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Silicon-Chip-Based Real-Time Dispersion Monitoring for 640 Gbit/s DPSK Signals

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Abstract—We demonstrate silicon-chip-based instantaneous chromatic dispersion monitoring (GVD) for an ultrahigh bandwidth 640 Gbit/s differential phase-shift keying (DPSK) signal. This monitoring scheme is based on cross-phase modulation in a highly nonlinear silicon nanowire. We show that two-photon absorption and free-carrier-related effects do not compromise the GVD monitoring performance, making our scheme a reliable on-chip CMOS-compatible, all-optical, and real-time impairment monitoring approach for up to Terabit/s DPSK signals.

Index Terms—Nonlinear optics, optical performance monitoring (OPM), optical planar waveguides, optical signal processing, spectral analysis.

I. INTRODUCTION

Nonlinear silicon photonics has attracted significant attention in recent years due to major advances in nanofabrication enabling a wide range of applications on integrated silicon-on-insulator (SOI) photonic circuits [1]. Another major motivation behind the interest in nonlinear photonic nanowires is the improved energy efficiency of nonlinear processes due to the significant enhancement of the nonlinear response $\gamma = \cdot 2\pi n_2 / \lambda A_{\text{eff}}$ (where n_2 and A_{eff} are nonlinear index and effective mode area, respectively), which is up to 5 orders of magnitude larger than in silica fiber [2]. These advantages make silicon an attractive nonlinear platform for on-chip all-optical signal processing, and there have been numerous impressive demonstrations, including all-optical switches [3], format con-

version [2], tunable delay and optical phase conjugation [4], [5], time-division demultiplexing [6], [7], ultrafast waveform measurement [8], wavelength conversion [9], [10], Raman laser [11], [12], light amplification [13], [14], and optical impairment monitoring via slow-light enhanced third harmonic generation [15].

Recently, we have exploited cross-phase modulation (XPM) [16] in a nonlinear waveguide to capture the radio frequency (RF) spectrum of an ultrahigh bandwidth optical signal [17]. This scheme has been used to characterize terabit/s data [18], [19] as well as multi-impairment monitoring of high-bit-rate intensity-encoded [20], [21] and phase-encoded signals [22]. Our initial experiments were reported in chalcogenide glass photonic integrated circuits. Besides chalcogenide, silicon would be a natural platform to consider for this scheme because of the mature CMOS fabrication process; however, silicon suffers from relatively high two-photon absorption (TPA), photogenerated free-carrier absorption (FCA) and photogenerated free-carrier dispersion (FCD) when operating at telecommunication wavelengths [23]–[25]. The TPA and FCA are nonlinear losses, which reduce the efficiency of the XPM process, while FCD modifies the material refractive index and can distort the modulation caused by the Kerr nonlinearity [26], [27]. This may result in reduced device efficiency and a cross-chirp that can distort the Kerr-based XPM and lead to errors in the RF spectrum in our optical performance monitoring (OPM) method. We recently demonstrated that TPA and free-carrier effects are insignificant at the power used for the RF spectrum monitoring of 640 Gbit/s ON–OFF keying signals [19], [28].

In this paper, we report on the first demonstration of a silicon-chip-based OPM scheme to instantaneously monitor dispersion of ultrahigh bandwidth phase-encoded signals. We exploit the ultrafast effect of XPM in a CMOS compatible silicon nanowire, in combination with a simple optical bandpass filter (BPF) and a relatively slow ($\sim \mu\text{s}$ response time) power meter to continuously monitor the group velocity dispersion (GVD) for a 640 Gbit/s return-to-zero differential phase-shift keying (RZ-DPSK) signal. As the fluctuations of GVD in optical communication links typically occur on times scales much greater than microseconds, this scheme allows for real-time monitoring of GVD. We numerically investigate the impact of TPA and free-carrier related effects and demonstrate that they do not affect performance of GVD monitoring. The potential integration of electronics with our CMOS compatible device offers the possibility for the creation of a terahertz bandwidth [28], low-cost, chip-based, real-time OPM device.

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Fig. 1. Conceptual diagram of silicon-chip-based instantaneous GVD monitoring. This scheme is suitable to be operated at nodes, after in-band EDFA in reconfigurable networks.

II. WORKING PRINCIPLE

Fig. 1 shows a schematic of all-optical performance monitoring using a silicon-chip-based RF spectrum analyzer. When the signal under test (SUT) and a continuous wave (CW) probe are copropagated through the nonlinear waveguide, the probe's phase is modulated via XPM [16]. The broadened spectrum around the probe frequency is proportional to the power spectrum of the input signal intensity waveform if the maximum cross phase-shift on the probe $\phi_{\max} = 2\gamma P_0 L_{\text{eff}} < 1$, where P_0 and L_{eff} are the signal peak power and effective length of a waveguide, respectively [29]. Note that this monitoring approach is suited for operation at nodes [after in-line Erbium-doped fiber amplifiers (EDFAs)] of reconfigurable optical communication networks to obtain sufficient input power.

The key device used in the experiment is a 1.5 cm long SOI nanowire. The waveguide is 450 nm wide by 260 nm thick, producing an effective transverse electric (TE) mode area $A_{\text{eff}} \sim 0.15 \mu\text{m}^2$, as extracted from finite element method mode solver (RSoft FEMsim) calculated using the definition provided in [16]. The nonlinear refractive index n_2 of the silicon waveguide is taken as $6 \times 10^{-18} \text{m}^2/\text{W}$, yielding a nonlinear coefficient γ of $\sim 160 \text{W}^{-1}\text{m}^{-1}$ at 1550 nm. We couple to the TE mode from lensed fibers (2.5 μm spot size) via inverse tapers. Coupling and propagation losses of this silicon waveguide are $\sim 8 \text{dB/facet}$ and $\sim 3 \text{dB/cm}$, respectively.

III. TPA AND FREE-CARRIER EFFECTS

In silicon, intensity-dependent nonlinear effects become more complicated due to TPA and TPA-induced free-carrier effects [26], [27]. The input peak power and pulsewidth of a signal are the determinant factors that decide the impact of TPA and free-carrier effects on performance of the device. We

numerically investigate the effects of TPA and free carriers by simulating this system using the nonlinear Schrödinger equation, given by [10], [26], [30]–[36]

$$\begin{aligned} \frac{\partial A_s}{\partial z} + \frac{\alpha}{2} A_s + \beta_{1s} \frac{\partial A_s}{\partial t} + j \frac{\beta_{2s}}{2} \frac{\partial^2 A_s}{\partial t^2} - \frac{\beta_{3s}}{6} \frac{\partial^3 A_s}{\partial t^3} \\ = j (\gamma_{ss} |A_s|^2 + 2\gamma_{sp} |A_p|^2) A_s \\ - \frac{1}{2A_{\text{eff}}} (\beta_{\text{TPA-ss}} |A_s|^2 + 2\beta_{\text{TPA-sp}} |A_p|^2) A_s \\ - N_c \frac{\sigma_s}{2} A_s - j N_c \frac{2\pi k_c}{\lambda_s} A_s \end{aligned} \quad (1a)$$

$$\begin{aligned} \frac{\partial A_p}{\partial z} + \frac{\alpha}{2} A_p + \beta_{1p} \frac{\partial A_p}{\partial t} + j \frac{\beta_{2p}}{2} \frac{\partial^2 A_p}{\partial t^2} - \frac{\beta_{3p}}{6} \frac{\partial^3 A_p}{\partial t^3} \\ = j (\gamma_{pp} |A_p|^2 + 2\gamma_{ps} |A_s|^2) A_p \\ - \frac{1}{2A_{\text{eff}}} (\beta_{\text{TPA-pp}} |A_p|^2 + 2\beta_{\text{TPA-ps}} |A_s|^2) A_p \\ - N_c \frac{\sigma_p}{2} A_p - j N_c \frac{2\pi k_c}{\lambda_p} A_p \end{aligned} \quad (1b)$$

where $A_{s,p}$ is the slowly varying amplitude of the pulse envelope of the signal s under test and the CW probe p , α is the linear loss coefficient, $\beta_{1s,p}$, $\beta_{2s,p}$, and $\beta_{3s,p}$ are the linear, second- and third-order dispersion parameters, respectively. The nonlinear coefficient γ_{ss} , γ_{sp} , γ_{ps} , γ_{pp} are, strictly speaking, different from each other, but the difference is small [16], [30]. Therefore, the four parameters are approximately $\gamma_s = 2\pi n_2 / \lambda_s \cdot A_{\text{eff}}$. $\beta_{\text{TPA-ps}}$ and $\beta_{\text{TPA-pp}}$ are the TPA coefficients at frequencies f_s and f_p , respectively; and $\beta_{\text{TPA-sp}}$ and $\beta_{\text{TPA-sp}}$ are the cross-TPA coefficients. These four terms are approximately equal to $\beta_{\text{TPA}} \approx 5 \times 10^{-12} \text{m/W}$. σ_s and σ_p are the FCA coefficients which are defined as [33] $\sigma_{s,p} = \sigma \cdot (\lambda_{s,p} / \lambda_r)^2 \approx 1.45 \times 10^{-21} \text{m}^2$, where $\lambda_r = 1550 \text{nm}$ is the reference wavelength. We take $k_c \approx 1.35 \times 10^{-27} \text{m}^3$ as the FCD coefficient. N_c is the free-carrier density generated by TPA, which is calculated as [26], [30], [33]

$$\begin{aligned} \frac{\partial N_c}{\partial t} = \frac{\beta_{\text{TPA}}}{2h\nu_s} |A_s|^4 + \frac{\beta_{\text{TPA}}}{2h\nu_p} |A_p|^4 \\ + \frac{2\beta_{\text{TPA}}}{h\nu} |A_s|^2 |A_p|^2 - \frac{N_c}{\tau_c} \end{aligned} \quad (2)$$

where the free-carrier lifetime τ_c is chosen to be 1 ns, which is the worst case estimate of the carrier lifetime in our waveguide [10], [37], [38].

Fig. 2(a) shows the numerical (solid curves) and experimental (dots) results of average output power versus input power for a 640 Gbit/s DPSK signal traveling through our waveguide. With low input power, TPA has a negligible impact and the generated free-carrier density N_c is low enough that free-carrier effects, including FCA and FCD, can be neglected. Therefore, the output power linearly increases with increasing input power. As the input power increases beyond $\sim 300 \text{mW}$, TPA and associated free-carrier effects become more severe and nonlinear losses lead to a clamping effect. By analyzing the accumulated cross-phase shift, we conclude that the maximum average input signal power should be kept below $\sim 110 \text{mW}$ (corresponding to a maximum nonlinear phase shift $\phi_{\max} = 1$) to satisfy the

Fig. 2. (a) Numerical (curves) and experimental (dots) results showing power transfer functions of a 640 Gbit/s DPSK signal. Frequency chirp of a 640 Gbit/s signal on the probe when a steady-state of the carrier density is reached with (b) $P_{ave} = 2$ W and (c) $P_{ave} = 110$ mW.

$\phi_{max} < 1$ condition. Within this operating regime, we find that TPA and free-carrier effects are negligible.

To illustrate the impact of cross-chirp induced by free-carrier effects on the captured RF spectrum, we plotted the calculated frequency chirp generated by a DPSK signal on the CW probe when the steady state of the carrier density is reached at input powers $P_{ave} = 2$ W [see Fig. 2(b)] and 110 mW [see Fig. 2(c)]. At the leading edge of the pulse at 2 W input power, Kerr-based XPM generates red-shifted frequencies while TPA and free carriers create an opposite phase shift, thus reducing the amount of Kerr nonlinear phase shift on the probe. At the trailing edge of the pulse, both Kerr- and free-carrier-based phase shifts generate blue-shifted frequencies on the probe, hence producing an asymmetrical spectral broadening shifted toward shorter wavelengths [see Fig. 2(b)], compromising the RF spectrum measurements. On the other hand, for $P_{ave} \leq 110$ mW, the free-carrier-induced cross-chirp on the probe is insignificant compared

Fig. 3. (a) Experimental setup of silicon-chip-based real-time chromatic dispersion monitoring for 640 Gbit/s DPSK signals. (b) Eye diagram. (c) Optical spectrum of a 640 Gbit/s signal.

to Kerr-based XPM as shown in Fig. 2(c). As such, TPA-induced free-carrier effects are negligible at the operating power required by this monitoring technique.

IV. EXPERIMENTAL SETUP AND RESULTS

Fig. 3(a) shows the experimental setup for real-time GVD monitoring for a 640 Gbit/s RZ-DPSK signal. A 40 GHz pulse train ($\lambda_s = 1548$ nm) from a mode-locked fiber laser with a pulsewidth (after a nonlinear pulse compression) of ~ 550 fs was data encoded by a Mach-Zehnder modulator, driven by a 40 Gbit/s pseudorandom bit sequence of $2^{31} - 1$ pattern length to produce a 40 Gbit/s RZ-DPSK signal. Four-stage optical time division multiplexing was used to generate a 640 Gbit/s RZ-DPSK signal, whose eye diagram and optical spectrum are shown in Fig. 3(b) and (c), respectively. We introduced a precise, known amount of chromatic dispersion via a programmable spectral pulse shaper (Finisar WaveShaper) [39]. The distorted signal ($P_{ave} \sim 70$ mW at the input or ~ 11 mW inside the waveguide, well below the previously determined maximum power condition) and a CW probe ($\lambda_p = 1565$ nm, $P_{ave} \sim 30$ mW, or ~ 4.7 mW inside the waveguide) were copropagated in the silicon waveguide. Polarization controllers were used to ensure coupling light to the TE mode of the chip. A second spectral pulse shaper, acting as a sharp optical filter centered at 1567.5 nm with 3 nm filtering bandwidth, and a slow power meter were used at the output of the waveguide to perform the real-time (microsecond measurement time) GVD monitoring.

Fig. 4(a) presents the RF spectra captured via this silicon-chip-based performance monitoring technique. The RF spectrum of a 640 Gbit/s RZ-DPSK signal comprises a strong fundamental 640 GHz clock tone and weaker tones at multiples of the bit-rate due to the pulsed nature of RZ signals. Note that the RF power is comparatively small for other frequencies. In the presence of GVD, the constructive and destructive interference between adjacent pulses distorts the temporal intensity of the SUT. This leads to intensity ripples in the time domain, corresponding to new frequencies generated in the RF spectrum

Fig. 4. (a) RF spectra of a 640 Gbit/s DPSK signal (with and without dispersion) captured via chip-based RF spectrum analyzer. These curves are displaced vertically by 20 dB for clarity. (b) Experimental. (c) Numerical results of the GVD measurements of a 640 Gbit/s DPSK signal.

around the clock tone. Therefore, the RF power of the frequencies between the CW probe and the fundamental clock tone increases gradually with increasing GVD [22], [40].

Fig. 4(b) and (c) shows the experimental and numerical results of the optical power versus applied GVD, respectively. The GVD monitoring range was determined to be ~ 0.7 ps/nm. Note that the current dynamic range of this monitoring was limited to ~ 6 dB [see Fig. 4(b)] in our experiment compared to ~ 10 dB dynamic range in the simulation results [see Fig. 4(c)] due to the limited sensitivity of the power meter used in this paper. The experimental measurement dynamic range could be improved with a higher sensitivity power meter and lower loss optical filters. The error bars in Fig. 4(b) were due to power fluctuations during the measurements.

V. DISCUSSIONS

We numerically investigate the GVD monitoring accuracy in the presence of TPA and free carriers while varying input power P_{ave} . As discussed in Section III, the contribution of TPA and free-carrier-related effects to the new frequencies generation around the CW probe becomes significant with high input power. This directly distorts the RF spectrum captured via

Fig. 5. (a) Numerical results showing the GVD monitoring errors due to TPA and TPA-induced free carriers compared to pure Kerr nonlinearity while varying input powers. (b) Example of a comparison between (solid line) true GVD and (dots) captured GVD with different input power.

our chip-based radio frequency system automation approach, hence providing inaccurate GVD monitoring. Fig. 5(a) illustrates the GVD measurement errors due to TPA and free-carrier effects compared to pure Kerr nonlinearity. The monitoring errors are defined as $\text{errors} = P_{\text{Kerr}} - P_{\text{Kerr_TPA_FC}}$, where P is optical power of new frequencies generated between the CW probe and fundamental tone at 640 GHz with the subscripts Kerr and Kerr_TPA_FC indicating simulations with the Kerr effect only and with Kerr effect, TPA, and free-carrier-associated effects, respectively. We observe that with an average power $P_{\text{ave}} \leq 110$ mW, the measurement errors are insignificant due to the relatively small free-carrier-induced cross-chirp on the probe compared to Kerr-based XPM. The result confirms the high measurement accuracy of our silicon-chip-based monitoring approach if the upper limit of the operating power for the condition of $\phi_{\text{max}} < 1$ is satisfied.

Fig. 5(b) illustrates the changing accuracy of the GVD monitoring with varied signal average power inside our silicon waveguide. Here, we compared the measured GVD to the true GVD values with different input powers to investigate the measurement errors owing to TPA and free-carrier effects. First, the S-shape GVD monitoring curve [see Fig. 4(c)] is simulated without TPA and free-carrier effects as reference values. The simulation is then repeated in the presence of TPA and free carriers. The optical power between the CW probe and an optical clock tone is captured and compared with the reference GVD monitoring curve for determining the GVD value within the system. At low input power (i.e., $P_{\text{ave}} = 11$ and 110 mW), a good agreement between measured GVD and true

GVD values is demonstrated with a measurement error of less than 3%. However, the measurement error becomes significant with higher input power ($P_{\text{ave}} = 2$ W). These numerical results agree well with our surmise that free-carrier effects will distort our measurements. The error bars in Fig. 5(b) are calculated with an assumption that the power meter has 0.5 dB power fluctuation.

We should note that this silicon-chip-based GVD monitoring method also works well for nonreturn-to-zero DPSK format [22]. In this paper, we only show GVD monitoring for 640 Gbit/s RZ-DPSK because we would like to highlight the broadband operation capability and high measurement accuracy (distortions of the RF spectrum of a SUT due to free-carrier-induced cross-chirp on the probe is negligible) of this approach.

The measurement dynamic range of our chip-based monitoring approach is relatively low compared to the results from [40] while similar input powers were used in both experiments. This is because the $(\gamma \cdot L)$ product of our silicon nanowire is ~ 9 times lower than that of the 2 km highly nonlinear fiber (HNLF) used in [40], thus resulting in significant less nonlinear phase shift in the propagation. On the other hand, the high dispersion $(D \cdot L)$ of a HNLF introduces large walk-off between an SUT and a CW probe, yielding much narrower monitoring bandwidth of ~ 200 GHz (numerical result), without considering the dispersion fluctuation along the fiber, compared to 1.6 THz bandwidth [28] of our monitoring scheme.

VI. CONCLUSION

We have experimentally demonstrated instantaneous GVD monitoring for ultrahigh bandwidth 640 Gbit/s DPSK signals based on XPM in a silicon chip. Our investigation shows that nonlinear losses due to TPA and TPA-induced free carriers do not affect our monitoring performance. The numerical studies also provided the operating power limit for our silicon-chip-based OPM. This scheme is a compact and potentially cost-effective OPM approach with terahertz monitoring bandwidth, requiring only a CMOS-compatible SOI waveguide, an optical BPF, and a relatively slow power meter, all of which can be integrated on a chip.

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