated directly from a RF signal generator.

Additionally, we also measured the phase noise of the generated microwave signals. Fig. 6.17 depicts the measured SSB phase noise for the generated 10 GHz microwave signal (red). For comparison, we have also included the SSB phase noise spectrum for a 10 GHz microwave signal generated by bandpass filtering the 10 GHz mode of the 2 GHz regenerative active mode-locked fiber laser using a cavity filter centered at 10 GHz (blue). Clearly, we observe a degradation in the phase noise and this can be attributed to the additional phase noise introduced by the components [high power EDFA, MZI, PD, etc.] used in our demonstrated system. We then compare these phase noise results to the SSB phase noise spectrum (black) corresponding to a 10 GHz microwave signal generated directly by a RF signal generator (Anritsu 68367C). Evaluating in terms of timing jitter Δt_j , we obtained a timing jitter of 100 fs (blue), 188 fs (red) and 145 fs (black) for the three different cases, respectively.

Compared to the third case (black), it is observed that there are two regions in the phase noise spectra of the first two cases which have higher phase noise. The higher phase noise at low offset frequencies (< 2 KHz) is mainly due to the use of non low-phase noise amplifiers in the regenerative active mode-locked fiber laser setup. The higher phase noise at higher offset frequencies (> 200 KHz) results from the ASE noise introduced by the high power EDFA used in the setup. Thus, in order to improve the current phase noise performance of our generated microwave signals to be comparable to at least the RF signal generator which we have used for comparison, one can improve the existing fiber laser design using low-phase noise amplifiers as well as a low noise figure high power EDFA in the microwave generation setup.

Based on the current experimental setup, we have shown that the power level of the generated microwave signals is approximately -20 dBm. Replacing the 27 km SMF by an 8 km SMF would improve the power level of the microwave signals. A simple approach to further shorten the required fiber length and reduce the loss in the system is to bandpass filter a wider bandwidth from the SC compared to the 5 nm bandwidth which we have used. Moreover, as a long SMF is bulky in size, one can make use of a LCFBG in place of the SMF to provide the same amount of dispersion and to reduce the system size.

6.4 Summary

In this chapter, we have proposed and demonstrated two photonic techniques which are based on the use of mode-locked lasers and which enable the generation of frequency-tunable CW microwave signals with significant suppression of unwanted frequencies components.

In the first part of the chapter, we have shown analytically and experimentally a new frequency-tunable CW microwave generation technique which is based on generating an array of uniformly-spaced optical combs with identical comb profiles and propagating the array of combs through a length of standard SMF, followed by photodetection. The proposed scheme does not require an external microwave reference source or modulators. For an initial optical comb with a fixed repetition rate as well as using fixed length of SMF, we have shown mathematically and through calculations that, by choosing appropriate values for the number of optical combs, number of comb lines in each optical comb as well as the frequency spacing between the adjacent optical combs, one can tune the frequency of the generated single-tone microwave signal to any integer multiple of the optical comb's repetition frequency which lies within the bandwidth of the PD.

We have experimentally demonstrated the generation of single-tone microwave signals up to 12 GHz using an initial 2 GHz optical comb and a FD-POP which spectrally-filters the initial optical comb into an array of uniformly-spaced optical combs. Frequency-tunability has also been experimentally verified by generation of microwave signals at 8 and 10 GHz. The generated CW microwave signals are noted to have ~ 30 to 33 dB suppression of the unwanted frequency components. The generated microwave signals are observed to have very good long-term stability.

To overcome the limited suppression of the unwanted frequency components as well as to avoid the use of a bulky and expensive component such as the FD-POP, a second technique is introduced in the second part of the chapter. In this part, we have shown analytically and experimentally the generation of frequency-tunable CW microwave signals based on the direct photodetection of a temporally-stretched and chirped pulse-train together with its time-delayed replica. Through numerical

simulations, it has been shown that by properly managing the dispersion in the system as well as the relative time-delay of the MZI, one can tune the frequency of the generated microwave signal to any integer multiple of the pulse source's repetition frequency which lies within the bandwidth of the photodetector.

We have experimentally demonstrated the generation of CW microwave signals up to 14 GHz using an initial 2 GHz mode-locked pulse-train. Frequency-tunability has also been experimentally verified by generation of microwave signals at multiples of the repetition frequency of the initial pulse-train. The highest frequency that can be generated using this scheme is only limited by the bandwidth of the photodetector. More than 50 dB suppression of the unwanted frequency components is achieved for the generated CW microwave signals.

Chapter 7 PHOTONIC GENERATION OF LINEARLY-CHIRPED MICROWAVE SIGNALS

7.1 Introduction

Leveraging on the advantageous features offered by photonic components, such as low loss, broad bandwidth, immunity to electromagnetic interference and high rate-distance product, the area of microwave photonics has attracted extensive research interest over the past few years [1]. Furthermore, the area of photonic generation of arbitrary microwave waveforms has also become increasingly important due to its diverse applications such as fiber wireless communication, pulsed radar, as well as electronic countermeasures.

With the conventional electronic arbitrary waveform generators limited to relatively low frequencies, much effort has been put into developing photonic techniques which can generate microwave waveforms at frequencies ranging from tens to hundreds of GHz. Some of these techniques include direct-space-to-time (DST) optical pulse shaping [77, 79], temporal pulse shaping [80-83] and optical line-by-line pulse shaping [84-87]. The operating principles and the drawbacks of these techniques have been reviewed in Chapter 3. The first technique requires bulky free-space optical components, the second one requires modulators and microwave sources, and the last one requires using a spatial-light modulator (SLM) that suffers from a slow response time. In most of these techniques, the upper-end frequency of the generated microwave waveforms is generally limited by the bandwidth of the photodetector used.

In recent years, the spectral shaping and frequency-to-time mapping technique has shown promise in generating microwave waveforms with complex arbitrary profiles, such as linearly-chirped waveforms [88-93]. Proposed and demonstrated in [90], Zeitouny *et al.* generated linearly-chirped microwave waveforms by using two LCFBG with different chirp rates in an unbalanced MZI based on the optical spectral shaping and linear frequency-to-time mapping technique. As the fabricated LCFBGs possessed a fixed 2nd-order dispersion term, it was shown that tuning the relative time-delay between the interferometer arms would change the centre frequency while keeping the chirp rate constant for the microwave waveform.

An alternative approach to generate linearly-chirped microwave waveforms was demonstrated by Wang *et al.* [91]. Unlike [90], spectral shaping was accomplished by an LCFBG-incorporated Sagnac loop mirror (spectral filter), which has a spectral response with an increasing or decreasing FSR, while linear frequency-to-time mapping was realized by a dispersive medium. It was shown that by tuning the time-delay in the Sagnac loop mirror, the centre frequency and chirped profile of the generated microwave waveform could be controlled. Specifically, depending on the polarity of the time-delay in the Sagnac loop mirror, the chirp rate could be tuned to either negative or positive values but its absolute value was kept constant. This limitation is clearly attributed to the non-tunable nature of the spectral filter's spectral response.

Besides linear frequency-to-time mapping, Ashrafi *et al.* have shown that nonlinear frequency-to-time mapping together with spectral shaping can also be implemented in an unbalanced MZI setup to generate linearly-chirped microwave waveforms [92]. Essentially, the nonlinear frequency-to-time mapping is induced by a dispersive medium having both the 2nd and the 3rd-order dispersion terms. More

importantly, it has been demonstrated that by varying the relative time-delay between the interferometer arms, this scheme permits the generation of microwave waveforms whose chirp rate can be tuned to negative, zero or positive values. Despite the tunability of the chirp rate, the centre frequency of the generated microwave waveforms however changes when the chirp rate is tuned.

Recently, Li *et al.* have proposed and demonstrated the generation of microwave waveforms with continuously tunable linear chirp rate based on the spectral shaping and linear frequency-to-time mapping technique [93]. In contrast to [91], the spectral filter is realized by an unbalanced MZI incorporating an optically pumped LCFBG whose spectral response can be tuned to have increasing or decreasing FSR. As the optically pumped LCFBG is written in an Erbium-Ytterbium co-doped fiber, pumping it with different pump powers changes its group-delay response and this leads to a shift in the FSR. It should be emphasized that, unlike all the previously mentioned schemes, the centre frequency and chirp rate can be tuned independently.

In general, for a fixed value of the 2nd-order dispersion term of the dispersive medium which induces the linear frequency-to-time mapping, the centre frequency of the microwave waveform changes with the relative time-delay between the interferometer arms, while the chirp rate is dependent on both the polarity of the relative time-delay as well as the GDD induced by the optically pumped LCFBG. It is shown experimentally that the chirp rate of the generated microwave waveforms changes from 79 to 64 GHz/ns when the injection current to the pumping laser diode increases from 0 to 100mA.

Fig. 7.1 Schematic of the proposed scheme for generating linearly-chirped microwave pulses. SMF: single-mode-fiber; LCFBG: linearly-chirped fiber-Bragg grating; OTDL: optical-tunable-delay-line; BPD: balanced photodetector.

In this chapter, we show analytically and experimentally the generation of linearly-chirped microwave waveforms based on an approach similar to that proposed in [93]. But, instead of an optically pumped LCFBG, an LCFBG which has a programmable GDD is used in our scheme. Our scheme also employs balanced photodetection and offers the following advantages. The dispersion of the LCFBG can be tuned from -900 ps/nm to $+900$ ps/nm, enabling the generation of linearly-chirped microwave waveforms with a wider chirp tuning range. The tuning range in our scheme also covers the zero-chirp case and the generated microwave waveforms are DC-free due to the balanced photodetection. The principle of operation of our scheme and its governing mathematics is presented in the next section, followed by the measured results and their comparison with simulation results.

7.2 Principle of operation and mathematical expressions

Fig. 7.1 shows a schematic of the proposed scheme. An ultrashort pulse $g(t)$ is first propagated through a length L of SMF whose GDD is defined as $GDD_S = \beta_2 L$, where β_2 is the 2nd-order dispersion term. The temporally-stretched and linearly-

chirped pulse is then fed to an unbalanced MZI which splits the pulse into two independent paths. As depicted in Fig. 7.1, the pulse propagating in the upper arm is further chirped by an LCFBG which has a magnitude response (reflection) $|r(\omega)|$ as well as a programmable GDD of GDD_F . On the other hand, after passing through an OTDL, the pulse propagating in the lower arm experiences a tunable time-delay Δt relative to the pulse in the upper arm. When the output pulses $E_1(t)$ and $E_2(t)$ from both arms combine, a temporal interference pattern with increasing or decreasing FSR is generated. With the interference patterns having a phase difference of $\pi/2$ for the outputs of the MZI, followed by O/E conversion by a balanced photodetector, a DC-free chirped microwave waveform can be obtained.

Assuming the initial pulse to be Gaussian-shaped, so that $g(t) = \exp(-t^2 / 2\tau_o^2)$, τ_o = half pulsewidth at $1/e$ maximum, and following the detailed derivation in [93], the individual photocurrents $I_1(t)$ and $I_2(t)$ are expressed as

$$I_1(t) \propto \left[A_1 \left| r\left(\frac{t_1}{GDD_T}\right) \right| G\left(\frac{t_1}{GDD_T}\right) \right]^2 + \left[A_2 G\left(\frac{t_2}{GDD_S}\right) \right]^2 + A_1 A_2 \left| r\left(\frac{t_1}{GDD_T}\right) \right| G\left(\frac{t_1}{GDD_T}\right) G\left(\frac{t_2}{GDD_S}\right) \left[2 \cos\left(\frac{t_2^2}{2GDD_S} - \frac{t_1^2}{2GDD_T}\right) \right] \quad (7.1)$$

$$I_2(t) \propto \left[A_1 \left| r\left(\frac{t_1}{GDD_T}\right) \right| G\left(\frac{t_1}{GDD_T}\right) \right]^2 + \left[A_2 G\left(\frac{t_2}{GDD_S}\right) \right]^2 + A_1 A_2 \left| r\left(\frac{t_1}{GDD_T}\right) \right| G\left(\frac{t_1}{GDD_T}\right) G\left(\frac{t_2}{GDD_S}\right) \left[2 \sin\left(\frac{t_2^2}{2GDD_S} - \frac{t_1^2}{2GDD_T}\right) \right] \quad (7.2)$$

where $A_1(t)$ and $A_2(t)$ are the amplitudes of $E_1(t)$ and $E_2(t)$, $G(\omega)$ is the Fourier transform of $g(t)$, $t_1 = t - \Delta t / 2$, $t_2 = t + \Delta t / 2$ and $GDD_T = GDD_F + GDD_S$. Here, it

should be noted that $g(t)$ is real-time Fourier-transformed under the assumption that the condition $|\tau_o^2 / 2GDD_T| \ll 1$ is satisfied. The output photocurrent $I(t)$ is given as

$$\begin{aligned}
 I(t) &= I_1(t) - I_2(t) \\
 &\propto A_1 A_2 \left| r\left(\frac{t_1}{GDD_T}\right) \right| \cdot G\left(\frac{t_1}{GDD_T}\right) \cdot G\left(\frac{t_2}{GDD_S}\right) \\
 &\quad \cdot \left[\cos\left(\frac{t_2^2}{2GDD_S} - \frac{t_1^2}{2GDD_T}\right) - \sin\left(\frac{t_2^2}{2GDD_S} - \frac{t_1^2}{2GDD_T}\right) \right] \\
 &\propto A_1 A_2 \left| r\left(\frac{t_1}{GDD_T}\right) \right| \cdot G\left(\frac{t_1}{GDD_T}\right) \cdot G\left(\frac{t_2}{GDD_S}\right) \cdot \left[\cos\left(\frac{t_2^2}{2GDD_S} - \frac{t_1^2}{2GDD_T} + \frac{\pi}{4}\right) \right]
 \end{aligned} \tag{7.3}$$

where it is observed that the instantaneous level of the generated microwave waveform oscillates according to the $\cos(t_2^2 / 2GDD_S - t_1^2 / 2GDD_T + \pi / 4)$ term while its envelope is dependent on the remaining terms.

The instantaneous frequency of the microwave waveform can be derived by differentiating the phase of the cosine term $\phi_{RF} = t_2^2 / 2GDD_S - t_1^2 / 2GDD_T + \pi / 4$ w.r.t. time and is expressed as

$$f_{RF,inst} = \frac{1}{2\pi} \frac{\delta \phi_{RF}}{\delta t} = \frac{1}{2\pi} \left[\left(\frac{GDD_S - GDD_T}{GDD_S GDD_T} \right) t - \left(\frac{GDD_S + GDD_T}{GDD_S GDD_T} \right) \frac{\Delta t}{2} \right] \tag{7.4}$$

Clearly, $f_{RF,inst}$ is observed to be linearly proportional to time and therefore the generated microwave waveform is linearly-chirped. Furthermore, the centre frequency of the waveform can be calculated when $t=0$ and is expressed as

$$f_{centre} = \frac{1}{4\pi} \left| \frac{\Delta t}{GDD_F + GDD_S} + \frac{\Delta t}{GDD_S} \right| \tag{7.5}$$

From (7.5), it is evident that f_{centre} is proportional to Δt . Since GDD_S is generally fixed, tuning Δt would change the f_{centre} of the linearly-chirped microwave waveform. Furthermore, the chirp rate of the microwave waveform can be written as

$$CR = \frac{\delta f_{RF,inst}}{\delta t} = \frac{1}{2\pi} \text{sgn}(\Delta t) \left[\frac{GDD_F}{(GDD_F + GDD_S) GDD_S} \right] \quad (7.6)$$

Unlike f_{centre} , the absolute value of the chirp rate is independent of the magnitude of Δt but tunable by changing the value of GDD_F . Moreover, both up-chirp and down-chirp are obtainable as the polarity of the chirp rate is dependent on the sign of Δt . Lastly, tuning the GDD_F of the programmable LCFBG to zero would also enable the generation of zero-chirp microwave waveforms in this scheme.

7.3 Measured results and discussion

Fig. 7.2 illustrates the experimental setup for realization of the proposed scheme. In the experiments, the initial pulse source is implemented by a homemade 68 MHz mode-locked laser which has a FWHM pulsewidth of 2.5 ps, a FWHM bandwidth of 2 nm and a centre wavelength of 1565.3 nm. A readily-available 27 km long SMF with standard dispersion parameter $D_\lambda = (2\pi c / \lambda^2) \beta_2 = 17$ ps/nm/km is utilized to temporally-stretch as well as linearly chirp the pulse-train. Hence, GDD_S can be calculated to be ~ -596.63 ps². Next, the temporally-stretched pulse-train is fed to an unbalanced MZI. For the realization of the programmable LCFBG in arm 1, a single channel is bandpass filtered from a commercially available tunable multi-channel dispersion compensation module which comprises multiple-channels of similar group-

Fig. 7.2 Experimental setup of the proposed scheme for generating linearly-chirped microwave pulses. SMF: single-mode-fiber; LCFBG: linearly-chirped fiber-Bragg grating; OTDL: optical-tunable-delay-line; VOA: variable optical attenuator; OBPF: optical bandpass filter; TDCM: tunable multi-channel dispersion compensation module; BPD: balanced photodetector.

Fig. 7.3 Magnitude response (reflection) of the filtered channel serving as the programmable LCFBG in the proposed approach.

delay characteristics. The magnitude response of the filtered channel has a 1-dB bandwidth of 0.5 nm and is centered at 1565.3 nm as depicted in Fig. 7.3.

Based on two superimposed FBGs connected to an optical circulator, the dispersion $D_F = (2\pi c / \lambda^2) GDD_F$ of the implemented TDCM can be tuned from -900 ps/nm to $+900$ ps/nm with a resolution of 1 ps/nm. In other words, GDD_F can be tuned from -1169.99 ps² to $+1169.99$ ps². As for arm 2, the OTDL is used to control the tunable relative time-delay between the interferometer arms, while the VOA is

used to maximize the fringe visibility of the interference pattern at the outputs of the MZI. The MZI outputs are sent to a balanced photodetector with a 42 GHz bandwidth for OE conversion and removal of any DC component. Finally, the generated DC-free linearly-chirped microwave waveform is RF amplified and observed using a 70 GHz sampling oscilloscope.

In the first experiment, we target the generation of a zero chirp microwave waveform by setting $\Delta t = 100 \text{ ps}$ and $GDD_F = 0 \text{ ps}^2$. The centre frequency and chirp rate are calculated as 27.2 GHz and 0 GHz/ns from (7.5) and (7.6), respectively. Fig. 7.4(a) illustrates the generated microwave waveform which can be observed to be DC-free. Based on the Hilbert-transform method, the corresponding instantaneous frequency is calculated and also plotted in Fig. 7.4(a) wherein the centre frequency and chirp rate are clearly indicated. It can be observed that the generated microwave waveform has nearly zero chirp and its centre frequency of $\sim 27.24 \text{ GHz}$ is in close agreement with the theoretically calculated centre frequency. Besides this, for comparison, a numerically simulated result based on the values of the experimental parameters is presented in Fig. 7.4(b), where we have assumed the LCFBG to have a 6th-order super-Gaussian profile and a 1-dB bandwidth of 0.5 nm. It is seen that, in terms of the generated waveform profile, centre frequency, and chirp rate, the experimental and simulation results are in good agreement.

Next, generation of linearly-chirped microwave waveforms is investigated. Fig. 7.5(a) illustrates the experimentally generated microwave waveform when we set $\Delta t = 0 \text{ ps}$ and $GDD_F = -519.94 \text{ ps}^2$. It can be seen that the generated microwave waveform has a nearly symmetrical profile centered at $\sim 0 \text{ GHz}$ while its instantaneous frequency changes linearly from down-chirp ($\sim -126.7 \text{ GHz/ns}$) to up-

Fig. 7.4 (a) Experimental and (b) simulation results for generation of zero chirp microwave waveform ($\Delta t = 100$ ps, $GDD_F = 0$ ps²). Green line: amplitude of the generated microwave waveform. Blue line: instantaneous frequency of the generated microwave waveform.

Fig. 7.5 (a) Experimental and (b) simulation results for generation of linearly-chirped microwave waveform ($\Delta t = 0$ ps, $GDD_F = -519.94$ ps²). Green line: amplitude of the generated microwave waveform. Blue line: instantaneous frequency of the generated microwave waveform.

chirp ($\sim +120.8$ GHz/ns). Again, our experimental results agree well with the simulation results shown in Fig. 7.5(b).

Fig. 7.6 (a) Experimental and (b) simulation results for generation of linearly-chirped microwave waveform ($\Delta t = -100$ ps, $GDD_F = -194.98$ ps²). Green line: amplitude of the generated microwave waveform. Blue line: instantaneous frequency of the generated microwave waveform.

Fig. 7.7 Experimental and (b) simulation results for generation of linearly-chirped microwave waveform ($\Delta t = 100$ ps, $GDD_F = -324.96$ ps²). Green line: amplitude of the generated microwave waveform. Blue line: instantaneous frequency of the generated microwave waveform.

In order to generate a microwave waveform which is either linearly up-chirped or down-chirped, Δt has to be set to an appropriate value such that the zero instantaneous frequency is shifted out of the waveform's main-lobe. This is

exemplified in Fig. 7.6(a) when we set $\Delta t = -100$ ps and $GDD_F = -194.98$ ps² in which a linearly down-chirped microwave waveform is generated. With a centre frequency of ~ 26.46 GHz, the instantaneous frequency of the waveform decreases linearly from ~ 41 GHz to ~ 10 GHz, equivalent to a down-chirp rate of ~ -67.37 GHz/ns. Lastly, Fig. 7.7(a) illustrates a generated microwave waveform which has a centre frequency of ~ 25.42 GHz and is observed to have a linear up-chirp of ~ 94.1 GHz/ns, with the instantaneous frequency increasing from ~ 4 GHz to ~ 43 GHz. This is realized when we set $\Delta t = 100$ ps and $GDD_F = -324.96$ ps². In the last two examples, the centre frequency and the chirp rate are again in good agreement with the corresponding simulation results shown in Fig. 7.6(b) and 7.7(b), respectively.

It is pertinent to mention here that the mismatch in the amplitude between the experimental and simulation results at high instantaneous frequencies is mainly attributed to the frequency response of the RF link (comprised of the balanced photodetector, RF amplifier and RF cables) as well as the limited bandwidth of the sampling oscilloscope. Use of an RF link with a high roll-off frequency as well as a higher sampling rate oscilloscope would provide a more accurate measurement of the microwave waveform in terms of its amplitude at high frequencies. Using (7.6) and the experimental parameters (GDD_F can be tuned from -1169.99 ps² to $+1169.99$ ps² with a minimum resolution of ~ 13 ps²) of the LCFBG, the chirp rate tuning range for the generated microwave signals can be calculated and is depicted in Fig. 7.8. Considering only a portion of the LCFBG's full GDD_F tuning range (-1169.99 ps² to $+260$ ps²), it is observed that the chirp rate can be tuned continuously across a wide range, extending from ~ -206 GHz/ ns to ~ 206 GHz/ ns. Potentially, higher chirp rates beyond 500 GHz/ns can be obtained by setting the GDD_F of the LCFBG to an

Fig. 7.8 Chirp rate tuning range of the generated microwave signal when GDD_F is tuned from -1169.99 ps^2 to $+260 \text{ ps}^2$.

appropriate value within its tuning range, however such chirp rates are difficult to be measured accurately due to the limitations in our current detection and measurement apparatus/equipments.

7.4 Summary

In this chapter, we have shown analytically and demonstrated experimentally a scheme for the generation of microwave waveforms of zero-chirp, up-chirp and down-chirp based on the temporal interference of two temporally-stretched pulses which are relatively time-delayed and have different frequency chirps. The scheme makes use of an unbalanced MZI, incorporating a tunable time delay and an LCFBG with tunable GDD.

Through analytically derived equations, it has been shown that by tuning to appropriate values the relative time-delay between the two arms of the MZI as well as the GDD of the LCFBG, one can generate a microwave waveform with desired chirp rate and centre frequency. The chirp rate can be tuned by changing the GDD of the

LCFBG, while the centre frequency can be tuned by changing the relative time-delay. Balanced photodetection has also been proposed for the removal of DC components in the generated microwave waveform.

Experimentally, we have shown the generation of several microwave waveforms with different centre frequencies, chirp rates ranging from ~ -126.7 GHz/ns to $\sim +120.8$ GHz/ns, and no DC component. Based on the specifications of the LCFBG which we have used as well considering the limitations in our current detection and measurement apparatus/equipments, our scheme has the potential to tune the chirp rate of the generated microwave waveforms continuously from ~ -206 GHz/ ns to $\sim +206$ GHz/ns.

Chapter 8 CONCLUSION AND FUTURE WORK

8.1 Achievements and Conclusion

In this thesis, two topics in the field of Microwave Photonics have been discussed. These are photonic A/D conversion and photonic generation of microwave signals. Both topics are important in advanced systems such as radar, surveillance, and wireless communication. The techniques considered in this thesis for both of these topics also involve the use of mode-locked lasers, providing another common link between the two topics in this thesis.

In the first part of the thesis (Chapters 1-3), we have reviewed the developments in the techniques for photonic A/D conversion and photonic generation of microwave signals. For photonic A/D conversion, the working principles together with the mathematical derivations for two particular techniques that have been found promising in the recent years, optically-sampled ADC as well as photonic time-stretched ADC (TS-ADC) has been presented. Furthermore, multi-channel interleaving schemes for optically-sampled ADC and continuous-time operation for photonic TS-ADC have also been described. Likewise, for photonic generation of microwave signals, the working principles of various approaches to generating CW and chirped microwave signals have been reviewed. The presented material describes fundamental concepts for the two topics, provides a review of the previous literature on these topics, and points out the shortcomings of the previously reported approaches so as to form the basis and explain the motivation for the research work reported in the rest of the thesis.

The second part of the thesis (Chapters 4-5) first presents the investigations on two photonic ADC schemes. For the first scheme, utilizing complementary dual-output modulation and balanced photodetection scheme, we have demonstrated a 2 GS/s optically-sampled ADC system which achieves a measured 5.52 SNR-bits, SFDR of 28 dB as well as ENOB of ~ 4.29 -bits for a single-tone signal at 200 MHz. By applying arcsine operation on the digitized data to suppress the 3rd-order distortion introduced by the nonlinearity of the dual-output MZM, we have obtained a significant improvement in the SFDR which is now measured to be ~ 62 dB. Measuring across a range of analog input frequencies and operating the current system in undersampling mode, we measured and obtained an average ENOB of ~ 5.34 -bits after linearization. Unlike some other reported schemes, our scheme requires only one set of well-calibrated electronic detection and quantization channel.

For the second scheme, utilizing polarization modulation and balanced photodetection scheme, we have demonstrated a photonic TS-ADC based on virtual time gating principle that can permit continuous-time operation. Digitizing a single-tone signal at 950 MHz, we have measured an ENOB of ~ 3.56 bits for a single channel which is sampled at 2 GS/s. Based on these results, we estimate that with adequate backend digitizing hardware, a four-channel continuous-time TS-ADC with a sampling rate of 8 GS/s can be realized to handle RF frequencies as high as 4 GHz. This scheme offers a simpler practical implementation than other reported schemes based on a complementary dual-output MZM.

Following the above mentioned investigations of the photonic ADC systems, we also present a scheme that utilizes dispersion and sectional compression to generate a Time- and Wavelength-Interleaved Pulse-Train (TWIPT) which potentially can be applied in high-speed optically sampled ADC system based on WDD parallel

processing. This scheme involves propagating an initial broadband ultrashort pulse into a length of SMF and subsequently through a tunable multi-channel dispersion compensation module which has similar tunable group-delay characteristics for all the channels. Through theoretical analysis and simulations, it has been shown that by selecting appropriately the length of the SMF as well as the dispersion of the channels one can generate the desired TWIPT. Experimentally, using an initial 2 GHz SC source, we have demonstrated the generation of a 10×2 GHz TWIPT with tunable pulsewidths. Compared to other reported approaches, our scheme offers simplicity in practical implementation and does not require modulators and microwave sources.

The third part of the thesis (Chapters 6-7) first presents two photonic techniques which enable the generation of frequency-tunable CW microwave signals. Both techniques utilize a mode-locked laser. The first technique uses a Fourier-domain programmable optical processor (FD-POP) to spectrally-carve out an array of uniformly-spaced optical combs with identical comb profiles from an initial SC source generated by a mode-locked laser. The array of combs is then propagated through a length of standard SMF, followed by photodetection. Through theoretical analysis and simulations, it has been shown that by choosing appropriate values for the number of optical combs, number of comb lines in each optical comb as well as the frequency spacing between the adjacent optical combs, one can tune the frequency of the generated single-tone microwave signal to any integer multiple of the optical comb's repetition frequency which lies within the bandwidth of the PD. Experimentally, using an initial 2 GHz SC source, we have demonstrated the generation of CW microwave signals at 8, 10 and 12 GHz with ~ 30 to 33 dB suppression of the unwanted frequency components relative to the desired microwave signal.

The second technique makes use of only passive components in the setup wherein a temporally-stretched and chirped pulse-train together with its time-delayed replica are heterodyned at a PD. Through theoretical analysis and simulations, it has been shown that by properly managing the dispersion in the system as well as the relative time-delay of the MZI, once again one can tune the frequency of the generated microwave signal to any integer multiple of the pulse source's repetition frequency which lies within the bandwidth of the PD. Experimentally, using an initial 2 GHz SC source, we have demonstrated the generation of CW microwave signals from 2 GHz up to 14 GHz with more than 50 dB suppression of the unwanted frequency components relative to the desired microwave signal. It should be emphasized that in both this and the previous technique, modulators and microwave sources are not required.

Following the investigations of the abovementioned photonic techniques to generate frequency-tunable CW microwave signals, we have also presented a scheme to generate linearly-chirped microwave signals based on the optical spectral shaping and wavelength-to-time mapping principle. We have experimentally demonstrated the generation of linearly-chirped microwave signals whose chirp rate as well as centre frequency can be tuned independently. In essence, a chirp rate ranging from ~ -126.7 GHz/ns to $\sim +120.8$ has been obtained in the experiment. Potentially, our scheme allows the chirp rate of the generated microwave waveforms to be tuned continuously from ~ -206 GHz/ ns to $\sim +206$ GHz/ns, covering the positive, zero and negative chirp rate cases.

8.2 Future Work

Although in this thesis we have presented a number of schemes pertaining to photonic A/D conversion and photonic generation of microwave signals which offer features not present in the previously reported schemes, there are some possible directions to extend the present work. These directions are mentioned in the following.

1. Making use of the 10x2 GHz Time- and Wavelength-Interleaved Pulse-Train (TWIPT), which has been reported in Chapter 5, as the initial pulse source as well as adopting the complementary dual-output modulation and balanced photodetection scheme that has been discussed in Chapter 4, one can explore a 20 GHz wavelength-interleaved optically-sampled ADC system.
2. In Chapter 5, we have demonstrated a 10x2 GHz TWIPT based on dispersion and sectional compression. However, it is noted that the generated TWIPT has a slight amplitude variation for pulses at different wavelengths. Such amplitude variation is undesirable when the TWIPT is used for optical sampling and therefore should be minimized. One method to do so is to use the FD-POP which has been used in Chapter 6 to ensure channel equalization. Besides this, nonlinear GVD due to the higher order dispersion terms may distort the pulse shape of the different wavelength channels as well as induce a non-uniform time-delay between adjacent wavelength channels. Therefore, it is important to take into consideration the wavelength dependence of the GVD. Again, one can also make use of the FD-POP to compensate for the higher order dispersion terms. Since the generated TWIPT is intended for use in an optically-sampled ADC system, the phase noise and intensity noise of each wavelength channel have to be characterized and investigated.

3. As described in Chapter 6, the direct photodetection of a temporally-stretched and chirped pulse-train together with its time-delayed replica allows us to generate frequency-tunable CW microwave signals with large suppression ratio over the unwanted frequency components. However, the current setup is bulky since it utilizes an active mode-locked laser, high power amplifier, HNLF and a large length of SMF. A more compact setup can be realized by implementing linear cavity passive mode-locked fiber laser as the initial pulse source and replacing the long SMF with an LCFBG with equivalent dispersion.

Lastly, in each presented scheme, we have also included suggestions on how the current performance can be improved. When improved with the suggested conditions, the issue of practicability of our schemes becomes another important issue to be thought about. In all of our demonstrated schemes, we have used discrete optical and electronic components and this eventually led to bulky and inconvenient systems. In order to be viable and useful alternatives to their corresponding electronic counterparts, the practical issues have to be addressed. For commercial applications, form-factor, cost, and performance are the key considerations. Therefore, photonic integration would be the key enabler for the microwave photonic versions to become competitive commercially. By integrating the different building blocks on a single chip, benefits such as miniaturization, less power consumption, robustness, functional scalability, reduced complexity as well as increased in volume manufacturing can be obtained.

In the context of our demonstrated schemes, integration of optical components would most likely be done in the indium phosphide platform as the wavelengths of the discrete optical components used reside in the telecommunication wavelengths. As for integration of optical and electronic components, silicon will likely be used as the

base material. One example of a device enabled by silicon photonics and photonic-electronic integration is a photonic ADC chip which has been demonstrated recently in [27]. In general, the chip comprises both photonic and electronic components, namely microring-resonator filters, silicon carrier-depletion modulator, and germanium photodetectors, electronic ADCs, and digital postprocessing circuits

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