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OPTICAL FREQUENCY DOMAIN REFLECTOMETRY SYSTEM BASED ON SEMICONDUCTOR LASERS

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OPTICAL FREQUENCY DOMAIN REFLECTOMETRY
SYSTEM BASED ON SEMICONDUCTOR LASERS

BY

ZHEYI YAO

A DISSERTATION SUBMITTED IN PARTIAL FULFILLMENT OF THE
REQUIREMENTS FOR THE DEGREE OF
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OF

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2020
ABSTRACT

In modern society, the amount of data being transferred is growing at an exponential rate, about 2.5 quintillion bytes data created per day. Such data can be realized in the form of include the audio, video, or text. Furthermore, due to the many advantages of the optical fiber, including the low-cost, immunity to electromagnetic interface, light weight, and high carrier frequency, it is taking over the industry. The question then arises as how to diagnose the inner fault and maintain the integrity becomes the crux for the telecommunication system based on optical fiber.

There are two main high-precision measurement methodologies to complete the task mentioned above: one is the optical time-domain reflectometry (OTDR), which has been widely applied in technologies such as light detection and ranging (LiDAR) in faraway situations. The other one is optical frequency-domain reflectometry (OFDR), which is also referred to as optical frequency modulated continuous wave (OFMCW) or chirped frequency measurement. Compared with the OTDR, the OFDR measures distance by comparing the phase change of the chirped incident optical signal and reflected signals to capture the distance information. Even though the OFDR has been proven to have many advantages over OTDR in terms of the high spatial resolution at μm level, low power consumption, and eye-friendliness, both the high complexity and cost significantly limit its applications. To overcome the above shortages in the OFDR system and expand this technology in the various applications, the following research was conducted.

Firstly, since the optical tuning source utilized in OFDR is too expensive and complex to be implemented in-situ measurements, there are some researches completed
in the tuning laser source control methods proposed to reduce the cost and complexity of this technology and fit the various applications.

Additionally, to enhance the accuracy of the OFDR measuring, some tuning rate error compensating algorithms were introduced. Benefiting from the improved spatial resolution, the physical unclonable function (PUF) in the optical fiber can be achieved as its unique identification to sustain the data transferring security.

Furthermore, even though the initial optical frequency (IOF) in OFDR has been disregarded in the distance measuring, it does influence the PUF reader precision. To fully analyze such effect, some research and IOF mitigating methods were proposed, in which a fast processing method for the PUF reader has been introduced by decreasing the calculating complexity.

From the above research, a low-cost, accurate, and effective OFDR system has been built that is suitable for various applications.
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Additionally, I am very grateful to have Prof. Ramdas Kumaresan, Prof. Daniel Roxbury, and Prof. Godi Fischer for being my dissertation committee members and Prof. Chenzhi Yuan for being my dissertation chair, and their support in correcting my dissertation. I also want to thank Prof. Haibo He, Prof. Qing Yang, Prof. Steven Kay, and Prof. Jien-Chung Lo for all their valuable advice for my course work and research projects URI.

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PREFACE

This dissertation is organized in the manuscript format consisting of six manuscripts as following:

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- **Manuscript 7:**
  Yao, Z., Mauldin, T., Xu, Z., Hefferman, G. and Wei, T., “Breaking limitations of fiber ID in traditional OFDR systems via compensation of initial optical frequency instability”, published in *Optics Letters*, 202
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INTRODUCTION

Considering that the laser source and measure are the components of OFDR, and the application demonstrates the ability and performance of the proposed system, the completed research can be organized as three categories to the goal as shown in Figure 0-1.

The following manuscripts are parts of the researching goal as:

a. Manuscripts 2, 3, 5 focus on the SCL control to achieve the low-cost optical source applied for the OFDR system, including the self-adaptive method, digital control, and double-sided design;

b. Manuscripts 1, 6 focus on the signal processing in the measurement, including the parallel algorithm and compensating error;

c. Manuscripts 1, 4, 7 focus on the applications of the OFDR, including the distributed temperature sensing, fiber identification, and compact system design.

Figure 0-1 The categories for the research works
MANUSCRIPT

Chapter 1 Real-time signal processing for sub-THz range grating-based distributed fiber sensing

by

Zheyi Yao, Tao Wei, Gerald Hefferman, and Kan Ren

published in

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Abstract

Distributed optical fiber sensors are an increasingly utilized method of gathering distributed strain and temperature data. However, the large amount of data they generate present a challenge that limits their use in real-time, in-situ applications. This article describes a parallel and pipelined computing architecture that accelerates the signal-processing speed of sub-terahertz fiber sensor (sub-THz-fs) arrays, maintaining high spatial resolution while allowing for expanded use for real-time sensing and control applications. The computing architecture described was successfully implemented in a field programmable gate array (FPGA) chip. The signal processing for the entire array takes only 12 system clock cycles. In addition, this design removes the necessity of storing any raw or intermediate data.

Introduction

Sub-terahertz-range fiber structure (sub-THz-fs) technology is a high spatial resolution (mm level) distributed fiber optic sensing technology that has been applied to both strain and temperature sensing applications [1, 2]. Uniquely, sub-THz-fs sensors obviate the necessity of a broadband swept laser source, such as an external cavity tunable laser (ECL), which are required by alternative coherent optical frequency domain reflectometry (C-OFDR)-based distributed sensing technologies [3]. It has been successfully demonstrated that a digitally-controlled swept laser source, built around a low-cost semiconductor laser, such as a distributed feedback laser (DFB) or a vertical-cavity surface-emitting laser (VCSEL) is capable of interrogating a sub-THz-fs array for high spatial resolution distributed temperature sensing applications [4, 5]. The all-electronic design of this swept laser source allows each sweep to be completed within
20 ms or less, enabling a maximum update rate of the entire sensing array above 50 Hz [4].

While this increased sweeping rate is a step forward towards real-time, in-situ monitoring, in previous work, 400ms was required to process the raw data and compute the distributed temperature profile for only 10 equivalent sensor nodes. The signal processing for each of these equivalent sensor nodes was independent, and a series of filtering and signal mixing algorithms were implemented using a computer to achieve the sensing result for each equivalent sensor. These computational tasks for each equivalent sensor are scheduled in a sequential fashion; thus, the total processing time is proportional to the total number of equivalent sensor nodes along the fiber sensor. Larger sensing networks therefore require a longer processing time using this method, preventing us from capitalizing on the full potential of the update rate, which ideally should only be limited by the physics – sweep velocity of the laser.

Sub-THz-fs based distributed sensing technology is particularly attractive for many real-time feedback control applications, such as unmanned vehicles and robotics, thanks to its low power, low cost, and more integrated interrogator [2]. However, the low processing speed imposes more delay between the sensor and the controller, which is hazardous in all feedback control systems. In certain applications, such as capturing the dynamic strain profile along a mechanical structure over time, where real-time monitoring is not required, one can store the raw data in memory chips, and process them at a later time. However, this post-processing approach requires large memory to store all the raw data, and high-bandwidth channels to transfer large amount of data from signal acquisition device to memory chips, leading to higher system cost and more
power consumption. Ideally, memory device is only necessary to store the end results, rather than raw data or intermediate data.

This work proposes a method to tackle the challenge via a parallel and pipelined computing architecture. Since the signal processing for each equivalent sensor in a sub-THz-fs array is independent, a fully parallel processing structure is introduced to simultaneously process all equivalent sensors. Furthermore, the unique processing algorithm in sub-THz-fs technology allows us to design a fully pipelined structure for each equivalent sensor. As a result, unlike post-processing approach or any conventional computer-based sequential processing approaches, this method obviates the need to store any raw or intermediate data. The proposed architecture was successfully implemented in a field programmable gate array (FPGA) chip. The signal processing for the entire array with 10 equivalent sensors takes 12 clock cycles. Given a system clock of 1MHz, the processing time is only 12μs.

**Operation mechanism**

![Figure 1-1](image.png)

*Figure 1-1* Schematic of the complete sensing system. (ADC, analog-to-digital converter; DAC, digital-to-analog converter; AGC, automatic gain control amplifier; PD, photodiode; MZI, Mach-Zehnder Interferometer.

The system schematic is illustrated in Figure 1-1. The entire system can be viewed in two major modules, the sweep velocity-locked laser pulse generator (SV-LLPG)
module and the sensing module. The SV-LLPG was built around an optical phase lock loop, detailed in a previous publication [4]. A DFB laser was employed as the optical source. A phase lock loop (PLL) was developed to lock the sweep velocity of the chirped laser source. The digital phase comparator and loop controller were constructed in the same FPGA chip, where the signal processing unit was implemented in, making the whole system highly integrated. The output frequency of the SV-LLPG during a sweep can be expressed as:

\[ f_L = f_0 + vt \]  

(Eq 1-1)

where \( f_L \) is the output frequency, \( f_0 \) is starting frequency, \( v \) is the sweep velocity, and \( t \) is time. Each sweep takes 10ms with a sweep velocity of 10GHz/ms, making an entire swept bandwidth of \( \sim 100GHz \). After the completion of one sweep, 10ms relaxation time is applied for the DFB laser to return to its initial state.

The sensing module, based on a conventional C-OFDR configuration, is built to capture the raw date reflected from a sub-THz-fs array. The computing architecture implemented in the FPGA chip is utilized to process the raw date to generate the sensing results from the entire array, and send the processed data to display and/or storage devices. The homodyne configuration was constructed using two 2×2 3dB couplers, depicted in the sensing module of Figure 1-1. The input light is split into two paths via the first optical coupler, with one path serving as the reference arm and the other path directed into the sensing arm that contains a sub-THz-fs array. The sensing arm was terminated using an anti-reflection cut. Reflected light from the sub-THz-fs was then combined with light from the reference arm through the second coupler. A balanced photodetector and a single channel ac-coupled 16-bit ADC was used to record the
combined signals. The sampling rate of the ADC was set to 1MSa/s. The digitized raw data were then fed into the FPGA chip.

The device under test (DUT) is a 20pt periodic weak reflection sub-THz-fs array with a 1mm pitch length, fabricated along a single-mode fiber (SMF-28, Corning, Inc.) using a Ti: sapphire femtosecond laser (Coherent, Inc.) [2]. The total ac-coupled voltage received by the ADC can be expressed as:

\[ v_t(t) = A \sum_{n=0}^{N-1} \cos[2\pi v(\tau_0 + n\Delta\tau)t + 2\pi f_0(\tau_0 + n\Delta\tau)] \]

(Eq 1-2)

where \( v_t(t) \) is the ADC captured waveform, \( A \) is the magnitude of each individual reflector in a sub-THz-fs array (ideally, all the reflectors should have the same reflection coefficient), \( n \) denotes the nth reflector in the array, \( N \) is the total number of reflectors, \( \tau_0 \) is the time delay difference between the reference arm and the optical path associated with the first reflector in the sensing arm, and \( \Delta\tau \) is optical time delay between neighboring reflectors. In our experiment, the sub-THz-fs array is considered as 10 cascaded sub-THz-grating sensor nodes, differentiated using a 4mm wide moving Butterworth bandpass filter (BPF) with a step size of 2mm. In this case, each sensor unit contains roughly 4 reflection peaks. The sensing mechanism is based on tracking the frequency shift of the equivalent sub-THz-grating signal. To extract the sub-THz-grating signal, a self-mixing technique is applied to the filtered waveform, corresponding to each individual grating. The filtered waveform is squared, and low pass filtered to obtain the down-converted sub-THz-fs signal. This signal processing method has been systematically investigated in the previous publication [2]. The ambient changes lead to optical path length changes between weak reflectors in a sub-
THz-fs array. Thus, a phase shift in the resulting interferogram can be used to measure the changes, such as temperature and strain, along the sensing probe.

Figure 1-2 depicts the schematic of the proposed computing architecture. The raw data, i.e. the waveform captured by the ADC, is fed into a FIFO in order to synchronize the data flow with the FPGA master clock. Then, the data stream is simultaneously distributed to N computing channels. Each channel is dedicated to process a specific equivalent sensing element – a sub-THz-grating. All the channels are independent to each other, and configured in a parallel fashion. Thus, all the equivalent sensors are processed simultaneously by this design. Within an individual channel, all the abovementioned signal processing algorithms are pipelined to ensure zero latency in the data flow, leading to a maximum processing speed. Each pipelined channel includes 4 computing modules.

The first processing module is a high order Butterworth BPF, which is responsible to select the equivalent sensing element from the sub-THz-fs array, as discussed in the
previous session. The system function of nth order bandpass filter, \( H(z) \), can be expressed as an infinite impulse response (IIR) filter:

\[
H(z) = \frac{b_n z^{-n} + b_{n-1} z^{-(n-1)} + \cdots + b_1 z^{-1} + b_0}{a_n z^{-n} + a_{n-1} z^{-(n-1)} + \cdots + a_1 z^{-1} + a_0}
\]

(Eq 1-3)

In this design, a 7th order BPF was employed to select the equivalent sensors. Accordingly, the output signal \( y(n) \) of bandpass filter can be expressed as:

\[
y(n) = \frac{1}{a_0} \left\{ [b_6 x(n - 6) + \cdots + b_0 x(n)] - [a_6 y(n - 6) + \cdots + a_1 y(n - 1)] \right\}
\]

(Eq 1-4)

Following the BPF module, a self-mixing module is constructed. This module utilizes the DSP slice resource in the FPGA chip to perform a square function to every data input in the pipelined data stream. The typical output of a self-mixing module is shown in Figure 1-3(a), where, apparently, high frequency components are included. Thus, low pass filter (LPF) is followed to strip out the high frequency components. The low frequency components from one sub-THz-grating can be expressed as:

\[
v_m = A \sum_{m=0}^{M-1} (M - m) \cos m(2\pi n \Delta t + 2\pi f_0 \Delta t)
\]
where $v_m$ is the LPF filtered reflection spectrum of the $m^{th}$ sub-THz-grating in the array, $M$ is the total number of reflectors in a grating structure. The typical output, i.e. the processed reflection spectrum of an equivalent sensor node, after a LPF module is shown in Figure 1-3(b). The self-mixing module takes 1 clock cycle to compute, and

![Figure 1-4](image)

*Figure 1-4* Reflection spectrum shift in response to external perturbation.

the LPF module takes 3 clock cycles to compute along the pipelined structure.

According to the sensing principle of sub-THz-gratings, assuming the phase change ($\Delta \tau$) within a sub-THz-grating is uniform, the change of target sensing quantity, strain or temperature, is linearly proportional to $\Delta \tau$. According to (Eq 1-5), the frequency shift of the sub-THz-grating in reflection spectrum, illustrated in Figure 1-4. Since there only exists one peak in the processed reflection spectrum of every sub-THz-grating, the location change of the maximum value point in the processed reflection spectrum server as an indicator of the relative change of the target sensing quantity. Thus, a peak position detector is designed and implemented in the pipelined signal processing structure to output the location of the maximum point from each sweep. The peak position detector takes 1 clock cycle to compute.
The peak locations of all sensors – all equivalent sensing nodes in the array – are obtained simultaneously, and uploaded through AXI communication link to an ARM core (Xilinx Zynq programming system, PS), and eventually displayed.

The IIR filter was constructed using fixed-point binary numbers. The digital structure of the IIR filter is illustrated in Figure 1-2. A 20-bit width data was used to ensure sufficient data resolution. The IIR filter coefficients are designed in MATLAB. The impulse response of two BPFs were shown in Figure 1-5, where the Fourier transform of the raw data, containing signals from all reflectors in the array, was plotted in the same figure. The Fourier transform of the raw data was also plotted in Figure 1-5.

The x-axes of Figure 1-5 was converted to optical time delay. It clearly shows that the designed BPF is capable to isolate the signal from each individual sub-THz-grating from the array. The processing time of the BPF module is 7 clock cycles.

**Experiments, results and discussion**

In order to investigate the proposed computing architecture and demonstrate its high-speed real-time signal processing capability, an experiment was conducted to measure the dynamic change of temperature along the sensing array. By quickly moving
a hot soldering iron tip along the fiber, a dynamic temperature distribution is generated. The soldering iron tip was set at ~1 cm away from the fiber, and the horizontal moving speed is set at ~200 cm/sec. Both ends of the device under test (DUT), i.e. 20 mm sub-THz-fs array, were fixed to an optical table. The lead-in fiber of the device under test (DUT) was secured to the sensing system as shown in Figure 1-7. 

Figure 1-7 plots the relative peak position shifts of 10 adjacent equivalent sensor nodes in response to dynamic temperature distribution along the DUT over a duration of 70 ms. The proposed computing architecture enables a real-time update rate of 50 Hz. Given the ADC sampling rate of 1 MSa/s and the laser sweep velocity of 10
GHz/ms, 1 pt in peak position shift is corresponding to 10 MHz. Thus, the maximum peak position shift is around 14 GHz, corresponding to a maximum temperature change of around 9.3 μs, as the temperature sensitivity of 1 mm was measured to be around -1.5 GHz/μs. And the negative frequency shift on Figure 1-6 are caused by the suppressing as the DUT is fixed on optical table.

**Conclusion**

To summarize, a parallel and pipelined computing architecture was proposed to achieve real-time signal processing for sub-THz-fs distributed sensing applications. The proposed architecture was successfully implemented in a FPGA chip. The signal processing for the entire array with 10 equivalent sensors takes only 12 system clock cycles. The proposed computing architecture maintains high spatial resolution, while provides high temperature resolution, allowing for expanded use for real-time sensing and control applications. Importantly, this design removes the necessity of storing any raw or intermediate data.
Chapter 2 Self-adaptive method for performance enhancement of sweep-velocity-locked lasers

by

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Abstract

Optical frequency domain reflectometry (OFDR) and frequency modulated continuous wave (FMCW) based sensing technologies, such as LiDAR and distributed fiber sensors, fundamentally rely on the performance of frequency-swept laser sources. Specifically, frequency-sweep linearity, which determines the level of measurement distortion, is of paramount importance. Sweep-velocity-locked semiconductor lasers (SVLL) controlled via phase-locked loops (PLLs) have been studied for many FMCW applications owing to its simplicity, low-cost, and low power consumption. In this letter, we demonstrate an alternative, self-adaptive laser control system that generates an optimized predistortion curve through PLL iterations. The described self-adaptive algorithm was successfully implemented in a digital circuit. The results show that the phase error of the SVLL improved by around one order of magnitude relative to the one without using this method, demonstrating that this self-adaptive algorithm is a viable method of linearizing the output of frequency-swept laser sources.

Introduction

Highly linear frequency-swept laser has been attracting increasing attentions for high spatial resolution, sensitivity, low-power, and eye-friendly, which has been applied in spectroscopy, frequency modulated continuous wave (FMCW), LiDAR and distributed fiber sensors [3, 4, 6-12]. Compared with the external cavity tunable laser (ECL) [3], sweep-velocity-locked semiconductor lasers (SVLLs) controlled via phase-locked-loops (PLLs) are becoming more popular for the simple design, low cost, low power consumption, and small footprint. In a typical PLL-based SVLL, a frequency/phase discriminator, typically an optical interferometer, converts optical
sweep velocity into a radio frequency signal, where this frequency is proportional with the sweep velocity [3, 10, 11]. A phase error detector is used to compare this radio frequency signal with a high-precision reference signal. The phase error is then sent to a loop controller to generate a control signal, which feeds back into the laser driver to complete the loop. The optical frequency output of semiconductor lasers has a highly non-linear relation with the injection current. Thus, to achieve required sweep-velocity locking, a pre-distortion curve is often used to cause the laser to output a quasi-linear sweep, and the control signal is added onto this curve to achieve a high-quality locking. The pre-distortion curve can be a predetermined voltage ramp signal, or it can be derived from the frequency response of a ramp signal [4] [6-10]. However, the traditional methods can only generate one constant sweep velocity and require the complex equipment to measure the laser source quality, which limit the applications of SVLLs [13]. In this letter, a self-adaptive method is capable of finding an optimized predistortion curve that closely approximates a locked sweep. To measure the performance of our design, we directly utilize the control signals generated in the control system without any extra complex instrument. As for the application, the laser source in this system can be easily implemented in the standard design without adding any additional components. In addition, the optimized predistortion curve can be adapted within two iterations to generate different frequency velocities.

**Operation Mechanism**

A phase locked loop (PLL) is a feedback system that generates a output signal whose frequency and phase are tracking the input ones, the simplest structure is shown in *Figure 2-1*, including phase detector, loop control and voltage controlled oscillator.
(VCO) [14]. And the schematic of SVLL we applied in this letter is illustrated in Figure 2-2. Compared with the traditional PLL, our SVLL contains two parts: the FPGA working as the phase comparator (PCMP) and loop control (LC) and other additional functions, the components from digital-to-analog converter (DAC) to analog-to-digital converter (ADC) working as the VCO. In other words, we're locking the frequency of ADC signal rather than laser itself.

In this design, the FPGA works as the core control unit and is constructed using three digital subsystems: a part of PLL system (phase comparator and loop control), self-adaptive predistortion generation system, and performance evaluation system. Figure 2-2 shows the functional blocks of all three digital subsystems.
First of all, the part of PLL system includes the Schmitt trigger (ST), phase comparator (PCMP), reference clock (REF), and loop control (LC) modules. The ST receives the digitized signal from the ADC, which converts the balanced photodetector (BPD) analog signal into a single-bit digital one (Sig in Figure 2-3). The REF generates a precise reference clock (Ref in Figure 2-3 with period $T$, frequency $f_R$). And the phase error between Sig and Ref is computed in PCMP. In order to suppress the high frequency noise in digital system, we apply XOR (or Type I) phase detector. After that, the phase error (XOR output in Figure 2-3) works as the input of the LC, and implies the performance of the laser source.

![Figure 2-3 XOR phase detector waveforms.](image1)

![Figure 2-4 The relation between Trig and laser current in time domain with different units. ($t_L$, the control period).](image2)
Figure 2-5 Self-adaptive approach for predistortion curve training. (a) hardware architecture; (b) finite state machine (FSM) of PRE-CTRL module that enables the self-adaptive approach (IDLE: idle state; RA_WB: read BRAM A and write BRAM B state, WA_RB: write BRAM A and read BRAM B state; CNT: counter).

Secondly, the self-adaptive predistortion generator contains predistortion control (PRE-CTRL), BRAM and output (OUT) modules. The Figure 2-5 shows the simplified architecture and finite state machine (FSM) of the proposed self-adaptive method for predistortion curve training. The working mechanism of this method can be thoroughly explained using Figure 2-5. At the beginning of predistortion curve training, the data stored in BRAM A and B, shown in Figure 2-5(a), are empty, and the process falls in the idle state (IDLE) in Figure 2-5(b). It is worth noting that throughout the entire PLL SVLL system, a trigger signal (Trig) is generated accompanying the laser sweep. The relation between the Trig and laser current curve is shown in Figure 2-4 where $t_L$ is the control period. During the sweep, the Trig equals to 1, and Trig pulls to 0 during relaxation stage, where the laser injection current is zero. In other words, the Trig works as a switch for laser. The FSM described in Figure 2-5(b) is triggered by the rising edge of Trig. When the first Trig signal comes in, the process falls into the RA_WB state, in which, PRE-CTRL reads the BRAM A, sends it over to OUT module to combine with the control signal from LC module to form the DAC output signal. Since the BRAM A is empty at the beginning stage, the DAC output signal is the same as the control signal from LC. Also, in this current RA_WB stage, PRE-CTRL module takes the DAC output
signal and writes it into the BRAM B. As the second Trig comes in, the process falls into WA_RB state, in which, PRE-CTRL reads the BRAM B, sends it over to OUT as a predistortion curve. The OUT module takes in the predistortion curve, combines it with the control signal from LC, and outputs to the DAC. Meanwhile, this DAC output signal is taken by the PRE-CTRL and stored in the BRAM A as the updated predistortion curve. This process goes on to switch between WA_RB and RA_WB. As a result, the predistortion curve is trained to form a high-quality quasi-linear sweep, which best approximates a velocity locked sweep.

Besides, the performance evaluation system is applied to judge the performance of locking, which includes phase error recording (PER), direct memory access (DMA), and double data rate synchronous dynamic random-access memory (DDR) modules. The phase error captured by PCMP is measured in PER, then sent to DMA, finally stored in DDR. In our design, we save the phase error every half Ref period at the falling and rising edges or Ref, or the sampling rate for the phase error is 2/T. For a precise locking with XOR, the phase error would be constant as $\pi/2$ [14].

As described before, we combine the control signal and predistortion curve generated by PRE-CTL to synthesize a digital output signal via OUT, and transfers the synthesized output signal to the laser driver through a digital-to-analog converter (DAC) chip [5]. The self-adaptive predistortion training system, which is the main contribution of this letter, collects the control signal from the LC module in the PLL system and updates the predistortion curve until the self-trained predistortion curve is able to produce a quasi-linear sweep.
Apart from the FPGA modules and DDR, the VCO of SVLL built by DAC, LD, SCL, CPL, MZI, BPD, and ADC plays a significant role in our design as well. The LD covert the DAC output signal to a current source to drive the SCL. There are low-pass filter and current converter in LD. Then the SCL output signal or laser original signal would be split into two branches using a 1/99 coupler. The majority (Laser output in Figure 2-2) of the laser light with frequency $f_0$, can be integrated with an external heterodyne system for various linear frequency modulation type applications. 1% of laser light is used for the purpose of sweep control. The smaller laser light passes through an MZI, in which the time difference between two arms is $\tau$, and is collected by the power-voltage converter, BPD. The combined MZI and BPD configuration converts the swept laser light into a radio frequency (RF) signal with frequency $f_B$, where the frequency of the RF signal is linearly proportional with the sweep velocity of the laser source1, which can be expressed as:

$$ f_B = \frac{df_0}{dt} \tau $$

(Eq 2-1) shows the sweep velocity of SCL is constant if $f_B$ is invariant.

**Experiments, results and discussion**

In order to demonstrate the proposed self-adaptive method and evaluate the performance of this approach, the abovementioned structure was implemented in an FPGA. In this experiment, the sampling rate for both ADC and DAC is 125 MSPS, while the resolution for both ADC and DAC is 14 bits, the master clock of all modules is 125 MHz, the $\tau$ of the MZI is 10 ns, $f_R$ is set 104.17 kHz, or 1 over 12,000 of master clock, the sweep duration is set at 10 ms, corresponding to an optical sweep velocity of
10.42 GHz/ms, and a sweep bandwidth of 104.2 GHz. Figure 2-6 shows the experimental results of the self-adaptive method described in Figure 2-5. A relaxation time of 10 ms is taken after each sweep, between the system launches the next sweep. In other words, \( t_L \) is 20 ms. Only 2 iterations, or 40 ms, were required to train the predistortion curve to a sufficiently good condition. After the first iteration, the predistortion stored in BRAM A generates a sweep, plotted in Figure 2-6 (a), which shows the short-time Fast Fourier transform (STFFT) of the RF signal, collected by the ADC over the laser sweep duration of 10 ms. Figure 2-6 (b) shows the sweep after the second self-adaptive iteration. It is worth noting that all the data taken in Figure 2-6 are purely generated from predistortion curve when active feedback loop was turned off. Figure 2-6 (c) shows the time domain phase error between Sig (generated by RF signal in ST) with 104.17 kHz Ref at the predistortion curve after the second iteration, taken by PER.

The high-quality predistortion curve trained by the proposed self-adaptive method is expected to significantly improve the performance of SVLL system. To demonstrate the improvement, a comparison experiment was conducted, plotted in Figure 2-7.
Figure 2-7 Comparison experiments. (a) STFFT under active sweep velocity locking with a ramp-like predistortion curve, (b) STFFT under active sweep velocity locking with a trained predistortion curve by the self-adaptive approach, (c) The FFT comparison between ramp-like distortion based locking and self-adaptive trained predistortion based locking, (d) phase error of (a), and (e) phase error of (b), (f) The power spectral density (PSD) comparison between ramp-like distortion based locking and self-adaptively trained predistortion based locking.

Figure 2-7 (a) and (d) plot the STFFT of the RF signal captured by the ADC and its corresponding phase error taken by the PER module with a ramp-like predistortion curve, while Figure 2-7(b) and (e) plot the STFFT of the RF signal captured by the ADC and its corresponding phase error taken by the PER module with a trained predistortion curve using the proposed self-adaptive method. It clearly shows that the SVLL system with self-adaptively trained predistortion curve outperformed the one with ramp-like predistortion. Figure 2-7 (c) shows the comparison of Fast Fourier Transform results of Ramp-like and Self-adaptively trained predistortions. We set the max results of FFT as 0 dB, the values of nearby noise are clearly depressed about 5 dB. Figure 2-7(f) compares the phase spectral density (PSD) of the phase error in both cases, there is about 1 order difference between them at low frequency.
The proposed self-adaptive method is highly versatile and can be easily adopted to some frequencies by setting the REF parameter, which only can be set even number in FPGA. To demonstrate this unique advantage, the experiment was modified to lock at a sweep velocity of 20.85 GHz/ms by changing the reference clock to 208.33 kHz (1 over 6,000 of master clock), $t_L$ to 10 ms, sweep time to 4 ms. And the results are plotted in Figure 2-8.

**Conclusion**

To conclude, the self-adaptive method in this letter is capable of finding an optimized predistortion curve that closely approximates a locked sweep; it directly utilizes the control signal without complex computations; it can be easily implemented in the standard digital design without adding any additional components; the optimized predistortion curve can be self-adaptive in real-time within two interactions, i.e. two sweeps, for the SCL at different velocities. In addition, as the frequency-swept laser is
the core for FMCW, OFDR, and LiDar, the self-adaptive method can be easily integrated in various applications for its low cost and universal digital design.
Chapter 3 Digitally integrated self-trained pre-distortion curve finder for passive sweep linearization of semiconductor lasers

by

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Abstract

This letter reports a digitally integrated, self-trained pre-distortion curve finder for linearly frequency-swept semiconductor lasers. In this method, we introduce an iteration machine to train the laser current curves by utilizing the feedback signals from the laser’s output. By measuring the phase and frequency information of lasers, we find that the pre-distortion finder in our design can generate various high sweep velocities (THz/ms) of lasers in 1 second with phase error range less than $\pi/2$. This method is universal for semiconductor lasers at any sweep frequency velocity.

Introduction

Highly linear frequency-swept lasers have been attracting increasing attention due to their ability to make sensitive measurements with high spatial resolution using low power, eye-friendly semiconductor laser diodes. Present applications for these frequency-swept lasers include tunable laser diode spectroscopy [15], as known as frequency-modulated (FM) spectroscopy, optical frequency-modulated continuous-wave (FMCW) LiDAR [16], optical frequency domain reflectometer (OFDR)-based distributed optical fiber sensing [3, 7, 17-20] and optical fiber key technology for identification and physical security (FiberID), one of the physical unclonable function (PUF) technologies [21]. Each of these applications requires the use of lasers capable of generating a range of well-controlled output frequencies in order to make high-fidelity frequency domain measurements. Various laser frequency control techniques have been investigated in order to generate highly linear, well controlled frequency-swept outputs [12]. There are two basic categories of control systems that linearize the frequency sweep of the laser, which can be broadly characterized as either active or
passive. Methods of active laser control utilize closed feedback loops, specifically optical phase-locked loops (OPLLs), to capture the frequency or phase error of the laser frequency sweep velocity. The error signal is then used to modify the drive current of the laser to linearize the sweep velocity of the frequency-swept laser source \[9\]. Alternatively, methods of passive control utilize a pre-determined, ramp-like injection current to drive the laser output \[22\]. Both active and passive methods fundamentally rely on a specifically tuned pre-distortion curve of their initial injection current in order to generate a constant sweep velocity and resulting linear frequency output. In the absence of a well-tuned initial pre-distortion curve, neither active or passive laser control methods are effective; thus, determining the parameters of a well-tuned pre-distortion curve is of central importance in electronically driven swept frequency lasers.

However, despite this critical importance, there have been no published systematic investigations into methods of generating well-tuned pre-distortion curves. An ideal version of such a method would have several requirements: the method should be applicable to a wide range of (ideally all) sweep velocities and semiconductor laser types; the method should require no prior knowledge of the frequency response of the laser; the generated sweep velocity, and resulting laser output, should be highly accurate; the calculation process should be as efficient as possible; and the required system should be of minimal complexity.

In this paper, we introduce a digitally integrated, self-trained pre-distortion curve generation method for semiconductor lasers (SCLs) to generate linear frequency-swept optical signals that meets the above requirements. In order determine the method’s feasibility and to further explore its unique engineering advantages, the method was
implemented using an electronic-photonic system illustrated in Figure 1. Systematic experimental investigation was then conducted at various sweep velocities. The detailed results of this investigation are described below. Overall, the method as implemented takes less than 1 second to generate well-tuned pre-distortion curves for SCLs.

Theory and Implementation

![System schematic](attachment:image.png)

*Figure 3-1* System schematic: DFB-SCL, distributed feedback semiconductor laser; CPL, coupler; MZI, Mach-Zehnder interferometer; BPD, balanced photodetector; ADC, analog-to-digital converter; DST, digital Schmitt trigger; TDC, time digital converter; DCL, digital loop control; DAC digital-to-analog converter.

A schematic of the method is shown in *Figure 3-1*. A distributed feedback (DBF) SCL serves as the laser source and is modulated using a time-varying current curve controlled by a digital control loop (DCL). A coupler (CPL) directs 2% of the laser power to a Mach-Zehnder interferometer (MZI); the resulting microwave frequency output of the MZI is then used to determine the sweep velocity of the laser output. The remaining 98% of the laser power can be applied outside of the loop as a linear frequency-swept laser source. As described in prior work [12], a small MZI delay, $\tau_d$, leads to a flat wideband frequency and phase response. Under the assumption that the SCL is operated with an ideal constant sweep velocity $v$, then the AC-coupled voltage output of the balanced photodetector (BPD) can be expressed as:
\[
 u(t) = \frac{A(t)^2}{8}\eta \cos[2\pi(f_0 + vt)t_d]
\]  

(Eq 3-1)

where \(A(t)\) is the amplitude of the electric field directed into the MZI as a function of time, \(\eta\) is the light-to-voltage conversion coefficient of the BPD, \(f_0\) is the starting frequency of the SCL laser sweep, \(v\) is the optical frequency sweep velocity, and \(t\) is time. Then the frequency we captured in the BPD signal can be expressed as:

\[
 f_{BPD} = v t_d
\]  

(Eq 3-2)

As \(A(t)\) changes over the laser driver current, \(I_{SCL}(t)\), we have the amplitude, \(A_{BPD}(t)\), of the BPD signal given as:

\[
 A_{BPD}(t) = \gamma(I_{SCL}(t) - I_{th})
\]  

(Eq 3-3)

where \(I_{SCL}(t)\) is the current in the SCL, \(I_{th}\) is the invariant threshold current of the SCL, and \(\gamma\) is the constant conversion coefficient in V/A.

The BPD signal, \(u(t)\), is then converted to a digital signal in the analog-to-digital converter (ADC) with the output \(S_{ADC}[n]\) expressed as:

\[
 S_{ADC}[n] = u(t)|_{t=nT_s}
\]  

(Eq 3-4)

where \(T_s = \frac{1}{F_s}\), \(F_s\) is the ADC sampling rate, and \(n\) is the sample number. We should apply the digital auto-gain current circuit to measure the \(f_{BPD}\), which is integrated in the digital Schmitt trigger (DST).

In the DST, the thresholds of the Schmitt trigger are set as:
where \( A_{BPD}[n] = A_{BPD}(t)|_{t=nT_s} \) and \( \alpha \) is the fixed phase shift for the DST. The minimal \( \alpha \) value that is required depends on the signal-to-noise ratio of the BPD signal; however, the DST would be insensitive to the \( f_{BPD} \) with increasing \( \alpha \). Thus, the choice of \( \alpha \) is a trade-off between frequency accuracy and high-frequency noise. The output of the DST can be described as:

\[
S_{DST}[n] = \begin{cases} 
0 & S_{DST}[n - 1] = 1, S_{ADC}[n] < -T_{ST}[n] \\
1 & S_{DST}[n - 1] = 0, S_{ADC}[n] > T_{ST}[n] \\
S_{DST}[n - 1] & \text{else}
\end{cases}
\]

(Eq 3-6)

where \( S_{DST}[n] \) is the DST output signal over sampling number \( n \), and \( S_{DST}[0] = 0 \). By applying the DST in our design, the digital signal \( S_{ADC}[n] \) is converted into a single bit signal \( S_{DST}[n] \). After \( S_{DST}[n] \) is generated, it is sent to the time digital converter (TDC) to determine the frequency errors (UP, DN), which are captured by comparing the period over the frequency parameter, \( p_f \), as:

\[
C[n] = \begin{cases} 
1 & S_{DST}[n]! = S_{DST}[n - 1] \\
C[n] + 1 & \text{else}
\end{cases}
\]

(Eq 3-7)

\[
UP[n] = \begin{cases} 
1 & C[n] > p_f \\
0 & \text{else}
\end{cases}
\]

(Eq 3-8)

*Figure 3-2 TDC timings at \( p_f = 4 \).*
\[ DN[n] = \begin{cases} 
(p_f - C[n - 1]) & C[n - 1] < p_f, S_{DST}[n] \neq s_{DST}[n - 1] \\
0 & \text{else} 
\end{cases} \]

(Eq 3-9)

where \( C[n] \) is the count over sampling number \( n \). Figure 3-2 shows the relations between \( S_{DST}[n], \) UP, and DN when \( p_f = 4 \). Compared with the OPLL design described in prior work [12, 23], the reference signal is no longer required and the reference frequency can be treated as:

\[ f_{REF} = \frac{F_s}{2p_f} \]

(Eq 3-10)

where \( F_s \) is both the ADC sampling rate and the digital system processing frequency.

There exist two traditional phase comparator types: type I (XOR, bitwise operation) and type II (PFD, phase frequency detector [24]); however, neither is suitable for this iterative design. In the case of the type I, if we apply this to the iterative design, the control loop is only sensitive for a single side, which means the frequency will increase over time. In the case of the type II, the input and reference signals are applied as clocks to capture the rising edge, which would cause a spur and ignore the falling edge information. The TDC implemented in our design overcomes both of these shortfalls.

Given the UP and DN signals from the TDC, the digital loop control (DLC) converts the frequency error information and sends the information to the digital-to-analog converter (DAC). The structure of the DLC is shown in Figure 3-3.

The digital integrator (DINT) function can be expressed as:

\[ DINT[n] = C_{INT} \sum_{i=0}^{n} UP[i] - DN[i] \]

(Eq 3-11)
where \( C_{\text{INT}} \) is the coefficient of DINT. As described previously, the control system introduces analog delay and digital latency caused by the SCL, CPL, MZI, and digital system. If we quantify them as \( d \) control clock cycles, then \( DINT[n] \) should correspond with the signal of the reading BRAM value, \( RB[n – d] \). Thus, the digital delay part (DLY in Figure 3-3) is introduced in order to compensate for the delay and latency. The writing BRAM values can then be expressed as:

\[
WB[n] = DINT[n] + RB[n]z^{-d}
\]

(Eq 3-12)

where \( z^{-d} \) is the delay compensation. Initial values of the RB are equal to the constant voltage corresponding with the SCL threshold current. In our design, we treat the exchange of WB and RB as one complete iteration. If the time of iteration reaches the setting value, \( N_i \), the exchange of WB and RB will stop, the output only depends on the RB values, and the current curve of the laser is not affected by the BPD signals. In other words, the phase error captured in the DINT would feedback to the WB for the following laser control period. Finally, by applying this method iteratively, we can capture the perfect pre-distortion curve in the RB, which remains the same after the iteration time reaches \( N_i \).

**Experiment**
In order to demonstrate the proposed self-trained pre-distortion curve generation method and evaluate the performance of this approach, the digital system structure described above was implemented using a field-programmable gate array (FPGA). The following hardware and system parameters were used during experimentation: SCL, Alcatel 1905 LMI, 1550 nm, current threshold 75 mA; MZI delay, \( \tau_d = 10\) ns; BPD BW 200 MHz; ADC sampling rate, \( F_s = 125 \) MHz; DLY compensation, \( d = 57 \); DAC sampling rate and system processing frequency equal to the ADC sampling rate.

To investigate the influence of various \( \alpha \) in the DST, experimental trails were conducted at \( \alpha = 0^\circ, 5^\circ, 10^\circ, \) and \( 30^\circ \) as illustrated in Figure 3-4. The other parameters remained unchanged: the frequency parameter, \( p_f \), was 5; the locking frequency, \( f_{\text{REF}} \), was 12.5 MHz; and the sweep velocity, \( v \), was 1.25THz/ms. The results are shown in Figure 3-4. The locking period was 25\( \mu \)s and each iteration required 1 ms to complete.
Over successive iterations, the current curves change except for the $\alpha = 0^\circ$ case, which is caused by the high frequency noise in the laser signal. Comparing Figure 3-4(b)-(d), it is clear that a large $\alpha$ is useful in suppressing the noise of the system. In addition to generating the current curves, the phase error information (UP, DN) given by (Eq 3-8)/(Eq 3-9) can be applied to quantify the performance of the curves. The absolute phase error results with different $\alpha$ values are shown in Figure 3-5. With the exception of the $\alpha = 0^\circ$ case, the phase errors decrease with successive iterations, reaching less than $\pi/2$ after 1000 iterations. In the idealized case of a perfectly tuned laser drive current, the phase error is a constant value, and is the consequence of a perfectly invariant frequency sweep velocity of the laser. As for the influence of $\alpha$, when considering the 500 iteration results (yellow results in Figure 3-5(b)-(d)), it is demonstrated that phase error decreases as $\alpha$ increases. As mentioned previously, it

![Figure 3-5](image)

*Figure 3-5* Phase errors with iterations at different DST parameters. (a) $\alpha = 0^\circ$; (b) $\alpha = 5^\circ$; (c) $\alpha = 10^\circ$; (d) $\alpha = 30^\circ$. 

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Figure 3-7 Short-time FFT results with different iterating curves, $p_f = 5$, $\alpha = 5^\circ$. (a) Initial current value result; (b) 300 iterating current curve result; (c) 500 iterating current curve result; (d) 1000 iterating current curve result.

takes 1 ms to complete one iteration using this technique; thus, a minimal phase error

Figure 3-6 Results for 0.89THz/ms frequency-swept velocity with $p_f = 7$, $\alpha = 5^\circ$. (a) Iterating curves; (b) Phase errors with iterations; (c) 500 iterating short-time FFT result; (d) 1000 iterating short-time FFT result.
range of less than $\pi/2$, and the resulting well-tuned pre-distortion curve, can be generated in 1 second.

*Figure 3-7* shows the short-time fast Fourier Transform (STFFT) results at different iterations when $\alpha = 5^\circ$, corresponding with the current curves shown in *Figure 3-4* (b).

Initially, the output of the MZI varies substantially, corresponding to a widely varying laser sweep velocity. With successive iterations, the STFFT demonstrates that the MZI output approaches a linear lock as the target frequency, 12.5 MHz, corresponding to a nearly linear laser sweep velocity. To exhibit the universality of our method, we implemented the experiments with frequency-swept velocities equal to 1.04THz/ms and 0.89THz/ms shown in *Figure 3-8* and *Figure 3-6*.

Several important engineering challenges remain. Chief among them is that, as this method is passive, it creates a single well-tuned curve for the specific parameters of the

*Figure 3-8* Results for 1.04THz/ms frequency-swept velocity with $p_f = 6, \alpha = 5^\circ$. (a) Iterating curves; (b) Phase errors with iterations; (c) 500 iterating short-time FFT result; (d) 1000 iterating short-time FFT result.
laser at a single point in time. Consequently, as the parameters of the laser change (e.g. temperature), a new curve will need to be generated.

**Conclusion**

Frequency-swept lasers have found important roles in a growing range of application areas. A well-tuned pre-distortion curve is a critical element of these systems; however, generating these current input curves has not been thoroughly investigated. This work investigates a method of creating well-tuned pre-distortion curves using a novel iteration-based system. From the results we obtained we can safely conclude that the pre-distortion curve generation method proposed in this letter can generate well-tuned current curves for THz/ms frequency-swept laser sources in 1 second with less than $\pi/2$ phase error range. This design can be utilized to improve the performance of applications such as spectroscopy, FMCW LiDAR, and fiber sensing.
Chapter 4 Low-cost optical fiber physical unclonable function (PUF) reader based on a digitally-integrated semiconductor LiDAR

by

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Abstract

This paper introduces an integrated fiber PUF verification system based on a semiconductor laser source at substantially lower complexity and cost than existing alternatives. A source sub-section consisting of a linear frequency-swept semiconductor laser is used in combination with an OFDR/LiDAR-based measurement sub-section in order to conduct fiber identification via measurement of the unique Rayleigh reflection pattern of section of optical fiber. When using these Rayleigh reflection patterns as physical unclonable functions, this technique results in a maximum equal error rate (EER) of 0.15% for a 5-cm section of optical fiber and an EER of less than 1% for a 4-cm section. These results demonstrate that the system can serve as a robust method fiber identification for device and communication verification applications.

Introduction

Transmission-lines (Tx-lines) are a group of devices or structures that are used to transmit and guide energy and signals from one location to another. The energy and signals that travel in Tx-lines are in the form of travelling waves. Based on the types of travelling waves they support, Tx-lines can be broadly categorized into three groups: electrical, optical, and acoustic. Optical Tx-lines support the propagation of optical travelling waves with wavelengths spanning the ultraviolet (UV), visible, and infrared (IR) ranges. Examples of these optical transmission lines include optical fibers, optical strip waveguides on photonic integrated circuits (PICs), optical rib waveguides on PICs, segmented optical waveguides, photonic crystal waveguides, plasmonic waveguides, laser-inscribed optical waveguides, light pipes, fiber-to-chip couplers.
The characteristic impedance of any Tx-line is determined by the unique inhomogeneity if its geometry and material properties along the path of wave propagation. This unique inhomogeneity results in a varied impedance as a function of location, referred to as the impedance inhomogeneity pattern. The impedance inhomogeneity pattern of any Tx-line is unique, unpredictable, uncontrollable, and highly random; as a consequence, it can be considered to be the unique physical “fingerprint” of the Tx-line, i.e., a physical unclonable function (PUF) [25-27]. Randomly introduced to a Tx-line during the fabrication process, the impedance inhomogeneity pattern is unclonable due to its high complexity, high randomness, and high entropy.

Optical fibers are among the most ubiquitous Tx-lines in contemporary use. The impedance inhomogeneity pattern in an optical fiber is referred to as Rayleigh backscatter pattern. Like all impedance inhomogeneity patterns, the Rayleigh backscatter profile of an optical fiber is randomly inscribed during its initial manufacture, and remains a functionally permanent feature of the optical fiber throughout its lifetime. The extremely small magnitude of these Rayleigh backscatter patterns, as well as their random pattern of inscription, makes spoofing attacks by potential adversaries impractical, increasing the security of optical fiber as a PUF mechanism. The main application areas of fiber PUFs include authentication, identification (ID), anti-counterfeiting, and secured communication. Optical frequency domain reflectometry (OFDR) is an interrogation technique that lies at the foundation of many optical frequency domain applications, from frequency-modulated continuous wave (FMCW) LiDAR to distributed optical fiber sensing. OFDR also allows for the
unique Rayleigh backscatter profile of an optical fiber to be resolved as a function of length with at sub-millimeter resolution. However, the high complexity and cost associated with presently available OFDR systems have made the use of Rayleigh backscatter profiles of optical fibers as PUFs impractical.

This paper describes a digitally integrated fiber PUF reader utilizing a semiconductor laser (SCL) which surmounts these limitations, allowing for the practical use of the Rayleigh backscatter profiles of optical fibers as PUFs. A systematic experimental investigation was conducted in order to determine the performance of this design and quantify its resolution and range. The detailed results of this investigation are described below. Overall, the system implemented in our design can utilize a 4-cm section of optical fiber as a robust PUF verification system with an error rate of less than 1% at substantially less complexity and cost relative to previous systems [28].

Principle

The design utilizes two subsystems, a source sub-system and a measuring sub-system, illustrated in Figure 4-1. The source sub-system provides a highly linear

![Figure 4-1 The schematic of integrated fiber verification system. Source sub-system: DBF-SCL, distributed feedback semiconductor laser with driver; CPL0, coupler; MZI, Mach-Zehnder interferometer; BPD1, balanced photo-detector; CZ, cross-zero circuit; CL, control loop; DAC, digital-to-analog circuit; Measuring sub-system: CPL1, CPL2, couplers; CIR, optical circulator; DUT, device under test; APC, an angle physical connect fiber connector; ADC, analog-to-digital circuit.](image-url)
A distributed feedback (DFB) semiconductor laser (SCL) serves as the laser source and is modulated by a time-varying current curve generated by the control loop (CL), which is described in detail in Figure 4-2. An optical coupler (CPL) directs 1% of the laser power into a Mach-Zehnder interferometer (MZI); the resulting microwave frequency output of the MZI is then used to determine the sweep velocity of the laser output and adjust the driving current to correct for nonlinearities in this velocity. The remaining 99% of the laser power is directed to the measuring sub-system as a linear frequency-swept laser source. The small MZI delay, $\tau_d$, introduced into source sub-system using this technique results in a flat wideband frequency and phase response [12]. Given that the SCL is operating with an ideal constant sweep velocity $v$, then we have:

$$u(t) = \frac{A(t)^2}{8} \eta_1 \cos[2\pi(f_0 + vt)\tau_d]$$

(Eq 4-1)
where \( A(t) \) is the amplitude of the electric field directed into the MZI as a function of time, \( \eta_1 \) is the light-to-voltage conversion coefficient of the BPD1, \( f_0 \) is the starting frequency of the SCL sweep, \( v \) is the optical frequency sweep velocity, and \( t \) is time.

The frequency captured by the BPD1 can be expressed as:

\[
f_{BPD1} = v \tau_d
\]

(Eq 4-2)

The cross-zero circuit (CZ) measures this frequency and has an output, \( S_{CZ}(t) \), expressed as:

\[
S_{CZ}(t) = \begin{cases} 
1, & u(t) > 0 \\
0, & u(t) < 0
\end{cases}
\]

(Eq 4-3)

The analog CZ signal is then directed to the digital control system, where the discrete CZ sequence can be described as:

\[
S_{CZ}[n] = S_{CZ}(t)|_{t=nT_s}
\]

(Eq 4-4)

where \( T_s = \frac{1}{F_s} \) and \( F_s \) is the CL digital system processing frequency.

There are two control methods in the CL part: active and passive [29]. In the CL, the passive control works as the current curve finder and provides the pre-distortion (RB) for active control. The time digital converter (TDC) in the passive control determines the frequency errors (UP, DN) measured by comparing the period of the BPD output over the frequency parameter, \( p_f \), as:

\[
C[n] = \begin{cases} 
1, & S_{CZ}[n] \neq S_{CZ}[n-1] \\
C[n-1] + 1, & \text{otherwise}
\end{cases}
\]

(Eq 4-5)
\[ UP[n] = \begin{cases} 1 & C[n] > p_f \\ 0 & \text{otherwise} \end{cases} \]

(Eq 4-6)

\[ DN[n] = \begin{cases} p_f - C[n - 1] & C[n - 1] < p_f, S_{CZ}[n] \neq S_{CZ}[n - 1] \\ 0 & \text{otherwise} \end{cases} \]

(Eq 4-7)

where \( C[n] \) is the count over sampling number \( n \). Figure 4-3 shows the relations between \( S_{CZ}[n] \), UP, and DN when \( p_f = 4 \).

Compared to the optical phase locked loop (OPLL) design described in prior work [4, 13, 29], a reference signal is no longer required, and the reference frequency can be treated as:

\[ f_{REF} = \frac{F_s}{2p_f} \]

(Eq 4-8)

Given the UP and DN signals from the TDC, the digital integrator for passive control (INTp) processes the frequency error information and sends a response to the adder. The INTp results can be expressed as:

\[ INT_p[n] = C_p \sum_{i=0}^{n} UP[i] - DN[i] \]

(Eq 4-9)
where \( C_p \) is the gain coefficient of \( \text{INT}_p \). For passive control, the writing BRAM (WB) values are generated by:

\[
WB[n] = INT_p[n] + RB[n]
\]

(Eq 4-10)

The initial data of the reading BRAM (RB) is equal to a constant value corresponding to the SCL threshold current. In our design, we treat the exchange of WB and RB as one complete iteration. If the iteration time reaches the setting value, \( N_t \), the exchange of WB and RB will stop, the RB will be invariant, and the RB works as the digital-to-analog convertor (DAC) input for passive control.

As mentioned in previous literature\[30\], passive control is only used to capture the initial curve in order to drive the linear frequency-swept laser. Since laser performance is affected by its surrounding temperature, active control is indispensable to maintain the linearity of the sweep velocity.

In traditional phase locked loop (PLL) design, an XOR gate processes all phase differences as a lagging frequency, leading to the frequency increasing over time. For an ideal frequency locking result, the XOR gate would seize the result with a 50/50 duty cycle. For active control, a double-sided XOR gate (DXOR) \[29\] is used to capture the phase error information \( DXOR[n] \), which is fed back to the digital integrator for active

![Figure 4-4](image)

*Figure 4-4* The timings of the double-sided xor gate (DXOR). Ref, reference signal; CZ, cross-zero signal; DXOR, the output signal of DXOR.
control (INTa) to adjust the driving current of the laser. The timings of the DXOR is illustrated in Figure 4-4. If we assume the period of the reference clock is $T$, when ideal locking is accomplished, the duty cycle of the DXOR is $50/50$ with a period of $\frac{T}{2}$. Then the INTa results can be expressed as:

$$\text{INT}_a[n] = C_a \sum_{i=0}^{n} D_{dxor1}[i] - D_{dxor0}[i]$$  \hspace{1cm} (Eq 4-11)

$$D_{dxor1}[n] = \begin{cases} 1 & \text{if } \text{DXOR}[n] = 1 \\ 0 & \text{otherwise} \end{cases}$$  \hspace{1cm} (Eq 4-12)

$$D_{dxor0}[n] = \begin{cases} 1 & \text{if } \text{DXOR}[n] = 0 \\ 0 & \text{otherwise} \end{cases}$$  \hspace{1cm} (Eq 4-13)

where $C_a$ is the gain coefficient for the INTa. The active control output for the DAC can then be expressed as:

$$\text{DAC}[n] = \text{RB}[n] + \text{INT}_a[n]$$  \hspace{1cm} (Eq 4-14)

where $\text{RB}[n]$, generated using passive control, serves as the pre-distortion for the active control.

After establishing a highly linear frequency-swept laser source using the source sub-system, we move to the measuring sub-system. Similar to (Eq 4-1), the BPD2 output, $u_R(t)$, can be expressed as:

$$u_R(t) = \frac{A_r(t)^2}{8} \eta_2 \cos \left[ 2\pi (f_0 + vt) \frac{d}{v_p} \right]$$  \hspace{1cm} (Eq 4-15)
\[ d = 2l + L_0 \]

(Eq 4-16)

where \( A_r(t) \) is the amplitude of the electric filed directed into CPL1 as a function of time; \( \eta_2 \) is the light-to-voltage conversion coefficient of BPD2; \( f_0 \) is the starting frequency of the SCL sweep; \( v \) is the optical frequency sweep velocity; \( t \) is time; \( L_0 \) is the length differential between the circulator (CIR) arm (seen in the upper part of the measuring sub-system) and the reference arm (seen in the lower part of the measuring sub-system); \( l \) is the length from the Rayleigh scatter point in the device under test (DUT) to CIR port 2; \( d \) is the total travel distance; and \( v_p \) is the speed of light in the fiber. Thus, the reflecting frequency is \( v \frac{d}{v_p} \).

From (Eq 4-15)(Eq 4-16), we can safely conclude that the corresponding fiber array at length from \( L_{min} \) to \( L_{max} \) in the DUT can be symbolized as:

\[ s_r(t) = \sum_{i=0}^{N_r} u_{R_i}(t) \]

(Eq 4-17)

where \( N_r \) is the number of reflecting points, \( u_{R_i}(t) \) is the reflecting signal with \( d = 2l_i + L_0 \), and \( L_{min} < l_i < L_{max} \).

Since the Rayleigh scatter positions are the result of a random distribution of material property variations introduced at fabrication, they represent the unique characteristic PUF of the optical fiber. From (Eq 4-15)(Eq 4-16)(Eq 4-17), we can refer to the reflecting frequency information from \( v \frac{2L_{min} + L_0}{v_p} \) to \( v \frac{2L_{max} + L_0}{v_p} \) as the ID for the fiber array.
In practice, data acquisition and signal processing all occur in a discrete manner; therefore, we get:

\[ s_r[n] = s_r(t)|_{t=nT_s} \]  

(Eq 4-18)

where \( T_s = \frac{1}{F_s} \) and \( F_s \) is the ADC sampling rate, which is the same as the processing frequency.

If we measure two intensity arrays and transform them so that they are functions of length, yielding \( \tilde{S}_r^{(1)}(m) \) and \( \tilde{S}_r^{(2)}(m) \), which are the discrete Fourier transforms of \( s_r^{(1)}[n] \) and \( s_r^{(2)}[n] \) respectively, we can use Fourier transform identities to complete the cross correlation between the spectra of any subset of points as [3]:

\[ S_r \left( \begin{array}{c} 1 \\ \ast \\ \ast' \end{array} \right) \otimes S_r \left( \begin{array}{c} 2 \\ \ast' \ast' \end{array} \right) = \frac{1}{2\pi N'} \sum_{m=m_1}^{m_2} \tilde{S}_r^{(1)}(m) \tilde{S}_r^{(2)*}(m) \exp \left( ijm \frac{2\pi}{N'} \right) \]  

(Eq 4-19)

where \( N' = m_2 - m_1 + 1 \) is the total number of points selected from the spatial array and \( S_r^{(1)'}, S_r^{(2)'} \) are the spectra of the selected section of fiber.

For verification processing, only one match is preferred; thus, we get the hypothesis test:

\[ \begin{align*} 
\mathcal{H}_0: & \quad s_r^{(1)}[n] \text{ and } s_r^{(2)}[n] \text{ are from different fibers} \\
\mathcal{H}_1: & \quad s_r^{(1)}[n] \text{ and } s_r^{(2)}[n] \text{ are from the same fiber} 
\end{align*} \]  

(Eq 4-20)

where \( \mathcal{H}_0 \) outputs “No”, and \( \mathcal{H}_1 \) outputs “Yes”.

Additionally, we define the fiber ID resolution, \( L_{res} \), in our system as:

\[ L_{res} = (L_{max} - L_{min})|_{EER<0.01} \]
where $EER$ is the equal error rate[31].

**Experiment**

In order to demonstrate the proposed integrated fiber ID verification system and evaluate the performance of this approach, the digital system structure described above was implemented using a field-programmable gate array (FPGA). The following hardware and system parameters were used during experimentation: SCL, Alcatel 1905 LMI, 1550nm; current threshold, 75mA; MZI delay, $\tau_d = 100\text{ns}$; BPD1, BPD2, with BW 200MHz; ADC sampling rate, $F_s = 125\text{MHz}$; DAC sampling rate and system processing frequency equal to the ADC sampling rate; $p_f = 5$; $T = 80\text{ns}$; $f_{\text{REF}} = 12.5\text{MHz}$; and frequency-swept velocity of laser $\nu = 125\text{GHz/\text{ms}}$.

To investigate the performance of passive and active control for the laser source, we use the phase error information (UP, DN) given by (Eq 4-6)(Eq 4-7) to quantify sweep velocity linearity. **Figure 4-5** shows the experimental results using both passive and active control methods over time. As the statement in the principle section

![Figure 4-5](https://via.placeholder.com/150)

**Figure 4-5** The results at different control methods. (a) Fast Fourier Transform (FFT) results; (b) Phase error results; (c) Power spectral densities (PSDs) of the phaser errors; (d) Short-time FFT (STFT) of MZI signal after completing current curve iterating (Initial passive); (e) STFT of MZI over some time (Passive). (f) STFT of MZI signal at active control.
describes, the performance of the laser changes over time even when using the same driving current. As depicted in Figure 4-5(a), although we initially capture a single peak MZI signal at 12.5 MHz with a magnitude of about 15 dB over the noise, as time progressed, the noise increased by about 5 dB. Figure 4-5(b)(c)(d)(e), demonstrate a similar effect; simultaneously, these results demonstrate a solution to this problem via active control. When compared with the passive FFT result in Figure 4-5(a), active control demonstrates the same initial performance as passive control with approximately a 15 dB signal to noise ratio (SNR). The phase error in the time domain (Figure 4-5(b)) and frequency domain (Figure 4-5(c)) demonstrate that active control can maintain the frequency-swept velocity of the laser. Therefore, we can implement the swept frequency laser technique as the laser source for the measuring sub-system in order to execute fiber ID verification.

By applying (Eq 4-15), we can convert the BPD2 frequency information to distance, equaling to the distance term, \( d \), as illustrated in Figure 4-6. There are substantial reflecting peaks at the beginning, \( L_0 \), and the terminal positions. In our measuring sub-system, \( L_0 = 0.854 \text{m} \), and there is an angled optical fiber connector (APC) at \( l = 15 \text{m} \). The expanded result illustrated in Figure 4-6(a) is the fiber PUF signal with a travel distance ranging from 1 m to 2 m which corresponds with a 50-cm fiber length.

If we treat the DUT optical fiber as having various fiber ID segments, then we can use (Eq 4-19)(Eq 4-20) to get the cross-correlations between them. When calculating the cross-correlation between the fiber ID segment shown in Figure 4-6(a) with another one using the same segment, we get the genuine result in Figure 4-6(b), in which a
significant peak is at the original point; however, when correlating the fiber ID with another reflection signal at a distance ranging from 2 m to 3 m, no peak results, as shown in Figure 4-6(c). From this result, we use the cross-correlation peak value as a threshold to verify the fiber ID.

Since the prior work [28] has demonstrated that changes in temperature (range from 30-55°C) resulted in an EER of <0.6%, we ignored the temperature influence in this paper. In order to quantify the verification accuracy of the fiber PUF reader, 10 independent fiber IDs, each with 5-cm in length at distances ranging from 1 m to 2 m, were studied. 640 measurements were gathered from each fiber ID, resulting in 6,400 IDs in total. Then there are $10 \times \binom{640}{2} = 2,044,800$ total scores that can be implemented to calculate the genuine distribution $P\{S|\mathcal{H}_1\}$ and $(10 \times 640^2 = 18,432,000$ total scores to get the impostor distribution $P\{S|\mathcal{H}_0\}$ which are defined in (Eq 4-20). To simplify processing, 5,000 scores in each distribution were randomly selected in order to define the related probability distribution functions (PDFs). Next, the receiver operating characteristic (ROC) curves were generated by setting the probability of false alarm ($P_{FA}$) to determine the corresponding probability of detection ($P_D$). In the security system algorithm, EER is used to evaluate the performance of the

![Figure 4-6](image_url)

*Figure 4-6* The experimental results. (a) Reflection signal captured from BPD2; (b) Genuine cross-correlation result; (c) Impostor cross-correlation result.
fiber PUF reader. Both the PDFs and ROC curve are shown in Figure 4-7. The EER in this experiment was calculated to be 0.15%.

In order to evaluate the performance of the fiber PUF reader using fibers of various lengths, experiments with 0.5-cm, 1.5-cm, 2.5-cm, 4-cm, and 5-cm fiber lengths were conducted, resulting in the generation of the ROC curves depicted in Figure 4-8. In these ROC curves, shorter fiber lengths cause a reduction in verification accuracy. Generally, to obtain a robust verification (EER < 1%), a fiber length greater than 4-cm should be implemented when using the system.

**Conclusion**

To summarize, this work demonstrates a viable system for the utilization of optical fiber Rayleigh backscatter profiles as PUFs with high accuracy and with minimal complexity and cost. From the observed results of this investigation, it can safely be

*Figure 4-7* The performance of PUF reader. (a) Genuine and imposter distribution; (b) Verification receiving operating characteristic (ROC) curve.

*Figure 4-8* The accuracy of the PUF reader.
concluded that the fiber PUF reader can recognize a 5-cm fiber ID segment with a maximum EER of 0.15%. The core of the optical fiber PUF reader is a SCL, which exists in all optical transceivers in any optical communication or sensing system. Thus, the low cost and digitally integratable nature of this design is suitable to be integrated with optical transceivers, which will further reduce the system cost, promote technology adoption, and broaden its application areas, such as the Fiber to the X (FTTX) technologies including fiber to the home (FTTH), fiber to the premise (FTTP), fiber to the building (FTTB), fiber to the node (FTTN), etc.
Chapter 5 A reconfigurable architecture for continuous double-sided swept-laser linearization

by

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published in

Abstract

This paper reports a reconfigurable architecture to generate and maintain the linearity of frequency-swept semiconductor lasers for both current rising (up-sweep) and falling (down-sweep) edges, respectively. This method combines both passive and active controls to achieve a high frequency-sweep linearity using an integrated digital architecture. Both the passive and active control utilize the same feedback—the interferometric signal of the laser’s output—which has a frequency proportional to laser frequency-sweep velocity. This architecture is highly reconfigurable, allowing key parameters to be changed, including sweep velocity, tuning range, and sweep duration, making it a highly versatile and flexible approach suitable for a range of semiconductor lasers and a variety of applications. This architecture was implemented on an FPGA to validate the concept and evaluate its performance. By precisely tracking the interferometric feedback signal, it was found that this approach allowed a semiconductor laser to generate and maintain a phase error of less than $\pi/2$ within the regions of interest at different sweep velocities for both up-sweep and down-sweep cycles. Given that the delay length of the interferometer is 226 ns, the instantaneous optical frequency error is within 0.12% at 1552 nm.

Introduction

Frequency-swept lasers play an increasingly significant role in a range of scientific and industrial applications [32, 33] because of its sensitive measuring ability, highly spatial resolution, and low-power consumption. There are some applications for these frequency-swept lasers: the precise spectroscopy [34]; optical frequency-modulated continuous-wave (FMCW) LiDAR [7]; optical frequency domain reflectometer
(OFDR)-based distributed sensing [13]; and the optical fiber identification (FiberID) system by capturing the optical backscatter-based physical unclonable function (PUF) [35]. For each of these applications, the highly linear laser source is required to generate the well-controlled optical frequencies to achieve the high-fidelity frequency domain measuring. Various laser frequency control methods and designs [12, 36] have been proposed to provide the optical well-controlled frequencies for the above systems. Furthermore, there are several methods that have been investigated as means of improving the performance of OFDRs, such as injection locking [37], which enhances the modulation bandwidth; integrated structures [38, 39], which narrows the laser linewidth; and increasing the laser power [40], which adds the carrier density.

Importantly, the tuning range [41] of these lasers determines both sensing resolution and distance. The tuning range limitation of a laser measurement is a consequence of both the structure and material of the laser itself. Since the tuning range cannot be arbitrarily broad, it is impossible to achieve a truly continuous measuring system by increasing this property alone. A potential solution to this problem is to utilize the tuning range when the wavelength is both increasing and decreasing continuously. For distributed feedback semiconductor lasers (DFB-SCLs), it’s convenient to sweep the frequency by modulating the forward current [42]. There are designs [4, 36] that only consider the rising edge of the laser’s driving current; the falling edge is not used to generate data, but rather to release the drive current in order to reset the system in preparation for repeated laser frequency-sweeps or measuring periods. In other words, only the laser’s up-sweep tuning is used for measurement purposes; potential measurements that could have been made using the down-sweep tuning are ignored. An
ideal current control method for frequency-swept lasers would have several requirements: the method should be reconfigurable so that it can be adapted to satisfy the demands for different lasers and different applications [36]; the method should maintain the linear frequency-sweep continuously through both passive and active control algorithms; and the control system should be uncomplicated to apply.

In this manuscript, we introduce a control method that utilizes both the driving current rising and falling edges for a DFB-SCL to generate and maintain the continuously well-controlled linear optical frequency, which meets all the above demands. Even though the laser amplitude changes as a result of the changing injecting current, it can be uniformed with signal processing. Since the laser power can be treated with a roughly linear relation with its injecting current [43], and the laser amplitude has a linear relation with the square root of its power, the laser amplitude can be uniformed by dividing the square root of the driving current to minimize the influence of the current. To determine the feasibility of the proposed method and further examine its

Figure 5-1 Schematic of the double-sided frequency-swept semiconductor laser system. CPL0, CPL1, CPL2, optical couplers; BPD1, BPD2, balanced photodetector; Source subsystem: DFB-SCL, distributed feedback semiconductor laser working as optical source; MZI, Mach-Zehnder interferometer providing the time delay; CZ, cross-zero circuit; CL, laser frequency-sweep control loop; DAC digital-to-analog converter; Measuring subsystem: CIR, optical circulator; APC, angled physical fiber connector; ADC, analog-to-digital circuit for laser driving current.
universal engineering advantages, the method was implemented using the electronic-photonic control schematic depicted in Figure 5-1 on a digital circuit. Various experiments were conducted systematically by using the same control algorithm. The detailed design theory and experiments of this investigation are described in the following sections. Overall, the proposed iterating algorithm for continuously linear frequency-swept SCLs takes less than 1 second to capture the current pre-distortions. Both the driving current rising and falling edges were successfully utilized to capture the position of reflective objects precisely.

**Theory and Implementation**

* A. Overview

The overall schematic of the control method is drawn in Figure 5-1. A distributed feedback (DBF) SCL provides the optical source and is modulated by a time-varying current curve controlled by the digital control loop (CL) outlined in Figure 5-2. CPL0 is applied here to separate 5% of the laser power to a Mach-Zehnder interferometer (MZI); then the interfering optical signal of the MZI is converted to electric radio frequency signal by BPD2 which is applied to derive the optical sweep velocity. The

![Figure 5-2 Control loop (CL) working structure. Passive control: TDC, time digital converter; INTp, passive digital integrator; RB, Reading from BRAM function; WB, Writing to BRAM function; Active control: DXOR, double-sided exclusive or gate; INTa, active digital integrator. CZ and DAC are with the same the functions shown in Figure 5-1.](image)
remaining 95% laser power from CL0 works as a linear frequency-swept optical source for the measuring subsystem. As mentioned in prior work [12], a small MZI delay, $\tau_d$, results in the flat phase and frequency response. During a linearly frequency-swept period, $T_c$, we assume the SCL optical frequency, $f_{opt}(t)$, as:

$$f_{opt}(t) = f_0 + vt + e(t)$$

(Eq 5-1)

where $f_0$ is the initial optical signal frequency, $v$ is the laser frequency slope over time, which is also termed as the frequency-sweep velocity (FSV), $e(t)$ is the corresponding frequency error in the laser frequency sweeping, and $t$ is the time. Furthermore, the MZI phase, $\phi_{MZI}(t)$, is the phase difference between its two arms [44] [45], which can be expressed as:

$$\phi_{MZI}(t) = \phi(t) - \phi(t - \tau_d)$$

$$= 2\pi \left( \int_0^t f_{opt}(m)dm - \int_0^{t-\tau_d} f_{opt}(m)dm \right)$$

$$= 2\pi \left( v\tau_d t + E(t) - E(t - \tau_d) + f_0\tau_d - \frac{v\tau_d^2}{2} \right)$$

(Eq 5-2)

where $\tau_d$ is the MZI delay time and $E(t)$ is the integral of $e(t)$, or $E(t) = \int_0^t e(m)dm$.

If $\tau_d$ is a small value, we can assume that:

$$e(t) = \frac{d}{dt} E(t)$$

$$\approx \frac{E(t) - E(t - \tau_d)}{\tau_d}$$

(Eq 5-3)
The terms $u_r$ and $u_f$ are defined as the frequency sweep velocities of the drive current’s rising and falling edges, respectively. Then we set:

$$u_r = -u_f$$  

(Eq 5-4)

where the negative sign connotes the inverse relation between two edges.

As a critical component in the control algorithm, the CL module illustrated in Figure 5-2 deserves further attention. It is shared in both current rising and falling edges, in which passive and active control methods are achieved. As a result, its principle is expressed in the following sections.

**B. Current rising edge**

To implement the self-trained pre-distortion current curve to achieve a highly linear frequency-swept laser output, a passive control subsystem is used. After capturing the BPD1 signal, $u(t)$, the cross-zero circuit (CZ) measures the BPD1 frequency by obtaining the output, $S_{CZ}(t)$, as:

$$S_{CZ}(t) = \begin{cases} 1, & u(t) > 0 \\ 0, & u(t) \leq 0 \end{cases}$$  

(Eq 5-5)

For the digital control system, the digitalized CZ sequence is described as:

$$S_{CZ}[n] = S_{CZ}(t)|_{t=nT_s}$$  

(Eq 5-6)

where $T_s = \frac{1}{F_s}$ and $F_s$ is the main processing frequency in the digital circuit.

There are both passive and active control subsystems at the rising edge of the current curve. As mentioned in our previous work [35, 36], we utilize the time digital converter (TDC) in the control system to measure the optical frequency slope by
comparing the CZ period with the expected one, then the phase error signals (UP, DN) are captured by comparing the CZ counter with the frequency parameter $p_f$. Using the current rising edge as an example, the working theory of the TDC can be expressed as:

$$
C[n] = \begin{cases} 
C[n-1] + 1, & S_{CZ}[n] = S_{CZ}[n-1] \\
1, & otherwise
\end{cases} 
$$

(Eq 5-7)

$$
UP[n] = \begin{cases} 
1, & C[n] > p_f \\
0, & otherwise
\end{cases} 
$$

(Eq 5-8)

$$
DN[n] = \begin{cases} 
p_f - C[n-1], & C[n-1] < p_f, S_{CZ}[n] \neq S_{CZ}[n-1] \\
0, & otherwise
\end{cases} 
$$

(Eq 5-9)

where $C[n]$ is the main clock cycle counter over the number $n$ and set to 1 at every switching point in CZ. Figure 5-3 gives an example of the $S_{CZ}[n]$, UP[n], and DN[n] if $p_f = 4$. The equivalent reference frequency in the TDC for the control loop can be treated as:

$$
f_{REF} = \frac{F_s}{2p_f}
$$

(Eq 5-10)

After digitizing, since the count $p_f$ corresponds with half of a period, i.e., $\pi$, the phase error in (Eq 5-2) can be expressed as:
\[ E[n] = \frac{\pi |\sum_{i=0}^{n}(UP[i] - DN[i])|}{p_f} \]

(Eq 5-11)

\( E[n] \) is then applied to derive frequency accuracy; from (Eq 5-3) we can determine the frequency error \( e[n] \) expressed in Hz, as:

\[ e[n] \approx \frac{E[n] - E[n - \tau_d F_s]}{\tau_d} \]

(Eq 5-12)

Since the \( f_0 \) is constant, then using (1) we define the frequency error rate, \( \gamma[n] \), as:

\[ \gamma[n] = \frac{e[n]}{vnfs} \]

(Eq 5-13)

By accumulating the frequency errors UP and DN from the TDC, the passive digital integrator (INTp) result can be processed as:

\[ INT_p[n] = c_p \sum_{i=0}^{n} (UP[i] - DN[i]) \]

(Eq 5-14)

where \( c_p \) is the adaptive gain coefficient for INTp. Then the integrating signal is sent to an adder as shown in Figure 5-2, working as the feedback in the laser control.

For the passive control process, the Writing BRAM (WB) is the combination of the INTp and the Reading BRAM (RB) and is generated by:

\[ WB[n] = INT_p[n] + RB[n] \]

(Eq 5-15)

The initialized values in the RB are equal to a constant current that corresponds to a current which is greater than the SCL threshold to obtain the oscillation. In the passive
self-training algorithm, we term the function exchanging between the WB and RB as one iteration. If the iterating number reaches the set threshold, \( N_l \), the function exchanging between the WB and RB stops. The reason why we treat this control as passive is that the INTp is only working on the following laser control period, not its corresponding period. Thereafter, the laser current only depends on the RB values and the current curve is not affected by the phase error information from the TDC. In this scenario, the RB works as the digital-to-analog converter (DAC) input for the passive control as:

\[
DAC_p[n] = RB[n]
\]

(Eq 5-16)

From the analysis in previous works \([35, 36]\), passive control is only applied to capture the laser’s pre-distortion curve. Since the SCL is so sensitive to its surrounding environment that the optical frequency performance is influenced as well, the active control is essential as a means of maintaining the linearity of the optical source.

The double-sided XOR (DXOR) gate is utilized to catch phase lead-lag information between the reference and the measurement as:

\[
DXOR[n] = \begin{cases} 
-1 & \text{REF}[n] = S_{cz}[n] \\
1 & \text{otherwise}
\end{cases}
\]

(Eq 5-17)

*Figure 5-4* Timing Diagram for the double-sided XOR gate (DXOR). REF, reference signal; CZ, cross-zero signal; DXOR, the output signal of the DXOR.
where $\text{REF}[n]$ has a frequency of $f_{\text{REF}}$. The timing of the DXOR is depicted in Figure 5-4. The output of the active integrator ($\text{INT}_a$) is expressed as:

$$\text{INT}_a[n] = c_a \sum_{i=0}^{n} \text{DXOR}[i]$$

(Eq 5-18)

where $c_a$ is the adaptive gain coefficient for the $\text{INT}_a$.

For an ideal optical frequency locking, the output of the $\text{INT}_a[n]$ should be 0 since the period of DXOR is $T = \frac{T}{2}$, where $T = \frac{1}{f_{\text{REF}}}$, with a duty cycle $\frac{50}{50}$.

In an active control subsystem, the current curve is generated by combing the $\text{RB}[n]$ and $\text{INT}_a[n]$ together, which is:

$$\text{DAC}_a[n] = \text{RB}[n] + \text{INT}_a[n]$$

(Eq 5-19)

where $\text{RB}[n]$, which is generated by the iterating algorithm in the passive control subsystem, works as the laser current pre-distortion in the active control.

C. Current falling edge

As mentioned above, the only difference between the drive current rising and falling edges is how the control system responds to the phase error information from the TDC. As a result, for the current falling edge in the passive control subsystem, (Eq 5-14) should be converted as:

$$\text{INT}_p[n] = -c_p \sum_{i=0}^{n} (\text{UP}[i] - \text{DN}[i])$$

(Eq 5-20)

where the minus symbol before the $\text{INT}_p$ coefficient is from (Eq 5-4).
Similarly, in the active control subsystem, the $\text{INT}_a[n]$ in (Eq 5-18) should be:

$$
\text{INT}_a[n] = -c_a \sum_{i=0}^{n} \text{DXOR}[i]
$$

(Eq 5-21)

The similarity shared by the current rising and falling edges makes it convenient for us to construct this double-sided laser locking system by using the same control module multiple times, leading to the system’s simplified design.

D. Measuring

After establishing a highly linear optical frequency source by using both the passive and active control subsystems, we focus on the measuring section in the system and utilize the double-sided frequency-swept laser as the source. Given the time difference between the forward and reflected optical signals in measuring subsystem shown in Figure 5-1 is $\tau$, and the SCL is operated with an ideal frequency slope over time, $v$, the BPD2 output, $u_R(t)$, converted by the optical interfering signal can be expressed as [12, 35]:

$$
u_R(t) = \frac{A_r(t)^2}{8} \eta_2 \cos[2\pi(f_0 + vt)\tau]
$$

(Eq 5-22)

$$
\tau = \frac{2l + L_0}{u_p}
$$

(Eq 5-23)

where $A_r(t)$ is the amplitude of the laser power directed into the CPL1 as a time-varying function which is with a rough relationship with squared root of the laser power [43], $\eta_2$ is the BPD2’s light-to-voltage conversion coefficient, $f_0$ is the SCL initial frequency.
at the beginning of the sweep, $v$ is the optical frequency slope over time, $t$ is time, $L_0$ is the reflection caused by the CIR port 2, which is equal to the length difference between the CIR arm (shown in the measuring upper part) and the reference arm (shown in the measuring lower part), $l$ is the fiber length from the angled physical connector (APC) to the CIR port 2, $u_p$ is the light speed in optical fiber, and the $\tau^2$ components in the Measuring subsystem are ignored due to their negligible values. Thus, the reflected frequencies are $v_r \frac{d}{u_p}$ and $v_f \frac{d}{u_p}$ respectively in the current rising and falling edges. If we set $v_r = -v_f$ by sharing the same $p_f$ in both cases, then we can capture such a signal with the same frequency twice to enhance the measuring accuracy.

**Experiment**

To validate the described continuous self-trained laser current curve generating method and evaluate its performance, the digital circuit parts in the control system introduced above were implemented using an FPGA (field-programmable gate array). The following hardware and system parameters were used during experimentation: DFB-SCL, Eblana EP1550-NLW-B, 1552nm, current threshold 40mA, about 35GHz frequency tuning range, linewidth 200kHz; BPD1, BPD2, with bandwidth 200MHz; ADC, DAC sampling rates and system processing frequency equal to $F_s = 125$MHz; and a set of $\tau_d$: 10ns, 100ns, and 226ns. As mentioned in (Eq 5.11)(Eq 5.13), the phase error $E[\pi]$, and frequency accuracy, $\gamma[\pi]$, can be used as measures of system performance.
To investigate the passive control performance, we first applied an MZI delay, $\tau_d = 100\text{ns}$, to determine iteration results. The phase error can be utilized as a standard with which to judge the performance of the frequency-swept laser. For an ideal linear sweeping source, both UP and DN should be 0, and the count $p_f$ corresponds with phase $\pi$.

*Figure 5-5* shows the results over various iterations with $p_f = 5$. From *Figure 5-5(a)*, we observe that the laser’s initial current is a constant value which is above the threshold. Since the short-time fast Fourier transform (STFFT) represents the frequency changing over time and the BPD1 frequency corresponds with laser frequency slope from (Eq 5-2)(Eq 5-10), the STFFT of BPD1 can be converted to the laser slope, or the FSV. The STFFTs of BPD1 with 50, 300, and 1000 iterations are shown in *Figure 5-5(d)-(f).* During each subsequent iteration, the driving curve is trained to better
achieve a linear frequency-sweep, which demonstrates the self-training algorithm. *Figure 5-5(f)* demonstrates that, after 1000 iterations, the MZI output approaches a linear lock at the target frequency, 12.5 MHz, corresponding to a nearly linear laser sweep velocity of 125GHz/ms at both edges of the driving current in the time regions of interest (ROI) with a frequency tuning range of 31.25GHz. In order to quantify the performance of the current curves for various iterations, the phase error information (UP, DN) given by (Eq 5-8)(Eq 5-9) can be applied. *Figure 5-5(b)(c)* illustrate the phase errors in ROIs observed with numerous iterations utilized at the current rising and falling edges respectively. After successful training, the absolute phase errors were observed to decrease to less than $\frac{\pi}{2}$, and the frequency error rate, $\gamma[\pi]$, to less than 0.05%. Since one period for the current curve is 500 $\mu$s, 0.5 s was required in order to use passive control to generate the final laser pre-distortion curve.

**B. Active control**

*Figure 5-6* Results for passive and active controls for a 125GHz/ms FSV, $p_f = 5$. (a) Passive control after 30 minutes; (b) Active control after 30 minutes; (.*1) s are the phase errors at ROI of the current rising edge; (.*2) s are the phase errors at ROI of the current falling edge; (.*3) s are the STFFT results of the BPD1.
As described in Section II, while a pre-distortion curve can be used to drive a highly linear frequency-swept laser after training, the linearity will subsequently decrease due to changing environmental parameters (e.g. temperature) over time. Active control allows for continued linearization despite these environmental changes. Performance differences between passive and active control after 30 minutes of elapsed time are depicted in Figure 5-6 using \( p_f = 5 \) and \( \tau_d = 100\text{ns} \).

Based on the STFFT results shown in Figure 5-6(*.3), it is difficult to distinguish a difference between the passive and active controls after 30 minutes. However, differences in phase error between these two methods quantitatively demonstrate the improvement which results from the active control, shown in Figure 5-6(*.1, *.2). Using these two methods for 30 minutes, phase errors using passive control change over time, while phase errors using active control are minimized to less than \( \frac{\pi}{2} \) for the majority of the time, and \( \gamma[n] \) is less than 0.05%.

C. Universal performance

From (Eq 5-2)-(Eq 5-10), it is apparent that the frequency-sweep velocity not only depends on \( p_f \), but \( \tau_d \) as well. To illustrate the universality of our design, various \( p_f \) and \( \tau_d \) values were implemented in order to obtain the desired pre-distortion curves. There are the results depicted in Figure 5-7 by using the \( \tau_d = 10\text{ns}, p_f = 10, \tau_d = 100\text{ns}, p_f = 10, \) and \( \tau_d = 226\text{ns}, p_f = 5 \) respectively. It is clear that the \( \tau_d \) has a noticeable influence on the performance of our system; specifically, that longer delay leads to declining results for the linewidth limitation of the laser. To illustrate this result clearly, two comparisons between these settings were made. Evaluating the \( p_f = 10 \) cases, it is clear that the phase errors in the case with \( \tau_d = 100\text{ns} \) shown in Figure
5-7(b.*) are greater than the case with $\tau_d = 10\text{ns}$ in Figure 5-7(a.*); looking at the $p_f = 5$ cases, the phase errors in the case of $\tau_d = 100\text{ns}$ shown in Figure 5-7(b)(c) are less than the case with $\tau_d = 226\text{ns}$ in Figure 5-7(c.*) respectively. Even with these differences, our design still performs well using a range of settings. Specifically, when assuming $f_0$ is 153THz, for the $\tau_d = 10\text{ns}$ design, the $\gamma[n]$ is less than 0.5%; for the $\tau_d = 226\text{ns}$ design, the $\gamma[n]$ is less than 0.12%.

D. Reflective measurements

As mentioned in the introduction section above, double-sided laser locking is proposed as a means of fully utilizing the frequency-sweep tuning range of a laser source, enhancing the OFDR performance and allowing for continuous measurement. (Eq 5-22) and (Eq 5-23) describe the relationship between the reflector position and the

![Figure 5-7 Results with different control parameters. (a) $p_f = 10$, $\tau_d = 10\text{ns}$; (b) $p_f = 10$, $\tau_d = 100\text{ns}$; (c) $p_f = 5$, $\tau_d = 226\text{ns}$; (*)1 s are the phase errors at current rising ROI part; (*)2 s are the phase errors at current falling ROI part; (*)3 s are the STFFT results of the BPD1.](image)
In our measuring design, $L_0 = 0.854$ m, and there is an APC at $l \approx 16$ m.

Various parameters were implemented in order to capture the APC reflection as shown in Figure 5-8, where all the APC terminations correspond with the same fiber length. Figure 5-8(a.*) illustrates the reflections captured at the current rising ROIs; Figure 5-8(b.*) illustrates the reflections captured at the current falling ROIs. From (Eq 5-4) and (Eq 5-22), the reflective signals at both edges occur at the same frequencies, and it is difficult to distinguish differences between the signals in the rising and falling edge cases (Figure 5-8 with the same label numbers).

There are still several important engineering challenges remaining in the proposed system. Chief among them is form the integral $p_f$, which limits the possible tuning frequency-sweep velocities for a given optical system. Furthermore, in order to generate more precise and flexible sweep velocities, a higher system frequency must be applied, increasing the cost and complexity of the system.

**Figure 5-8** Reflections at different settings. (a) ROI at current rising edges; (b) ROI at current falling edges; (*1) s are the results with $p_f = 10, \tau_d = 100$ ns; (*2) s are the results with $p_f = 5, \tau_d = 100$ ns; (*3) s are the results with $p_f = 5, \tau_d = 226$ ns.
Conclusion

Since the salient role of the frequency-swept laser is found in a growing range of application areas, it attracts the focus of many researchers. A fully utilized laser tuning range contains both the current rising and falling edges; however, present techniques do not utilize the falling edge for sensing purposes. This work investigates a method to generate and maintain the well-tuned double-sided current curves utilizing both rising and falling currents the frequency-swept SCLs. The results obtained in this investigation demonstrate that the control method described in this work can complete well-tuned double-sided current curves and maintain them for 30 minutes using GHz/ms FSV. In the 226ns delay case, the frequency error rate observed to be within 0.12%. This design can be implemented to enhance the performance in a wide variety of applications, including spectroscopy, FMCW LiDAR, and fiber sensing.
Chapter 6 An integrated OFDR system using combined swept-laser linearization and phase error compensation

by

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Abstract

This paper describes a flexible, compact, semiconductor laser (SCL)-based optical frequency-domain reflectometry (OFDR) design, incorporating swept-laser linearization functions (both passive and active), in-situ phase error evaluation function, and phase error compensation function. At the core of the proposed design, a fiber-optic Mach-Zehnder interferometer (MZI) and its electronic digitization components (photodetector and zero-cross circuit) simultaneously perform all abovementioned functions. Benefitting from the unique digital design, this system combines all functions in a compact fashion without increasing complexity and overhead. Experimental results demonstrate enhanced accuracy using the integrated design as compared to prior methods, as well as improved cost-effectiveness, reduced complexity, and optimized form factor.

Instruction

In the optical fiber sensing area, there are two basic categories: optical time-domain reflectometry (OTDR) [33] and optical frequency-domain reflectometry (OFDR) [32]. Since the spatial resolution (SR) in OTDR is limited both by the optical pulse width and system sampling interval [46], this technology is generally implemented in the long-haul situations. Different from the measurement in OTDR, the signal in OFDR is represented by the optical frequency spectrum, and SR is determined by the laser frequency tuning range [47]. As a result, it is much more convenient to achieve the high SR in OFDR than OTDR. In addition to the high SR, OFDR offers other engineering advantages [13, 48] such as significant multiplexing capability and low-energy consumption. Consequently, a range of important applications have OFDR technology
at their foundation, including frequency-modulated (FM) spectroscopy [34], frequency-modulated continuous-wave (FMCW) LiDAR [7, 17, 48] optical coherence tomography [49], and advanced intravascular medical imaging [50]. A linear frequency swept laser source provides the foundation of all OFDR systems, and the linearity of this the frequency sweep is the primary factor limiting both coherence length and signal-noise-ratio (SNR). Research efforts have focused on generating a highly linear sweep based on analog control [9], digital control [36], as well as hybrid methods [12]. While these techniques have been shown to significantly enhance the sweep linearity of semiconductor lasers (SCLs), residual phase error, and its deleterious effect on measurement precision, remains. On another path, several techniques have been investigated as a means of reducing phase errors to equivalently enhancing sweep linearity, including reducing external clock sampling error [51, 52], introducing interpolation in processing [53], and compensating the optical phase error [54], all of which rely on postprocessing techniques. While these techniques are effective in reducing the deleterious effect of phase error on measurement precision, they require additional electronic hardware for post-processing that is complex, bulky, and both energetically and economically costly [12]. These drawbacks limit the adoption of OFDR techniques in application areas where it may otherwise provide significant technical benefit. Consequently, the question of how to build an affordable, precise, and compact OFDR system represents an important area of technical research.

This paper describes an OFDR system designed to overcome the afore-mentioned limitations. By integrating both linearization and phase error correction techniques into a single combined platform, the OFDR system is able to achieve linear frequency
sweeps with minimized phase error without sacrificing system accuracy or introducing additional components.

Theory and Implementation

A. Overview of the compact OFDR

In order to construct a standard, high-precision OFDR system, four functional modules are required, as shown in Figure 6-1(a). These modules are:

1. the linearization control (L-Ctrl) module, which captures the feedback (FB) signal from the frequency-sweep laser source (FSL) and sends a corresponding control (Ctrl) signal back to the FSL;
2. the raw data measurement (RDM) module, which stores frequency information from the reflected signals;
3. the sweep linearity evaluation (LE) module, which quantifies the sweep linearity of the FSL;

![Figure 6-1](image-url) Design of a standard high-precision OFDR system: (a) functional block diagram (FSL: frequency-sweep laser source; L-Ctrl, linearization control module; LE, linearity evaluation module; CMP, phase error compensation module; RDM, raw data measurement module; IM, improved measurement); (b) schematic system illustration (DFB-SCL, distributed feedback semiconductor laser; CPL, optical coupler; MZI, Mach-Zehnder interferometer; BPD, balanced photodetector; ADC, analog-to-digital converter; PFD, phase frequency detector; LPF, low-pass filter; FB, feedback signals; Ctrl, control; APC, angled physical connector).
4. and the phase error compensation (CMP) module, which corrects residual phase error in the RDM to maximize measurement precision.

All prior implementations of such a system have required a complex collection of system elements as illustrated in Figure 6-1(b).

When analyzing these designs to determine which modules require similar or identical system hardware elements, we can find that the Mach-Zehnder Interferometer (MZI), balanced photodetector (BPD), and analog-to-digital converter (ADC) are all used in the RDM, LE, CMP, and L-Ctrl modules, suggesting that a single iteration of these subsystems can be shared across modules. To achieve this aim, the LE and CMP are combined in the L-Ctrl as a multi-function control (M-Ctrl) module shown in Figure 6-2(a) to achieve a single low-overhead OFDR system; in other words, the M-Ctrl module not only accomplishes frequency-sweep linearization control for the distributed feedback semiconductor laser (DFB-SCL), but also provides the frequency linearization quantification and accurate timing information in a frequency-sweep laser for improved measurement (IM). Additionally, to reduce the design cost further, the ADC utilizes a

![Figure 6-2](image)

Figure 6-2 Design of a compact OFDR system: (a) functional block diagram (M-Ctrl, multi-function control module; FSL, CMP, and RDM are the same as shown in Figure 6-1(a)); (b) schematic system illustration (ZC, zero-crossing circuit; TDC, time-digital converter; DCTL, digital control; DFB-SCL, CPL, MZI, BPD, ADC, FB, Ctrl, and APC are the same as in Figure 6-1(b)).
zero-crossing circuit (ZC) as shown in Figure 6-2(b). A detailed description of the components that make up this system will be expressed in the following sections.

B. Linearization control functions

The frequency of a DFB-SCL is directly determined by its driving current; however, this relationship is non-linear. Thus, when constructing an OFDR system using a DFB-SCL as the linear frequency-swept optical source, linearization control is required. There exist two general linearization strategies: passive and active. Passive control, also known as pre-distortion generation [22], uses successive iterations of a current pre-distortion curve to drive a DFB-SCL, with refinements to the curve added with successive iterations [36, 55]. In contrast, active control [9, 17] corrects phase error in real time during each injection current ramp via an optical phase lock loop design. This section explains the realization of both active and passive frequency-sweep linearization using the compact design shown in Figure 6-3.

Under the assumption that a DFB-SCL is modulated in order to output a constant frequency-sweep velocity, \( v \), the AC-coupled BPD, which resides in the M-Ctrl, is
applied to convert the optical signal to an electrical one. Its output, $u_c(t)$, can be expressed as:

$$u_c(t) = \frac{A_c^2(t)}{8} \eta \cos[2\pi(f_0 + vt)\tau_d + e(t)]$$

(Eq 6-1)

where $A_c(t)$ is the optical amplitude, which is positive, directed into the MZI changing over time, $t$; $\eta$ is the light-to-voltage conversion coefficient of the BPD; the $\phi(t) = 2\pi[(f_0 + vt)\tau_d + e(t)]$ part is termed as the phase of this analog signal, $f_0$ is the starting optical frequency of the SCL; $\tau_d$ is the time delay introduced by MZI, and $e(t)$ is the frequency variation in SCL linear sweeping.

Followed by the BPD, the digitized zero-crossing (ZC) signal, $ZC[n]$ is described as:

$$ZC[n] = u_c(t)|_{t = \frac{n}{F_s}} > 0$$

(Eq 6-2)

where $F_s$ is the digital system frequency. The $ZC[n]$ shares the same frequency as $u_c(t)$ by converting the analog signal to binary values, which is achieved by setting the Schmitt trigger phase $\alpha$ in the former work [36] as 0.

In the M-Ctrl module, a time-digital converter (TDC) is utilized as the phase comparator to capture the lead-lag (UP, DN) relationships between the $ZC[n]$ and the equivalent reference signal as follows:

$$C[n] = \begin{cases} 1 & ZC[n] \neq ZC[n - 1] \\ C[n - 1] + 1 & \text{else} \end{cases}$$

(Eq 6-3)
\[ UP[n] = \begin{cases} 1 & C[n] > p_f \\ 0 & \text{else} \end{cases} \]

(Eq 6-4)

\[ DN[n] = \begin{cases} \frac{p_f - C[n] - 1}{2} & C[n] - 1 < p_f, ZC[n] \neq ZC[n - 1] \\ 0 & \text{else} \end{cases} \]

(Eq 6-5)

where \( C[n] \) is the parameter used to count the sampling number, and \( p_f \) is the frequency parameter of the TDC, determining the equivalent reference frequency as \( \frac{E_s}{2p_f} \).

Figure 6-4 illustrates the relations among these signals in TDC when \( p_f = 5 \).

From (Eq 6-2), the ZC frequency, \( f_{ZC} \), can represent \( v \) as [35, 36]:

\[ v \approx \frac{f_{ZC}}{\tau_d} \]

(Eq 6-6)

where the \( \approx \) is caused by the \( e(t) \) in (Eq 6-1). Thus, only when \( f_{ZC} \) is a constant value over time can we define that there is a linear frequency-sweep laser source.

During the control period, UP and DN from the TDC represent the lower and higher relation between the frequency-sweep velocity and desired value, respectively. The accumulating error in the ZC can then be expressed as:

\[ E[n] = \sum_{i=0}^{n} (UP[i] - DN[i]) \]

(Eq 6-7)
where $i$ is the discrete-time index for these digital signals.

When applying passive control, the coefficient, $C_a$, for the active integrator (INTa) is set as 0. As a result, the laser performance depends on the reading BRAM (RB) only. The iteration algorithm is illustrated in the following way: in the $k^{th}$ laser chipping period, the laser is driven by $RB^{(k)}$, and the other signals in this period can be expressed with the label $(k)$ as:

$$WB^{(k)}[n] = C_p E^{(k)}[n] + RB^{(k)}[n]$$

(Eq 6-8)

where $C_p$ is the coefficient for the passive integrator (INTp).

Then, for the following laser frequency sweep period, $(k + 1)$, the laser driving current is determined by the $RB^{(k+1)}$ as:

$$RB^{(k+1)}[n] = WB^{(k)}[n]$$

(Eq 6-9)

Since the FB signal in period $k$ only affects the laser driving current in following period $k + 1$, this method is defined as passive control, and the data transfer between the RB and WB is defined as one iteration. When the number of iterations reaches a specified value, $K$, iteration stops, and the RB is implemented as the final current pre-distortion driving current curve.

In contrast, when using active control, which is introduced to mitigate the environment influence, there is no iteration at all, and the laser current, $D[n]$, is the combination of the INTa and the pre-distortion as:

$$D[n] = RB^{(K)}[n] + C_a E[n]$$

(Eq 6-10)
where $RB^{(k)}$ is the pre-distortion and $C_a$ is the coefficient of INTa.

The M-Ctrl module is comprised of all necessary components for both passive and active controls.

**C. Linearity evaluation function**

Since the short-time fast Fourier transform (STFFT) of the BPD signal in the M-Ctrl captures frequency change over time, it can be directly utilized to evaluate the laser frequency-sweep velocities by (Eq 6-2). However, there are some limitations: firstly, extra ADC components are needed to measure the BPD signal; additionally, since the STFFT averages the frequency information over several signal periods, it cannot capture small changes in real time. To overcome these drawbacks, a LE method [12] was used to evaluate the frequency-sweep linearity by employing an additional MZI subsystem. Although this method is effective, it substantially increases system complexity. This section describes a novel, consolidated LE which captures real-time evaluating information.

As mentioned in section II B, the accumulation error function $E[n]$ is used to adjust the laser driving current. For ideal control results, $E[n]$ is equal to 0, meaning that the generated optical source is driven via a perfect current curve which results in a constant frequency-swept velocity, $v$. In other words, $E[n]$ represents the phase error in the optical frequency-sweep source. Given the definition of the TDC, $p_f$ corresponds with a half period. From this, we can express the phase error $\phi_{err}[n]$ in the SCL as:

$$\phi_{err}[n] = \frac{E[n]}{p_f} \pi$$

(Eq 6-11)
Due to the system’s digital circuit design, it is convenient to evaluate frequency-sweep linearity by applying the UP and DN signals from the TDC to calculate the phase error $\phi_{err}[n]$ as $(Eq \ 6-7)(Eq \ 6-11)$ without introducing a second MZI subsystem. Thus, frequency linearity evaluation can be achieved using the M-Ctrl module alone without the need for further components.

**D. Phase error compensation**

During laser chirping, even when both the passive and active controls are implemented for the SCL, $\phi_{err}[n]$ still occurs due to unavoidable physical and environmental effects. As a result, correction algorithms [12, 54] are introduced to compensate for these errors in order to improve the OFDR measuring precision. These compensation algorithms, however, increase the complexity and cost of the OFDR system. To reduce the overhead from such algorithms, the M-Ctrl outputs both the laser control signals for the ZC and accurate timing information (Index in Figure 6-3) during a frequency-sweep, which allows it to be implemented as a ‘clock’. Put simply, the phase error compensation algorithm is resampling the RDM at each zero-crossing event [28, 56].

Since the cosine-like analog signal, $u_c(t)$, consists of a cosine signal with phase $\phi(t)$ and magnitude $\frac{A_c^2(t)}{8}\eta$, which is greater than 0, the zero points, $n'$, for $u_c(t)$ in $(Eq \ 6-1)$ are same with the zero points in its cosine signal. Considering the prosperity of the cosine wave, the phase at $n'$ can be expressed as:

$$\phi[n'] = 2\pi \left[ f_0 + \nu \frac{n'}{F_s} + e \left( \frac{n'}{F_s} \right) \right] \tau_d$$

$$= \left( m + \frac{1}{2} \right) \pi$$
where \( m \) is an integer, \( F_s \) is the system sampling frequency.

As the optical signal in the RDM from the DFB-SCL is divided by CPL0, both sources in two subsystems share the \( f_0 \), \( e(t) \), and \( v \) simultaneously. Similar to the M-Ctrl signal in (Eq 6-1), if we assume there is an echoed point at distance \( R \) in the RDM, the corresponding reflective signal, \( M[n] \), captured in the ADC can be described as:

\[
M[n] = \frac{A_M^2(t)}{8} \eta \cos \left( 2\pi \left( f_0 + vt + e(t) \right) \left( \frac{2R + L_0}{u_p} \right) \right) \bigg|_{t=\frac{n}{F_s}}
\]

(Eq 6-13)

where \( A_M(t) \) is the optical amplitude, which is positive, directed into the CPL1 changing over time, \( t \); \( L_0 \) is the constant difference between the reflected and forward optical paths in Figure 6-2, and \( u_p \) is the propagation velocity in fiber. If applying the CMP index, \( n' \), (Eq 6-13) would be modified as:

\[
M[n'] = \frac{A_M^2 \left( \frac{n'}{F_s} \right)}{8} \eta \cos \left[ \left( m + \frac{1}{2} \right) \pi \left( \frac{2R + L_0}{\tau_d u_p} \right) \right]
\]

(Eq 6-14)

As a result, the measuring inaccuracies caused by \( e(t) \) and \( f_0 \) in (Eq 6-13) would be eliminated. Thus, the phase error compensation function can be achieved without requiring the extra MZI subsystem.

**Experiment**

**A. Overview**

To demonstrate the performance of the proposed low-overhead compact OFDR design, a digital system was fabricated using a field programmable gate array (FPGA).
The following parameters and equipment were implemented during the experiments as shown in Table 6-1.

Table 6-1 Parameters or equipment in experiments

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Value or name</th>
</tr>
</thead>
<tbody>
<tr>
<td>DFB-SCL</td>
<td>Eblana Photonics EP1550-NLW-B</td>
</tr>
<tr>
<td>BPD bandwidth</td>
<td>200MHz</td>
</tr>
<tr>
<td>Temperature controller</td>
<td>LDT-5910, 24.5°C</td>
</tr>
<tr>
<td>(\tau_d)</td>
<td>226ns</td>
</tr>
<tr>
<td>(F_r)</td>
<td>125MHz</td>
</tr>
<tr>
<td>CPL0</td>
<td>5/95</td>
</tr>
<tr>
<td>CPL1/CPL2</td>
<td>50/50</td>
</tr>
<tr>
<td>(P_f)</td>
<td>5</td>
</tr>
<tr>
<td>(\nu)</td>
<td>55.3GHz/ms</td>
</tr>
<tr>
<td>(K)</td>
<td>10</td>
</tr>
<tr>
<td>(L_0)</td>
<td>0.6m</td>
</tr>
</tbody>
</table>

where LDT-5910 was utilized to maintain the DFB-SCL temperature during the experiments.

B. The linearization controls

In order to investigate the performance of linearization control in the M-Ctrl, experimental trials were conducted.

As mentioned in the theory section, the passive control is used to generate the laser current pre-distortion and repeated iteration results in the increasing sweep linearity. Figure 6-5(a) illustrates the current curve iterating progress. It is obvious that the initial RB is a constant value which is above the laser current threshold and with the iteration, there is an almost constant current curve generated. To roughly estimate the performance of the iteration, STFFT results with 2 and 10 iterations are depicted in Figure 6-5(b) and (c), respectively. After 10 iterations, there is a nearly constant sweep velocity from 100 to 500\(\mu\)s from the Figure 6-5(c). Thus, a driving pre-distortion current for the DFB-SCL can be generated using the self-training algorithm.
However, the DFB-SCL is sensitive to the environment, which means passive control, or laser controlled by the current pre-distortion, is not sufficient to maintain the laser performance. For this reason, active control is utilized to maintain the laser frequency-sweep linearity under various environmental changes such as temperature. As mentioned in (Eq 6-10), the active control would respond to the feedback signal to adjust the laser driving current curve simultaneously. Figure 6-6 illustrates the current curve and STFFT result in (a) and (b), respectively in the active control. When comparing the STFFT results from active control and passive control shown in Figure 6-5(c) and Figure 6-6(b), it is difficult to distinguish any difference between the two; thus a quantitative evaluation method is required to rigorously evaluate laser frequency-sweep linearity over time, which was developed and implemented as described in the following section.

C. Linearity evaluation

Considering that the STFFT is not sufficient to quantitatively evaluate the sweep linearity, an additional evaluation system was proposed to quantify laser performance as expressed in Section II. As shown in (Eq 6-11), only if the $\phi_{err}[n]$ is a stable value
or oscillates in a narrow range, the laser would provide the highly linear frequency sweep for the OFDR system.

Figure 6-7 depicts the $\phi_{\text{err}}[n]$ at both the passive and active control periods. Considering that the laser driving current at the beginning is so finite that the optical signal has neither adequate power nor frequency-sweep linearity, the OFDR measurement cannot be completed. Then we define the following time as regions of interest (ROI) in which the accumulating phase error should be nearly constant. For example, Figure 6-7(b) draws the $\phi_{\text{err}}[n]$ at the ROI from 200 to 500$\mu$s which are the zoom-in results of the whole phase error shown in Figure 6-7(a) during the control period with different iterations. Compared with the STFFT results illustrated in Figure 6-7(c) and Figure 6-6(b), the $\phi_{\text{err}}[n]$ is much more veracious to evince the laser performance, since the ROI is from 200$\mu$s rather than 100$\mu$s. After 10 iterations, a highly linear frequency-sweep velocity has been achieved with a $\phi_{\text{err}}[n]$ below $\pi$ at the ROI. However, given the sensitivity of the DFB-SCL to its environment, even with the same driving current, laser performance deteriorates over time. Figure 6-7(c) shows a comparison of phase errors between the passive and active control after running the systems for an extended period of time. Here, the $\phi_{\text{err}}[n]$ are suppressed to less than $\pi$
at the ROI utilizing active control mostly, while the phase errors increase with time for passive control system without extra recalibration. Thus, the active control plays a significant role to mitigate the affect caused by the environment to maintain the laser frequency-swept linearity.

D. Phase error compensation

As mentioned in the former section, only an idealized FSL, can achieve \( \phi_{err}[n] = 0 \) at the ROI. Even when the passive and active controls are implemented to generate and maintain optical frequency-sweep linearity, intrinsic phase error is inevitable as shown in the evaluation results. Compensation is thus employed as a technique of further reducing the effects of \( \phi_{err}[n] \) on the measurement, improving measurement accuracy. In the algorithm used in this experiment, the index from the ZC is utilized as the ‘clock’ to resample the RDM in order to generate the IM (Figure 6-2(b)).
Considering Nyquist’s law and that the $\tau_d = 226\text{ns}$, in order to utilize the compensation algorithm, the reflecting position can be no greater than 25m. To demonstrate the compensation techniques, experiments at various optical cable lengths, terminated with an angled physical connector (APC), were conducted.

After employing the fast Fourier transform (FFT) to the signals captured from the reflections, the frequency can be converted to fiber length by \((\text{Eq 6-13})\)(\text{Eq 6-14})\), the results of which are illustrated in \textit{Figure 6-8}. Using the reflections generated by the terminal APCs placed at 3m, 11m, and 20m along the optical fibers, the compensating algorithm can be used to detect the reflected signal via OFDR with increased precision. The sideband noise caused by the $e(t)$ and the corresponding SRs are illustrated in \textit{Table 6-2}.

\textit{Table 6-2} Spatial resolutions in experiments

<table>
<thead>
<tr>
<th>APC position</th>
<th>SR in RDM (cm)</th>
<th>Sideband noise in RDM (dB)</th>
<th>SR in IM (cm)</th>
<th>Sideband noise in IM (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>3m</td>
<td>2</td>
<td>-30</td>
<td>2</td>
<td>&lt;-40</td>
</tr>
<tr>
<td>11m</td>
<td>4</td>
<td>-20</td>
<td>2</td>
<td>&lt;-40</td>
</tr>
<tr>
<td>20m</td>
<td>6</td>
<td>-10</td>
<td>2</td>
<td>&lt;-40</td>
</tr>
</tbody>
</table>

From these results, it is noticed that the $e(t)$ affects the SR in the RDM when the distance to the discontinuity within the fiber under test increases. After compensating the $e(t)$, the SR in the IM is maintained for fiber lengths up to 20m.

A secondary benefit of the system’s flexibility and compact design is a simplified ability to modulate design parameters in order to achieve various frequency-sweep velocities. As mentioned above, the resampling clock provided in the M-Ctrl module limits the maximum OFDR range. In order to extend this range, we modify $\tau_d = 1.025\mu\text{s}$ with all other parameters held constant except for the laser tuning rage which
is 6.5GHz. The results generated after this change, in which the reflections are measured from an APC placed at a distance of 94m along an optical fiber, are illustrated in Figure 6-9 with 5cm SR. Using the RDM, high levels of noise make the reflection position difficult to detect; however, the result of the IM case provides a clear reflection position. These results make clear that the CMP algorithm is required for the OFDR system optimization, especially for long distance measurements.

**Conclusion**

OFDR systems have found significant roles in a growing range of sensing areas. Phase error correction is a critical element allowing for increased precision of these systems; however, the need for additional components and increased complexity has limited the expansion of this technology. This work investigates a multi-function system integrating both swept-laser linearization and phase error compensation. From the experimental results obtained using this technique, it can be concluded that this technique would not only reduces required hardware substantially, but also increases measuring accuracy by suppressing the sideband noise over 20 dB. Additionally, experimental results demonstrate that the IM offered by the proposed system is able to maintain a SR of 2cm for fiber lengths up to 20m with minor degradations at 94m, in
which the SR reduces to 5cm. This technique has the potential to expand the use of
OFDR sensing systems to a wide array of novel applications areas.
Chapter 7 Breaking limitations of fiber ID in traditional OFDR systems via compensation of initial optical frequency instability

by

Zheyi Yao, Thomas Mauldin, Zhenyu Xu, Gerald Hefferman, and Tao Wei

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Abstract

As the security of optical fiber lines in these data centers has attracted growing attention. Optical frequency domain reflectometry (OFDR) has been demonstrated as a means of identifying specific segments of optical fiber; however, OFDR measurements are limited in length due to initial optical frequency (IOF) variations. This letter describes a detailed analysis of IOF, and introduces a method to mitigate it in an OFDR system constructed using a semiconductor laser (SCL). Additionally, an algorithm described which minimizes the calculating density necessary for OFDR-based optical fiber verification, reducing the calculating time required by an order relative to prior techniques. Experiments demonstrate that the described method can be effectively applied to a range of application areas, ranging from centimeter to meter lengths of optical fiber, with an error equal rate (EER) of less than 1%.

Introduction

An increasingly vast amount of digital information is being transferred as audio, video, and text for economic, political, commercial, and military use [57]. As applications such as deep learning, cloud computation, and artificial intelligence continue to expand, the number and utility of data centers can be expected to increase further still [58-60]. Optical fiber is frequently used as a means of data transmission both within and between data centers due to its several unique advantages [61, 62], which include low-cost, high data rate, and relative immunity to electromagnetic interference.

As the number and value of data centers has increased, so too has the need for the data transferred within and between centers to be secured and verified, which led to a
wide range of techniques of securing, such as from radio frequency identification (RFID) [63], biomedical fingerprints [64], and magnetic stripes [65]. Fiber identification (fiber ID) represents the application of physical unclonable function (PUF) [26] technology to the verification of optical fiber, which is inscribed with a unique molecular-level impedance inhomogeneity pattern in manufacturing process [35]. Generally, the impedance inhomogeneity pattern in fiber is referred as Rayleigh backscatter [66].

Recent research regarding a low-cost fiber PUF reader [35] based on a semiconductor laser (SCL) has reduced the cost barrier to expand this technology, making it possible to integrate fiber ID directly into existing optical fiber communication systems and sensing infrastructure with modest overhead. However, even though the technology has shown promising results, there remain limitations that require additional engineering investigation. For example, in previous work, fiber ID achieved high performance only when relatively short optical fiber is under test, failing at longer distances. Even though fiber ID of long sections can be completed by utilizing the combination of short subsets, such additional calculation increases the complexity and cost.

In this letter, an analytical model is first introduced to determine the influence of the initial optical frequency (IOF) variations on the fiber ID system. Additionally, an algorithm is described which mitigates this influence. Next, a universal fiber ID method which aims to reduce computational complexity is described and experimentally evaluated.

**Principle**
In general, there are two subsystems which comprise a compact source and reflectometry measurement systems. To suppress the frequency tuning rate error, an extra compensating technique between the two abovementioned subsystems is utilized, as shown in Figure 7-1. As demonstrated in prior work, a SCL is used as a linearly tuned optical frequency source (Src) which generates a frequency, \( v(t) \), as:

\[
v(t) = \gamma(t) t + v_0
\]

(Eq 7-1)

where \( \gamma(t) \) is the laser optical frequency tuning rate measured in Hz/s, \( t \) is time, and \( v_0 \) is the IOF of the SCL.

Given that the time delay in the multi-control (M-Ctrl) module is \( \tau_d \), the frequency of the BPDs output can be expressed as [36]:

\[
f_{BPD}(t) = v(t) - v(t - \tau_d)
\]

(Eq 7-2)

under the assumption that \( \tau_d \) is so small that \( \gamma(t - \tau_d) \approx \gamma(t) \).
For an ideal linear frequency-swept laser, the frequency tuning rate is a constant value over the tuning period. However, when there is a variation in the environmental temperature or a shift in driving current, a tuning error in the optical frequency is introduced. Taking this into account, the optical frequency tuning rate can be expressed as:

\[ \gamma(t) = \gamma_0 + e(t) \]  
(Eq 7-3)

where \( e(t) \) is the optical frequency tuning error.

In order to minimize the influence of \( e(t) \), an error compensating algorithm is utilized, which is described in prior work [67][50]. Using this technique, the tuning rate of the SCL can be treated as an invariable parameter as indicated by the results in the improved measurement (IM). Furthermore, the optical phase, \( \phi(t) \), of the Src is:

\[ \phi(t) = 2\pi \left[ \int_0^t v(t) dt \right] = 2\pi \left( \frac{1}{2} \gamma_0 t^2 + v_0 t + C \right) \]  
(Eq 7-4)

where \( C \) is an integral constant for the optical phase.

The reflected signal is detected in the IM, after assuming that there is a reflection that corresponds with the time delay, \( \tau = \frac{2R + L_0}{c} \), where \( R \) and \( L_0 \) are the reflecting position and the difference between the circulator (CIR in Figure 7-1) port 2 and the
referencing arm (CPL1 to CPL2), respectively. The phase, $\Phi(t)$, for such an interference signal between the referencing and reflecting optical signals is:

$$
\Phi(t) = \phi(t) - \phi(t - \tau) + \psi_r
\approx 2\pi f_r t + \theta_r
$$

(Eq 7-5)

$$
\theta_r = \text{mod} \left( \frac{2\pi \nu_0}{\gamma_o} f_r + \psi_r, 2\pi \right)
$$

(Eq 7-6)

where $f_r = \gamma_o \tau$ is the interference frequency utilized to capture the reflecting distance, $R$, $\psi_r$ is the phase shifting caused by the reflection, and $\theta_r$ is the interfering phase reminder of $2\pi$ that is the combination of two parts: the time delay $\tau$ at the coefficient of optical initial frequency $\nu_0$, and the reflecting phase shifting $\psi_r$. The 2nd-order component of $-\frac{1}{2} \gamma_o \tau^2$ is so small that it can be neglected; thus, this reflection can be expressed by using the $\psi_r$, which cannot be measured accurately. However, $\theta_r$ can be extracted by using the Hilbert transformation (HT) [68]. After taking the magnitude of the reflection signal into account, the interference signal in time domain can be described as:

$$
S_{IM}(t) = A_r(t) \cos[\Phi(t)]
= \eta_r A_{Sc}(t) \cos(2\pi f_r t + \theta_r)
$$

(Eq 7-7)

where $A_r(t)$ and $\eta_r$ are the amplitude and reflecting coefficient corresponding to the interference signal with time delay $\tau$, which can be calculated from the Hilbert transform as $\eta_r e^{i\theta_r}$. As in previous work, since $f_r$ is sufficient to measure reflector position, the phase information, $\theta_r$, was ignored.
Considering that the reflection signal within an optical fiber is comprised of Rayleigh backscatter that is unintentionally and unavoidably randomly inscribed during its initial manufacture, the corresponding ID sequence, \( Q \), of \( \eta_r e^{j\theta_r} \) can be utilized as:

\[
Q: [\eta_1 e^{j\theta_1}, ..., \eta_i e^{j\theta_i}, ..., \eta_N e^{j\theta_N}]
\]

(Eq 7-8)

where \( N \) is the number of Rayleigh backscatter point sources captured within the fiber.

If assuming a hypothesis test as:

\[
\begin{align*}
\mathcal{H}_0: & \quad Q_{(1)} \text{ and } Q_{(2)} \text{ are from different subsets} \\
\mathcal{H}_1: & \quad Q_{(1)} \text{ and } Q_{(2)} \text{ are from the same subset}
\end{align*}
\]

(Eq 7-9)

where \( \mathcal{H}_0 \) outputs “No”, \( \mathcal{H}_1 \) outputs “Yes”, and the subscript \((1)(2)\) in the equation represent the ID sequence sources, to get the relation between two fibers with the ID sequences, \( Q_{(1)} \) and \( Q_{(2)} \), the maximum value \( P_{(1)(2)} \) in cross-correlation, \( C_{(1)(2)}[i] \), is defined as [35]:

\[
C_{(1)(2)}[i] = \frac{1}{2N-1} \sum_{n=1}^{N} \eta_{(1)n} e^{j\theta_{(1)n}} (\eta_{(2)i+n} e^{j\theta_{(2)i+n}})^* 
\]

(Eq 7-10)

\[
P_{(1)(2)} = \max \{ C_{(1)(2)}[i] \}
\]

(Eq 7-11)

where \( N \) is the number of captured reflecting points and \( i \in (-N, N) \) is the index in the cross-correlation result.

For a genuine fiber (when the reflecting ID sequences at different measurements represent the same properties in the fiber subset), \( P_{(1)(2)} = C_{(1)(2)}[0] \). Thus, the time complexity of the algorithm drops from \( O(N^2) \) in (Eq 7-10)(Eq 7-11) to \( O(N) \) [69] as:
From (Eq 7-6)(Eq 7-10), both the $\eta_r$ and $\theta_r$ affect the result of the optical fiber identification. Compared with the $\eta_r$, which is a permanent prosperity of the fiber, $\theta_r$ is a combination of the IOF, $v_0$, and the Rayleigh backscatter pattern of the fiber. Importantly, the IOF is unstable as illustrated in Figure 7-3, in which the $T$, $B = \gamma_0 T$, $k$, and $v_{(k)0}$ are the tuning period, tuning range, period counting number, and corresponding IOF of the Src, respectively.

To capture precise identification results, an optical initial frequency compensating algorithm is proposed to mitigate influence caused by the initial optical frequencies, $v_{(k)0}$, shifting. As illustrated in Figure 7-2, the beginning of the FUT is utilized to compensate for IOF shifting and is defined as the calibrating element. For the reflection signals in the calibrating element, even though it is impossible to directly measure IOF $v_{(k)0}$, the shift, $\Delta v_{(k)(k+1)0}$, between $v_{(k+1)0}$ and $v_{(k)0}$ can be calculated [70].

Considering the time-shifting property of Fourier transform:

$$\mathcal{F}\{x(t-a)\} = e^{-j2\pi af}X(f)$$

(Eq 7-13)
where \( X(f) \) is the Fourier transformation of \( x(t) \) and \( a \) is the shifting time, the various IOFs introduce linearly increasing phase shifts to the interfering signal. Thus, the time shift for the calibrating element can be applied to compensate for the variations in IOF.

After comparing (Eq 7-13) with (Eq 7-6), the \( v_{(k)0} \) compensation is converted to capture the shifting time \( a \) as the \( \frac{2\pi \Delta v(k)(k+1)0}{\gamma_0} \) can be treated as a coefficient of \( f_r \). For the reflection signal in the time domain from the calibrating element, the time index for the maximum value in the cross-correlation between the two neighboring measuring results represents the time shifting value \( a \). Thus, the subset relationships can be achieved by (Eq 7-13) from the intrinsic Rayleigh backscatter, reducing computational complexity.

**Experiment**

In order to empirically evaluate the accuracy of the proposed frequency model of the Src and the optimized algorithm for verifying the relations of the fiber subsets, experimental testing was undertaken using the parameters and components shown in Table 7-1.

**Table 7-1** Experimental equipment and parameters

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Value or name</th>
</tr>
</thead>
<tbody>
<tr>
<td>DFB-SCL</td>
<td>Eblana EP1550-NLW-B, 1550nm</td>
</tr>
<tr>
<td>( \gamma_0 )</td>
<td>12.6 GHz/s</td>
</tr>
<tr>
<td>( T )</td>
<td>1.2 ms</td>
</tr>
<tr>
<td>( \beta )</td>
<td>15.15 GHz</td>
</tr>
<tr>
<td>( L_0 )</td>
<td>0.11 m</td>
</tr>
<tr>
<td>APC</td>
<td>84 m</td>
</tr>
</tbody>
</table>

While the RDM illustrates that there is a reflection roughly at the APC position, the IM shows the APC position clearly with approximately 3 cm spatial resolution and \(-4 \text{ dB}\) relative magnitude. Considering that the return losses of an APC and Rayleigh
scattering [66] are about 65 dB and 80 dB, respectively, backscatter with over −20 dB relative magnitude can be measured.

The comparison between the analog-to-digital converter (ADC) results using raw data measurement (RDM) versus improved measurement (IM) is depicted in Figure 7-4.

As (Eq 7-6) illustrates, the interference phase reminder, $\theta_r$, captured in the HT is the combination of two parts: Rayleigh backscatter, $\psi_r$, and the IOF, $\frac{2\pi n_0}{\gamma_0} f_r$. There is a

Figure 7-4 The comparison of the RDM and IM results at APC.

Figure 7-5 The performance of the proposed initial optical frequency (IOF)-compensating algorithm between neighboring ($k$), ($k - 1$) measurements: (a) Hilbert angle at ($k$) measurement, (b) phase difference from the Hilbert transformation (HT), (c) phase difference from HT after IOF compensation; (d) statistical distribution of the semiconductor laser (SCL): probability density function (PDF).
linear relationship between fiber distance, $R$, and the time delay, $\tau$, as $\tau = \frac{2R + L_0}{c}$. For a SCL with 1550 nm wavelength, since the optical frequency is over 190THz, it is difficult to measure an accurate $\theta_{(k)\tau}$ which was heavily affected by the optical initial frequency as shown in Figure 7-5(a). However, from the datasheet of the SCL, the initial frequency varies by approximately 100 MHz, which indicates the phase reminder shifting is less than $2\pi$ with $\tau = 10$ ns. Then the shift of the interference phase reminders can be expressed by the difference in the HT angle as Figure 7-5(b) where the slope of the fitting line is caused by the $u_{(k)0} - u_{(k-1)0} = 45.95$ MHz. As for the optical initial frequency compensating algorithm, the shifting time $a$ is equal to 3.68$\mu$s, corresponding to $\gamma_{0a} = 46.48$ MHz. The difference between the two frequencies is from the limited system sampling rate and the fitting accuracy. After time shift processing, the angle difference in the HT is depicted in Figure 7-5(c) in which the fitting line is constant. Thus, the proposed algorithm is sufficient to mitigate the influence of the IOF. Due to the random shifting of the IOF, such neighboring IOF difference $(u_{(k)0} - u_{(k-1)0})$ are expected to follow a Gaussian distribution, as illustrated in Figure 7-5(d), which demonstrates the experimentally observed probability density function (PDF) with mean 0.0082 MHz and standard deviation of 168.922 MHz. The shift of the optical initial frequency should be with a standard deviation of 119.446 MHz, which is at the 100 MHz level as described in SCL datasheet; in other words, this method can be applied to more accurately measure the optical frequency stability of a SCL implemented in an OFDR system.
To demonstrate the performance of (Eq 7-13), proposed to reduce calculation in fiber ID, plenty of experiments were completed. Taking the subset length of 5cm as an example, Figure 7-6 draws the performance of the genuine and imposter results by traditional by (Eq 7-10)(Eq 7-11) and proposed methods by (Eq 7-12) where the x-line illustrates the relative peak values of the genuine and imposter sequences, $Q_{(1)(2)}$ from the $\mathcal{H}_0$ and $\mathcal{H}_1$ in Figure 7-6(a)(b). As for the accuracies illustrated by the EERs, compared with the traditional method (0.12%), the proposed algorithm is with greater EER rate (0.37%); however, the averaging time in the two methods are 251.1μs and 21.2μs respectively for each identifying process. Thus, the proposed method has reduced the calculating time over 1 order of magnitude without sacrificing much accuracy.

As mentioned in the principle section, both the reflecting coefficient, $\eta_r$, and phase shift, $\psi_r$, from the intrinsic Rayleigh backscatter pattern can be utilized as a means of fiber ID. To verify this principle, various experiments were undertaken with subset lengths of 3 cm, 10 cm, 1m, 10 m, and 20 m; the results of which are depicted in Figure 7-7 and Table 7-2.
Generally, a robust identification system is expected to achieve an EER of less than 1%. Taking the results from the traditional method as an example, neither the 3 cm nor over 1 m results can accomplish the robust EER less than 1%. The reasons behind such phenomenon are different: the 3 cm result is caused by insufficient characteristic information in the small subsets, while the results for 1 m and beyond are from the influence of the initial optical frequencies. This conclusion is safely reached from the fact that the reflecting coefficient only method, in which only the $\eta_r$ is utilized to represent the subsets, works for the over 1 m subsets shown in Figure 7-7(b). As for the proposed method, with results illustrated in Figure 7-7(c), it is clear that both cm and m level experiments result in the robust performances with EERs less than 1%, except the 3 cm experiment. Thus, the proposed method is universal for the fiber identifying system. Even though optical fiber is utilized in hyper-scale data centers with lengths on the order of 100 m, the proposed algorithm shows the potential to efficiently mitigate against the risk of physical attack. Furthermore, the cm level optical fiber ID can be

<table>
<thead>
<tr>
<th>Subset length</th>
<th>Traditional method EER (%)</th>
<th>Reflecting coefficient only method EER (%)</th>
<th>Proposed method EER (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>3cm</td>
<td>6.52</td>
<td>38.82</td>
<td>9.43</td>
</tr>
<tr>
<td>10cm</td>
<td>0.12</td>
<td>23.79</td>
<td>0.36</td>
</tr>
<tr>
<td>1m</td>
<td>12.82</td>
<td>&lt; 0.1</td>
<td>&lt; 0.1</td>
</tr>
<tr>
<td>10m</td>
<td>38.62</td>
<td>&lt; 0.1</td>
<td>0.33</td>
</tr>
<tr>
<td>20m</td>
<td>46.55</td>
<td>0.2</td>
<td>0.63</td>
</tr>
</tbody>
</table>

*Figure 7-7* The ROC curves for various experiments: (a) traditional method; (b) reflecting coefficient applied only method; (c) proposed method.
inscribed in the devices such as secure tags in order to maintain the security as described in prior work [35].

Conclusion

To summarize, this work demonstrates and experimentally evaluates a system that directly measures the stability of the semiconductor laser without the need for complex and expensive equipment, and proposes an algorithm to decrease the calculating complexity by more than one order of magnitude in the processing time in 5 cm subsets of fiber without sacrificing accuracy. While the core of the OFDR system is the constant tuning rate of the Src, the IOF plays an important role when precise results are required. Thus, the described IOF identifying method surmounts the limitations of prior OFDR systems used to measure PUFs, and further broadens its application to other areas, including optical fiber sensors, laser spectroscopy, and SCL performance characterization.
List of References


[38] D. Huang et al., "High-power sub-kHz linewidth lasers fully integrated on silicon," *Optica*, vol. 6, no. 6, pp. 745-752, 2019.


