Ultra-Compact Non-Travelling-Wave Silicon Carrier-Depletion Mach-Zehnder Modulators Towards High Channel Density Integration

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Abstract—Silicon Mach-Zehnder modulators (MZMs) utilizing carrier depletion usually adopt traveling-wave (TW) electrodes for their long phase shifters, which have inherent challenges in miniaturization, limiting the integration density of monolithic transceivers. In this work, we experimentally demonstrate non-TW carrier-depletion MZMs that give better compromises between compact footprint, modulation depth, and high speed. They adopt densely meandered phase shifters and thus offer an ultra-compact lumped-element MZM for using carrier depletion. Up to 28.1 Gb/s NRZ eye diagrams and ~26 GHz electro-optic bandwidth were experimentally achieved. The proposed lumped circuit model well explains the experimental data, indicating that impedance mismatching is the dominant factor in determining experimental bandwidth and a LC resonance contributes to the bandwidth improvement. Driving configurations with better impedance matching are proposed to further improve bandwidth. This novel MZM can enhance the integration density of modulators in wavelength-division multiplexing transceivers.

Index Terms—Electrooptic modulators, Silicon photonics.

I. INTRODUCTION

OPTICAL transceivers are continuously miniaturized for low-cost, low-power, and high-density optical interconnects [1]. Silicon photonics are believed as a leading technology to realize such transceivers due to its integration compatibility and will be more competitive than ever when co-packaging or monolithically integrating with electrical circuits becomes available [2]–[4]. On the one hand, integrating silicon photonics with electronics is becoming a disruptive hardware solution to build ultra-dense high-complexity electro-optic circuits. On the other hand, multiple wavelength lines of 400G and beyond are required for future inter/intra-data center interconnects to satisfy continuous increases in transmission capacity. If adopting dense wavelength-division multiplexing, several tens or even more wavelength channels will be required. Therefore, it is very important to keep silicon photonic transceivers as compact as possible from the views of integration density, easy packaging, low power, and low cost.

Most components in silicon photonic transceivers can be made with small footprints, while Mach-Zehnder modulators (MZMs) based on carrier depletion are of large sizes. They are usually several millimeters (mm) long [5], occupying large chip areas and limiting miniaturization of transceivers. There are compact silicon modulators such as ring modulators [6], slow-light modulators [7], [8], and MZMs using carrier injection [9], [10]; whereas MZMs using carrier depletion have advantages in the wide spectral bandwidth, high speed, and low dispersion so that they remain as a competitive choice in commercialized silicon photonic transceivers [1], [11]. But due to the low modulation efficiency, they usually need long phase shifters, for which traveling-wave (TW) electrodes in form of coplanar waveguides are one of the most common electrode choices (traveling-wave driving) [5]. Another choice is to adopt distributed electrodes (multiple lumped element driving) [12]. Both choices have similar large footprints. For the former, it is difficult to obtain impedance and radio frequency (RF)-optical velocity matching simultaneously over a broad band [13], [14]. If adopting slow-wave electrode structures [15], [16], better velocity matching can be achieved. However, the RF loss is unavoidable along several-mm-long TW electrodes loaded with pn junction capacitance and quickly increases with the phase shifter length and frequency, which is believed as an inherent limiting factor of bandwidth [13], [14], [17]. This also means that further increasing the electrode length is not so effective to improve the high frequency modulation efficiency and velocity mismatching will become more severe as well. For the latter, the phase shifter is divided into multiple small segments and each segment can be driven at very fast speed by an individual RF signal, thus it is free of the RF-loss problem. The driver is of high complexity because there are many RF signal channels and RF-delay adjustment between channels is required to compensate optical delay in each segment. Silicon TW-MZMs based on carrier depletion have encountered bottlenecks in breaking through bandwidth and efficiency limitations. Recently advanced modulation formats with high baud rates (>70G) have been demonstrated using low-bandwidth TW-MZMs (<40 GHz in [18], [19], ~47 GHz in [20]). It becomes more and more difficult for silicon modulators to compete on baud rates with other modulators such as lithium niobate [21], [22] and InP modulators [23] due to intrinsic material limitation. Tremendous efforts have been spent...
on hybrid integrated modulators on silicon using non-CMOS (complementary metal oxide semiconductor) process to seek breakthroughs, while all-silicon modulators remain competitive in large-scale integration, low cost, and high-volume production by leveraging the fully CMOS-compatible process, which are believed as the core advantages of silicon photonic transceivers. Therefore, silicon photonic transceivers will more and more rely on enhancing integration density (i.e., channel parallelization of wavelength or mode) of modulators to increase overall link capacity to compete with those on other platforms. The carrier-depletion MZMs have been extensively studied in the past as an important all-silicon modulator, while its miniaturization remains challenging compared to ring modulators and those using carrier injection and slow light.

Decreasing the phase shifter length of TW-MZMs is better for lowering the RF loss; but the modulation amplitude will also be decreased. Thus, the problem is how to compromise a small MZM footprint to make the RF loss negligible and a reasonable phase shifter length to guarantee necessary modulation amplitude. For this purpose, the lumped-element MZM could be a potential choice, for which we could adopt a single lumped-element driving instead of travelling-wave and distributed-electrode driving to keep the driver simple. In [14], using meandered phase shifter to realize lumped-element MZMs was first proposed and simulated for silicon epitaxially-grown diodes loaded with a comb-like electrode. In [24], using the same concept as that in [14], we experimentally demonstrated a non-travelling-wave (NTW) MZM (NTW-MZM) by employing a densely meandered phase shifter, which for the first time reported the experimental results of a lumped-element MZM modulator based on carrier depletion without utilizing slow-light effects [7], [8], as far as we know. Recently, in [25], a lumped-element MZM using an electrode structure different from that modeled in [14] was experimentally presented with about 8–12 GHz bandwidths within 2–5 V biases. These NTW-MZMs have quite compact footprint. As for our case, it has an ultra-compact footprint (<500 \( \mu \)m in width and <250 \( \mu \)m in length), which is almost equivalent to the electrical pad size of TW-MZMs, only 1/8~1/16 of a usual TW-MZM, and also smaller than those of distributed [26] and multi-segmented electrode types [12]. Due to the small area, RF driving signals have no attenuation along the phase shifter, which can relatively improve the modulation efficiency for shorter phase shifters. However, compared to TW-MZMs, the high frequency bandwidth is still low. For optimizing such lumped-element MZMs, we examine the transit time effects on eye diagrams and high frequency responses under various diode conditions and perform bandwidth comparison with TW-MZMs. We also study the impedance matching with considering the effect of LC resonance on bandwidth since in our experiment we found the inductance of comb-like electrodes can contribute to bandwidth enhancement via inducing a LC resonance. These two aspects have not been fully studied yet, especially for the conventional CMOS-compatible horizontal and vertical \( pn \) diodes.

In this paper, we present the updated results of NTW-MZMs with bandwidth much improved compared to previous works [14], [25] and our initial work [24]. A comprehensive study on NTW-MZMs is done from aspects of design, simulation, and experiment. The paper is organized as follows. Section II presents fundamental issues of NTW-MZM, including modulator design, transit time determined bandwidth, and what \( pn \) junctions can fit better into NTW-MZMs than TW-MZMs. The effects of transit time on eye diagrams and frequency responses under various diode conditions are discussed and comparison with TW-MZMs are presented. Section III presents the fabricated devices, experimental results, circuit models, and discusses the impedance matching and contributions of LC resonance in enhancing bandwidths. Finally, Section IV concludes the paper.

II. DESIGN AND SIMULATION

A. Modulator Design

Figs. 1(a) and 1(b) show the schematic structure and layout of the proposed silicon NTW-MZM, respectively. This device can be fabricated on photonic silicon-on-insulator (SOI) wafers with a SOI thickness of 200~300 nm. It is a MZ interferometer structure composed of two multimode interference couplers and two arms. The feature lies in that the arms are consisted of multiple parallel segments of straight rib waveguides (in slab region) which are connected at both ends by the fully etched Si-waveguides through the low-loss rib-channel conversion structures. The \( pn \) junctions are embedded in these rib waveguides as the phase shifters. Both horizontal and vertical \( pn \) junctions [27] are applicable. Two adjacent segments of rib waveguides are close to each other to share a middle doping region. The segment length \( L_1 \) and distance \( d_s \) of rib waveguide segments are 140.5 \( \mu \)m and 11.3 \( \mu \)m, respectively. So, the total length of phase shifter is 1.124 mm for each arm using eight segments. For such a phase shifter, we have designed two kinds of MZMs with different electrode designs. Our initial design shown in Fig. 1 was presented in [24] where the contact holes to the doping regions are alternatively put at the center and two ends of rib waveguides for accessing the signal (S) and ground (G) electrodes, respectively. This design is 490 \( \mu \)m in width and 150 \( \mu \)m in length, nearly equivalent to the electrode
pad size of TW-MZMs. Our improved design uses comb-like electrode, which was also fabricated and will be discussed in Section III. With such a phase shifter design, we can realize a lumped-element carrier-depletion MZM with both a small footprint and a reasonable length of phase shifter. Note that the $S$ electrodes are connecting to the $n$-side of diodes.

### B. Transit Time Determined Bandwidth

The bandwidth of TW-MZMs cannot reach the resistance-capacitance (RC) bandwidth of diodes due to the factors such as RF loss, RF-optical velocity mismatching, and impedance mismatching. For the lumped-element NTW-MZM, the bandwidth is determined by optical transit time corresponding to the phase shifter length and impedance mismatching. In this section, perfect impedance matching condition is assumed, and we only consider the phase modulation induced by carrier depletion upon various diode parameters. Thus, the bandwidth discussed in this section can be regarded as the theoretical limit of NTW-MZM. The influence of impedance matching will be discussed in the experimental part in Section III.

The transit time determined bandwidth comes from the mismatch between the RF frequency and the transit time for light passing through the phase shifter, which is different from the RF-optical velocity mismatching in TW-MZMs because it is not related to the RF effective index. Here we investigate the bandwidth with considering both RC parameters and transit time. For this purpose, we iteratively solve the well-known Kirchhoff’s current equation (Eq. (1)) and an integral equation (Eq. (2)) of phase shift $\Delta \phi$ in time domain. $V_s$ and $V_c$ are the RF source voltage and the actual voltage on junction capacitor that causes the optical phase change, respectively. $R_{pn}$ and $C_{pn}$ denote the series resistance and capacitance of the $pn$ diodes, respectively. $c_0$, $n_{pn}$, and $\eta$ are the light speed in vacuum, group index of silicon rib waveguide, and modulation efficiency in unit of radian/(V·μm), respectively. $\tau_0$ is the transit time for light passing through the phase shifter at its group velocity. $V_s$ can be sinusoidal wave $V_{pp} \sin(2\pi ft)/2$ for frequency response simulation or pseudo-random bit sequence (PRBS) for eye diagram simulation ($V_{pp}$ is the peak-to-peak driving voltage).

Given a phase shifter length that has a corresponding $\tau_0$, we can calculate the frequency response $r(f) = 20 \log |\Delta A/\Delta \phi_{mn}(f)|/\Delta A/\Delta \phi_m(0)|$ for each frequency $f$, where $\Delta A$ is the optical modulation amplitude of MZM upon a phase modulation $\Delta \phi_m$.

$$C_{pn} \frac{dV_c}{dt} = \frac{V_s - V_c}{R_{pn}}$$

$$\Delta \phi(t) = \int_{t-\tau_0}^{t} \eta c_0 V_c(t) \frac{dt}{n_g}$$

Before simulation, we determine $R_{pn}$, $C_{pn}$, and $\eta$ for three diode conditions to be studied: a horizontal $pn$ junction (hPN) with a doping density of $1 \times 10^{18}$ cm$^{-3}$ and two vertical $pn$ junctions (vPN) with doping densities of $1 \times 10^{18}$ and $2 \times 10^{18}$ cm$^{-3}$. Fig. 2(a) shows the schematics of the hPN and vPN, both of which are usually used in TW-MZMs [2], [27]. They are assumed to be fabricated on 220-nm SOI wafers and have the same waveguide parameters (rib width = 600 nm, slab thickness = 110 nm). The junction positions are at the horizontal and vertical centers of SOI layers for the hPN and vPN, respectively. Uniform doping is assumed for simplicity and $p$ and $n$ concentrations are equal. Using Lumerical simulators [28], we calculated $C_{pn}$ and phase change $\Delta \phi$ in relation of the reverse bias ($V_b$), as shown in Figs. 2(b) and 2(c). To get the continuous equations of $C(V_b)$ and $\Delta \phi(V_b)$, the simulated $C_{pn}$ and $\Delta \phi$ were fitted by the Eqs. (3) and (4), respectively, in which we used a building-in potential of 0.75 V. The fitting results are shown by the black lines in Figs. 2(b) and 2(c) and the corresponding fitting parameters are summarized in Table I.

![Fig. 2. (a) Horizontal pn (hPN) and vertical pn (vPN) junctions used in simulation. Bias dependences of (b) capacitance and (c) phase change. The black lines are fitting lines using Eqs. (3) and (4), respectively. M: metal.](attachment:image.png)

**TABLE I**

| $N$ cm$^{-3}$ | $\Delta \phi(V_b)$ | $C(V_b)$ |
|----------|------------------|---------|
|          | $a$   | $b$     | $c$   |
|          | $a'$  | $b'$   | $c'$  |
| hPN      | $1 \times 10^{18}$ | 0.014   | 270.20 | $-7.45 \times 10^{-4}$ |
|          | 0.020 | 149.13 | $-1.35 \times 10^{3}$ |
| vPN      | $2 \times 10^{18}$ | 0.023   | 100.15 | $-1.95 \times 10^{-3}$ |
|          | 120.67 | 1.36 $\times 10^{-5}$ | 0.069 |
| vPN      | $1 \times 10^{18}$ | 253.39 | 2.24 $\times 10^{-5}$ | 0.038 |
| vPN      | $2 \times 10^{18}$ | 272.88 | 3.55 $\times 10^{-5}$ | 0.083 |
Fig. 3. (a) Frequency dependent phase shift $\Delta \phi$ and (b) corresponding frequency response of the NTW-MZM using the hPN with a doping density of $1 \times 10^{18}$ cm$^{-3}$. The dotted line indicates $-3$ dB.

First, we discuss the frequency response of NTW-MZM using the hPN with a doping density of $1 \times 10^{18}$ cm$^{-3}$. The frequency response was simulated using $V_{pp} = 0.16$ V. The simulated phase shift $\Delta \phi$ versus frequency and the corresponding frequency response are shown in Figs. 3(a) and 3(b), respectively. The phase shifter lengths of 1 and 2 mm were simulated to clarify the influence of transit time on bandwidth. The 2-mm phase shifter has two times higher phase modulation at the low frequency side ($f \rightarrow 0$) but shows a much lower 3-dB bandwidth than the 1-mm one. The 3-dB bandwidths are about 44.3 and 23.3 GHz for 1 and 2 mm, respectively. This bandwidth difference comes from the difference in the transit time. With increasing the frequency, when the half period time $(1/2f)$ becomes shorter than the transit time $\tau_0$ ($n_a L/c_0$), the accumulated phase shift in the positive half period of RF voltage will be cancelled in the negative half period in coming. Full cancellation will occur when $f = 1/(2\tau_0)$ at which all of the time passing through the phase shifter exactly equals the full RF period. Thus, $\Delta \phi = 0$ occurs $\approx 40$ GHz for $L = 2$ mm ($c_0/n_a L = 39.5$ GHz). As seen in Eqs. (1) and (2), the calculated bandwidths have already included the contribution from RC constant. Because the RC bandwidth $(>80$ GHz) is very high, as seen in Fig. 3(b), these bandwidths are mainly dominated by the transit time.

Next, we examine the eye diagrams in response to above frequency responses under driving by a large signal ($V_{pp} = 3$ V). $2^7-1$ PRBS pattern was used in simulation and the waveform was generated by introducing pulse edge response $exp[-(t/\tau)^2]$ ($\tau = 18$ ps) [29]. This response corresponds to a $\sim 12$ ps rising/falling time (20$%$–80$%$) which agrees well with most pulse pattern generators. Figs. 4(a) and 4(b) show the simulated eye diagrams for 1 and 2 mm, respectively. Both lengths show clear eye opening at 25 Gb/s and the extinction ratios (ER) are 2.9 and 6.7 for 1 and 2 mm, respectively. The eye can keep open for up to 56 Gb/s for 1 mm even though the ER degrades with increasing the bit rate; but for 2 mm, the eye can be opened for up to 40 Gb/s. The large-signal eye diagram results are consistent with above small-signal bandwidth analysis. Thus, the tradeoff between ER and bandwidth can be simply done by changing the total phase shifter length through changing the segment number in Fig. 1(a). Decreasing the phase shifter length can enhance the bandwidth, but the ER also decreases. Therefore, we have two questions for further improvement: (1) how to enhance the ER with a similar bandwidth; and (2) how to enhance the bandwidth with an equivalent ER. For these goals, we plan to apply the vPN shown in Fig. 2(a) to the NTW-MZM, which will be discussed in the next.

C. Vertical PN for NTW-MZM

Here we discuss what kinds of $pn$ junctions can offer better tradeoff of figure of merits for NTW-MZMs. For traditional TW-MZMs, various diode structures were proposed to improve the modulation efficiency [30], [31]. The modulation efficiency can be enhanced by increasing the overlap integral between the depleted charge profile and the optical mode profile. Provided that the optical mode is constant, the increase in overlap means an increase in the junction area or charge density, either of which will induce the increase in junction capacitance. Thus, the decrease in bandwidth is usually observed when we try to utilize these diodes with large electro-optic (EO) overlaps to improve the modulation efficiency. However, such diodes can fit better for the NTW-MZM than the TW-MZM because the bandwidth of NTW-MZM is determined by the transit time (i.e., phase shifter length) provided that the RC bandwidth is large enough compared to that determined by the transit time. In other words, using these high-efficiency diodes, we can achieve an enhanced ER with the same phase shifter length and maintain a similar bandwidth because the transit time is not influenced by the doping profile and the junction capacitance. On the other side, we can use even shorter phase shifters to enhance the bandwidth and meanwhile maintain an equivalent ER. The vPN diode is one of such high-efficiency diodes [27], [31] and here we apply the vPN diode shown in Fig. 2(a) to the NTW-MZM.

The simulated EO responses and eye diagrams are shown in Figs. 5 and 6, respectively, for various diode conditions. The obtained 3-dB bandwidths (BW) and ER (at 25 Gb/s)
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D. Bandwidth Comparison Between NTW and TW MZM

The RF loss and transit time determine the theoretical bandwidth limits for TW-MZMs and NTW-MZMs, respectively. The bandwidths of real devices are always less than these limits due to impedance mismatching and RF-optical velocity mismatching. Here we compare these two theoretical bandwidth limits, the RF loss dominated bandwidth of TW-MZM and the transit-time dominated one of NTW-MZM. The major source of RF loss at high frequencies is the dielectric loss resulting from the loaded pn diodes [14], [16], which can be calculated by solving the differential equation (Eq. 5 in [14]) of voltage across the diode to the propagation distance using the diode parameters in Section II.B.

The calculated RF losses of TW-MZMs and corresponding EO responses are shown in Figs. 7(a) and 7(b), respectively. In Fig. 7(a), the RF loss of the vPN is about 2.5–3 times higher than that of the hPN for the frequency up to 60 GHz. The high RF loss of the vPN mainly originates from its large junction capacitance as shown in Fig. 2(b) because the RF loss is proportional to the squared capacitance [32]. Thus, the TW-MZM using the vPN shows much lower 3-dB bandwidths than that using the hPN due to the large area of depletion layer; and (2) the loss per unit length will be doubled if we double the doping density for both hPN and vPN. Thus, we can compare the relative loss figure between the situations in Table II.

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Fig. 7. (a) Calculated frequency dependent RF loss of TW electrodes loaded with the hPN and vPN at a doping density of $1 \times 10^{18}$ cm$^{-3}$. (b) Comparison on the EO responses between NTW and TW MZMs.

bandwidth estimated from RF loss for the 2-mm vPN-based TW-MZM is $\sim 18$ GHz that is close to our experimental one ($\sim 17$ GHz) of a TW-MZM using a similar vPN diode [31].

We cite the EO response of the 1-mm hPN-based NTW-MZM from Fig. 5 in Fig. 7(b) for comparison. For the hPN, the 1-mm NTW-MZM is observed to have a bandwidth larger than that of the 2-mm TW-MZM, but smaller than that of the 1-mm TW-MZM. But for the vPN, the bandwidth of 1-mm NTW-MZM ($\sim 37$ GHz as seen in Fig. 5) can be $\sim 12$ GHz higher than that of the 1-mm TW-MZM and $\sim 19$ GHz higher than that of the 2-mm TW-MZM. In addition, the 1-mm vPN NTW-MZM shows a comparable bandwidth and modulation depth to the 2-mm hPN TW-MZM. Therefore, the vPN can offer better performance if being applied to NTW-MZMs than to TW-MZMs. The vPN suffers severe RF loss that can induce great bandwidth degradation in TW-MZMs, but the RF loss can be avoided in NTW-MZMs.

III. EXPERIMENT

A. Device and Measurement

With the segmented rib waveguide structure as shown in Section II.A, two designs of NTW electrodes were fabricated for comparison, as shown in Figs. 8(a) and 8(b). We use Dev A and Dev B to present the NTW-MZM in Fig. 8(a) and 8(b), respectively. Dev A is exactly same as that shown in Figs. 1(a) and 1(b), which adopts the GSGSG configuration and the electrical pads are put on the top of segmented rib waveguides (indicated by slab region in Fig. 1(b)). This design offers the smallest footprint but introduces additional GS capacitance. This GS capacitance mainly comes from the formed vertical capacitor between doping layer accessing the ground and the S metal on top of phase shifter. Dev B adopts GSSG configuration using the comb-like electrode with a comb strip width of 7 μm. The electrical pads are put outside the rib waveguides, while the footprint keeps compact as well ($< 250$ μm in length). Additional GS capacitance in this design is expected to be negligible since the vertical capacitor as formed in Dev A is removed and the capacitor between comb strips has smaller area and larger dielectric thickness than the vertical capacitor in Dev A, which is estimated only about 14 fF. These two devices will be analyzed to clarify the dominant factors in determining high-frequency performance. Both devices were fabricated on the 220-nm silicon-on-insulator wafers with 3-μm-thick box layers using the standard AIST-SCR 12-inch silicon photonics line [33]. The widths of connecting wire waveguides and rib waveguides are 430 nm and 600 nm, respectively. The slab and clad thicknesses are $\sim 110$ nm and $\sim 1.5$ μm, respectively. AlCu is used for the electrode that has a thickness of $\sim 1.5$ μm. For optical coupling, inversely tapered structures were fabricated at both facets of chips. So far, we only fabricated the NTW-MZM based on the conventional horizontal pn junction that is same as our previous work [34]. The average doping density is about $5 \sim 8 \times 10^{17}$ cm$^{-3}$.

The measurement system is shown in Fig. 8(c). The details of measurement can be referred to in [24], [34]. The laser light ($\sim 1.55$ μm) emitted from a TLS was tuned to TE polarization and then coupled into the device through a tapered fiber. The output light was taken by another tapered fiber and then input to an optical power meter or EDFA for high frequency measurement. After EDFA, the light was selectively connected to a high-speed oscilloscope with a 30-GHz optical plug-in module for eye diagram measurement or EDFA for high frequency measurement. The converted electrical signal from

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Fig. 8. Optical microscope photograph of the fabricated NTW-MZMs: (a) GSGSG (Dev A) and (b) GSSG (Dev B) configuration. (c) Measurement system. TLS: tunable laser source. PC: polarization controller. DUT: device under test. PM: optical power meter. EDFA: Er-doped fiber amplifier. PPG: pulse pattern generator. OSC: oscilloscope. PD: high speed photodetector. DC: dc source. Amp: electrical amplifier. VNA: vector network analyzer.
PD was returned to VNA via a high frequency electrical amplifier. High-frequency probes were used for signal input and output after calibrated by a standard open-short-through-load substrate. An external 50Ω terminator was loaded also through a high-frequency probe. We performed single-arm driving on the left arm instead of push-pull driving and all measurements were taken around the quadrature point of MZM that was controlled using the right arm. The EO responses were calibrated using the properties of probe and high-speed PD. As for the dc performance of the phase shifter under the current diode condition, we evaluated \( V_\pi \) around 2.1–2.3 V/cm and the loss around \( \sim 20-23 \) dB/cm for 0–3 V biases. All biases used below denote reverse biases.

**C. EO Bandwidth**

The measured EO responses are shown in Figs. 10(a)–10(c) that correspond to the measured eye diagrams in Figs. 9(a)–9(c), respectively. The 3-dB bandwidths are extracted from these EO response curves and plotted in relation of bias in Fig. 10(d). Dev A has a bandwidth of \( \sim 13 \) GHz at 3–4 V biases, as seen in Fig. 10(a) [24], explaining the its eye degradation at 25 Gb/s. However, with the same total \( pn \) junction length (\( \sim 1.1 \) mm) as Dev A, Dev B in Fig. 8(b) shows improved bandwidth, \( \sim 26 \) GHz at 3–4 V biases, as seen in Fig. 10(b), which contributes to its much better eye performance in Fig. 9(b). This bandwidth improvement is believed partially resulting from the reduction in GS capacitance and will be further explained in next Section III.D. If increasing the total \( pn \) junction length to \( \sim 2.0 \) mm, the bandwidth decreases to \( \sim 14 \) GHz at 3–4 V biases, as seen in Fig. 10(c). As shown in Fig. 10(d), it is well known that increasing the bias can increase the bandwidth due to the junction capacitance reduction (see Fig. 2(b)).

In experiment, there are two factors left in above analysis that influence the EO bandwidth. (1) The first is the connecting waveguides between the segments of rib waveguides. Because the connecting waveguide length also contributes to the transit time, the bandwidth will be decreased. The connecting waveguide lengths are \( \sim 350 \) μm and \( \sim 650 \) μm for the 1.1 mm and 2.0 mm devices in Fig. 9, respectively. The estimated bandwidths after counting this additional length (\( \sim 32 \) GHz for 1.1 mm and \( \sim 18 \) GHz for
junction parameters, we calculated \( V_R \) is considered due to the comb lines are the series resistances of the junction across the junction capacitor \( C_j \). 6 V for both Dev A and Dev B, this LC resonance can enhance the EO bandwidth because it will increase the partial voltage across junction capacitor, which introduces a peaking effect in optical frequency response. The contribution of this inductance to enhance the EO bandwidth will be clearly seen in the following discussion. Furthermore, we simulated Dev B with the comb-like electrode using the SONNET software [35] to examine the appropriateness to treat it as a lumped element. Fig. 12(a) shows the model built in SONNET, where we leave the right arm as metal only but load ideal R and C components with same parameters as above \( pn \) junction (Fig. 11(b)) into the left arm. The simulated one-port S parameters are shown in Fig. 12(b) where we did not observe any ripples in both S11 (with RC loaded) and S22 (metal only) that might occur due to the comb-like strips. Thus, Dev B is small enough to be treated as a lumped element as seen from the RF signal side.

The characteristic impedance extracted from the measured S parameters for the bias of 6 V and those from circuit model calculation and SONNET simulation are shown in Fig. 12(c). The impedance calculated by circuit model shows good consistence with the experimental data for both Dev A and B. Furthermore, for Dev B, the circuit model in Fig. 11(b) is well reproduced by SONNET simulation. At the low frequency side (approaching to zero frequency), the device is close to open circuit due to the existence of capacitors, while with increasing the frequency, the impedance quickly decreases. It is this impedance decrease that allows the use of capacitor \( C_j \) as a lumped element. Fig. 12(d) describes the voltage transfer efficiency and its frequency dependence can directly reflect the EO response. In our measurement, both ports of Figs. 11(a) and 11(b) are terminated with 50 \( \Omega \), thus \( V_s/V_c \) equals 0.5 for both devices at the low frequency side because
the devices are nearly open. $V_c/V_s$ decreases with increasing the frequency due to the impedance decrease for both devices, but Dev B has larger $V_c/V_s$ than Dev A in a wide frequency region so that the EO bandwidth of Dev B is larger than that of Dev A, as shown in Fig. 10. For NTW-MZMs, the phase change is proportional to the integration of voltage as expressed by Eq. (2), so the bandwidth at which $V_c/V_s$ decreases to the 0.707 times zero-frequency value indicates the 3-dB EO bandwidth. Such bandwidths in Fig. 12(d) are $\sim$14 GHz for Dev A and $\sim$28 GHz for Dev B, both of which correctly reflect the experimental EO bandwidths in Figs. 10(a) and 10(b), respectively, while these bandwidths do not match well with those estimated from transit time with connecting waveguide taken into considerations.

To explain the bandwidth enhancement from Dev A to Dev B, we also calculate $V_c/V_s$ for Dev B without considering the inductance $L_c$, which is shown in Fig. 12(d) for comparison. It can be seen that the increase of impedance in Dev B due to the reduction in GS capacitance contributes to $\sim$6 GHz bandwidth enhancement and the existence of $L_c$ can further increase $V_c$ within a wide frequency region due to LC resonance, which contributes to $\sim$8 GHz bandwidth enhancement. The overall frequency response sums up a gradually decreased component and a peaking component that can be more obvious if changing the impedance matching conditions as seen in next section. Under the same termination, when increasing the pn junction length to 2.0 mm, the impedance will decrease and the LC resonance frequency also decreases, thus induces a decrease in EO bandwidth even though the RC product remains unchanged. Therefore, we conclude that (1) Dev B is better than Dev A due to its larger impedance and the existence of LC resonance that can enhance EO bandwidth; (2) for a fixed electrode design of NTW-MZMs, the impedance mismatching dominates the experimental bandwidth in our measurement.

E. Driving Configuration for Impedance Matching

To improve the bandwidth for NTW-MZMs, it is necessary to improve impedance matching condition that can be realized by modifying driving configuration even without changing the device itself. As discussed above, our experimental results and calculation are done with a 50-$\Omega$ terminated 2-ports configuration. Obviously, this is not optimum matching condition for high frequencies because the impedances at high frequencies are much lower than 50 $\Omega$. We can only perform 50-$\Omega$ matching in our cable-based measurement setup in Fig. 8(c), however, it is possible to design low-impedance matching (e.g., 25 $\Omega$) in integrated modulator drivers. Furthermore, in conventional TW-MZMs, single push-pull (SPP) driving (two $pn$ diodes are connected in series) is widely used to improve the bandwidth [18],[36]. SPP driving scheme can also be applied to the NTW-MZM for better impedance matching because the effective capacitance is reduced. Thus, we examine this driving scheme under both 50 $\Omega$ and 25 $\Omega$ matching below. We define our current driving configuration shown in Fig. 11(b) with a terminated port 2 as the 2-port configuration, while that leaving the port 2 open as the 1-port configuration. The port 1 is connected to the driver, so it has a terminated resistor at default.

1) 50-$\Omega$ matching. Even though Dev B shows better EO bandwidth than Dev A, SPP can be done for both devices to achieve larger bandwidths. Here, taking Dev B as an example, we compare $V_c/V_s$ under SPP driving with that of Dev B under current single arm driving because the frequency dependent $V_c/V_s$ directly reflects EO bandwidth as mentioned above. Figs. 13(a) and 13(b) show the calculated $V_c/V_s$ at 50-$\Omega$ matching for 2-port and 1-port configurations, respectively. For single arm driving (blue lines), the 1-port configuration shows lower bandwidth ($\sim$13 GHz) than the 2-port one ($\sim$28 GHz) because its $V_c$ at zero frequency is doubled compared to the 2-port one due to the lack of voltage dividing resistor. If adopting SPP driving (red lines), the bandwidth can be increased to $\sim$43 GHz and $\sim$29 GHz for the 2-port and 1-port configurations, respectively. This bandwidth increase mainly comes from the reduction of $V_c$ at the low frequency side because the RF voltage is divided by two diodes. At zero frequency (complete open circuit), $V_c$ of SPP driving is reduced to a half that of single arm driving, but at higher frequencies this reduction ratio is relatively increased because the overall RF voltage across two diodes is increased due to the impedance increase from series connected capacitance and resistance.

2) 25-$\Omega$ matching. If the device is matched to 25 $\Omega$, the bandwidth will be improved greatly compared to 50-$\Omega$ matching. Figs. 13(c) and 13(d) show the calculated $V_c/V_s$ at 25-$\Omega$ matching for 2-port and 1-port configurations, respectively. The bandwidth can be increased from current $\sim$28 GHz at 50-$\Omega$ matching to $\sim$43 GHz at 25-$\Omega$ matching at the 2-port configuration and from current $\sim$13 GHz at 50-$\Omega$ matching to $\sim$29 GHz at 25-$\Omega$ matching for the 1-port configuration. This bandwidth improvement does not sacrifice the modulation efficiency because the ratio $V_c/V_s$ is enhanced over a wide frequency range compared to 50-$\Omega$ matching, which can be seen by comparing blues lines in Figs. 13(c) and 13(d) to those in Figs. 13(a) and 13(b), respectively. Thus, low-impedance matching is a simple
way to improve both bandwidth and efficiency for NTW-MZMs. Similarly as in 50-Ω matching, adopting SPP driving can further improve the bandwidth to \( \sim 48 \) GHz and \( \sim 43 \) GHz for 2-port and 1-port configurations, respectively. Furthermore, the 1-port configuration offers larger \( V_T/V_s \) than the 2-port one for both 25-Ω and 50-Ω matching, so the 1-port configuration can offer better modulation efficiency. The contributions from LC resonance have been already included in above results in this section and its peaking effect varies with impedance matching conditions, as seen in Fig. 13. The LC resonance induced bandwidth enhancement is more obvious for low impedance matching, SPP driving, and 2-port configuration.

Therefore, according to all above discussions, there are three remaining works in future study for improving the performance of NTW-MZMs: (1) SPP driving with or without low impedance matching; (2) Designing NTW-MZMs with shorter or without connecting waveguides; and (3) Applying vertical pn junctions. Through these studies, we could achieve better tradeoff between high-frequency performance and compact footprint for NTW-MZMs.

### IV. Conclusion

We experimentally demonstrated carrier-depletion NTW-MZMs on silicon photonic platform. They adopt densely meandered phase shifters and thus offer an ultra-compact lumped-element MZ modulator for using carrier depletion. Up to 28.1 Gb/s NRZ eye diagrams and \( \sim 26 \) GHz EO bandwidth were experimentally achieved, showing improved high-frequency performance than previous works. The proposed lumped circuit model well explained the experimental S-parameters and frequency dependent impedance, indicating that the impedance mismatching is the dominant factor in determining experimental bandwidth and the LC resonance can contribute to bandwidth enhancement for NTW-MZMs using a comb-like electrode with suitable inductance. Driving schemes that could enable better impedance matching was proposed to enhance bandwidth. This NTW-MZM could enhance the integration density of modulators for wavelength-division multiplexing transceivers and offer better tradeoff between compact footprint, modulation depth, and high speed for utilizing carrier depletion.

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