I. INTRODUCTION

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Micro-ranging and 3D micro-imaging have direct applications in industrial measurement systems. There are existing tools for this purpose, but their large form-factors limit their usage in manufacturing industries. A chip-scale solution can make this technology accessible to other areas, such as enhancing the 3D-imaging capability of modern 3D printers to create high-fidelity 3D copy machines. Medical applications that would benefit from this technology include corneal imaging for contact lens fitting and 3D vision for robotic microsurgery.

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There are two main classes of chip-scale 3D imaging technologies. Techniques in the first category use image processing; these include stereovision [1], [2], structured-light cameras [3]–[5], and multi-focus imaging [6]–[8]. These methods are based on correlating the features on photographs of a scene, such as the view angle, curvature of the illumination pattern, or blur and gradient of focus, with the scene's 3D geometry. These image processing techniques typically rely on mainstream hardware, including photographic cameras and processor chips, and are therefore suitable for inexpensive depth estimations. However, their dependence on features of optical pictures, such as object edges, makes them inaccurate and hence inadequate for measuring distances to certain classes of objects, such as those with white shiny surfaces or smooth curvatures.

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The second category of 3D imagers is based on measuring the round-trip delay of an ultrasonic or electromagnetic wave to the target. Ultrasonic rangefinders can operate with very low power consumption, making them suitable for detection and ranging in mobile devices [9], but their lateral and depth resolution are both limited by their approximately mm-scale ultrasonic wavelengths.

Ultrasonic waves with shorter wavelengths exhibit exponentially larger attenuation in air, rendering them impractical for high-precision 3D imaging. Radars and lidars use electromagnetic waves in radio and optical spectra, respectively. Light waves have much shorter wavelengths than RF waves; thus, lidars can offer better lateral resolution and depth precision than radars, making them more suitable for the aforementioned

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3D micro-imaging applications.

The most straightforward lidar architecture is based on measuring the round-trip delay of a light pulse to the target [10], [11]. As illustrated in Fig. 1, this type of lidar consists of a pulsed light source (e.g., laser), a detector, and the electronic timing circuitry. The simplicity of the optical parts in a pulsed lidar makes it attractive for many applications. However, to achieve precision in the range below 10 μm, the electronic circuits must be able to measure the round-trip delay of the light pulse with a precision of approximately 70 fs. This is a difficult task for present-day electronic circuits, and toward this objective, frequency-modulated continuous-wave (FMCW) lidar [12]–[14] can relax the required precision in the round-trip delay measurement in exchange for a more sophisticated optical architecture.

One of the difficulties in implementing FMCW lidars is the precise modulation of the laser frequency. In previous works, this task has been accomplished using bench-top optical and electronic devices [15], [16]. In this work, we present an electronic-photonic integrated circuit for laser frequency modulation that demonstrates better precision than previously reported results from the bench-top systems and enables chip-scale 3D micro-imaging.

This paper is organized into seven sections. Section II presents the operating principle of the FMCW lidar. Sections III and IV describe the operation principle and the design procedure, respectively, of the electro-optical PLL (EO-PLL) proposed for laser frequency modulation. Practical aspects of the implementation, including simulation and fabrication, are discussed in Section V. Section VI shows the experimental setup and the measurement results.

II. Frequency-Modulated Continuous-Wave (FMCW) Lidar

Fig. 2(a) shows the basic architecture of the FMCW lidar, which basically consists of a tunable laser and a coherent receiver. The optical frequency of the tunable laser is linearly modulated with time. The frequency-chirped light hits the target, and the light's reflection is collected in the receiver and combined with a local branch of the laser light. As shown in Fig. 2(b), the time delay between the local light and the reflection

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causes a frequency difference between the two light signals. Consequently, a beat tone at the frequency of this difference appears on the interference of the two light beams and is detected by a photodiode. The target distance can be determined by measuring the beat frequency of the photocurrent according to the following equation:

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$$f_R = \frac{2\gamma}{c} \cdot R \qquad (1)$$

where f_R is the measured beat frequency, γ is the slope of the laser frequency modulation, c is the speed of light in free space, and R is the range to the target.

A. Required Timing Accuracy for FMCW versus Pulsed Lidar

It was previously mentioned that the electronic timing precision required for micro-ranging can be more relaxed for an FMCW lidar compared to a pulsed one. As shown in Fig. 1(b), in a pulsed lidar, the ranging precision for a given electronic clock period can be found from the following equation:

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$$\delta R = \frac{1}{2} c \cdot \delta\tau_R = \frac{1}{2} c \cdot T_{clk} \qquad (2)$$

where δR is the ranging error, c is the speed of the light, $\delta\tau_R$ is the timing precision, and T_{clk} is the period of the electronic clock. According to this equation, to achieve ranging precision below 10 µm using a pulsed lidar, the electronic circuitry should resolve the pulse time-of-flight with a precision of approximately 70 fs. For FMCW lidar, however, the situation is different. To compare the FMCW and pulsed lidars directly, consider the waveforms shown in Fig. 3. The goal is to perform a single ranging measurement within the sample time, T_s. Assume that the beat tone has cycles within this time. To measure the beat frequency, or equivalently the beat period ($T_d = 1/f_d$), an electronic clock can be used as a reference. The timing error caused by the clock period will equally distribute among the m successive cycles of the beat signal. The ranging error can then be approximated as given below:

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$$\frac{\delta R}{R} = \frac{\delta T_R}{T_R} = \frac{T_{clk}/m}{T_s/m} = \frac{T_{clk}}{T_s} \qquad (3)$$

where $\delta R/R$ is the normalized ranging error, $\delta T_R/T_R$ is the normalized beat period measurement error,

T_{clk} is the period of the electronic clock, T_s is the measurement time for one sample, and m is the number of beat cycles per T_s. To illustrate the difference, let us consider a numerical example. Assume that a ranging measurement must be done at a rate of 1 MP/s for a target at a distance of 10 cm. Based on (3), with an FMCW lidar, the required timing precision for the electronic circuits would be 100 ps, which is more than 1400 times larger than the approximately 70 fs precision required for the pulsed lidar. If multiple pulses were to be averaged to relax the precision requirement for pulsed lidar, more than two million pulses would have to be fired during the 1 μs measurement period, which is not practical. From this discussion, the primary advantage of the FMCW lidar for high-precision range-finding and 3D imaging is clear.

B. Laser Phase Noise and Photodiode Shot Noise

The phase noise of the tunable laser and the shot noise from the photodiode are the two primary limiting factors on the ranging precision and maximum operating distance of an FMCW lidar. Therefore, this section presents a comprehensive analysis of the effect of these two noise sources on the performance of the FMCW lidar.

1) Laser Phase Noise: Linear modulation of the laser frequency should ideally result in a single-tone sinusoidal current in the photodiode of the coherent receiver. However, the phase noise of the laser also causes phase noise in the spectrum of the photocurrent. An analysis of this phenomenon is given in [17]. In that work, it is assumed that the laser's frequency noise has a white spectrum. This is a valid assumption for fixed-frequency lasers, in which the majority of frequency noise is the result of spontaneous emissions in their active regions. However, for a tunable laser, in which the noise is dominated by the noise from the band-limited tuning process, it is more accurate to assume a colored frequency noise spectrum for the laser. The work presented in this paper uses an edge-emitting carrier-injection tunable distributed Bragg reflector (DBR) laser with a central wavelength of approximately 1530 nm [18]. Fig. 4(a) shows a schematic

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of the device, which has two electrical ports. One of the ports provides bias current for the laser diode to generate the light. The other port is used to tune the laser frequency by injecting a charge carrier in the laser's DBR section and modifying its refractive index. As shown in Fig. 4(a), the tuning port can be modeled with a p-n junction at which a change in current from 10 to 15 mA can change the laser frequency by approximately 120 GHz. The transfer function from the tuning current to the laser frequency is a first-order low-pass filter with a 60-MHz bandwidth. Within this bandwidth, the spectrum of the laser frequency noise is dominated by the shot noise from the tuning current. At higher frequencies, the spontaneous emission in the active region becomes more dominant; however, for the purposes of this analysis, this effect can be neglected. The average current in the tuning section is 12.5 mA, which results in a shot noise density of 4×10^{-21} A^2/Hz. Through the 24 GHz/mA tuning gain, this value of shot noise translates into an optical frequency noise level of $f_n^2 = 2.3\times10^6$ Hz^2/Hz at DC.

2) For round-trip delay to the target equal to τ_R, and defining $\xi_R = 1/(\pi^2 f_n^2 \omega_p^2 \tau_R^2)$, the spectral density of the photocurrent in an FMCW lidar employing this laser is derived as detailed in the Appendix, resulting in:

$$S_i(\omega) = \frac{i_0^2}{2}\left\{e^{\frac{-2}{\xi_R\cdot\omega_p}}\cdot sinc^2\left(\frac{T_{ramp}\cdot\omega}{2}\right) + \frac{\xi_R}{1+(\xi_R\omega_p/2)^2}\cdot\left[1-e^{\frac{-2}{\xi_R\cdot\omega_p}}\cdot\left(\cos\left(\frac{\omega}{\omega_p}\right)+\frac{1}{\xi_R\cdot\omega}\cdot\sin\left(\frac{\omega}{\omega_p}\right)\right)\right]\right\} , \quad (4)$$

where the first term is the signal spectrum measured during a single modulation ramp, the second term is the noise spectrum, and i_0 is the amplitude of the sinusoidal photocurrent in the absence of the laser phase noise. By adding the photodiode shot noise to the spectral density of the photocurrent given here, the value of the signal-to-noise-ratio (SNR) in the coherent receiver can be found.

3) *Scattering Loss and Photodiode Shot Noise:* The detector shot noise can be modeled by adding noise to the spectrum of the photocurrent as it was calculated in the previous step. The level of shot noise is determined from (5), based on the average current in the detector, including the dark current and the photocurrent generated by the combination of light power reflected back from the target and the local

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branch of light that mixes with the reflection:

$$i_{n-shot}^2 = 2q_e I_D = 2q_e[I_{dark} + R_{PD}(P_L + P_r)] \approx 2q_e R_{PD} P_L. \quad (5)$$

where I_{dark} is the dark current of the photodiode, and R_{PD} is its responsivity; P_L is the optical power from the local branch in the coherent receiver, and P_r is the optical power collected from the target reflection. In most cases, the photocurrent from the light in the local optical branch is dominant and the given approximation can be used. To compare the contribution of the laser phase noise and the detector shot noise, the amplitude of the photocurrent i_0 must be found. For a receiving aperture area of A_r, a target at a distance of R, a reflectivity of α, and Lambertian scattering, the value of Pr in terms of transmitted power P_t can be found as shown below:

$$P_r = \frac{\alpha \cdot A_r}{\pi R^2} \cdot P_t = \frac{\alpha \cdot A_r}{\pi \cdot c^2 \cdot \tau_R^2} \cdot P_t \quad . \quad (6)$$

where c is the speed of light in air, and the photocurrent amplitude, i_0, is given below:

$$i_0 = R_{PD}\sqrt{P_L \cdot P_r} \quad . \quad (7)$$

Using this result and considering the near-carrier flat component of the noise spectrum in (4), and for $\tau_R^2 \ll \frac{1}{4\pi\omega_p f_n^2}$, the spectral noise level caused by laser phase and frequency noise can be approximated as follows:

$$i_{n-Lpn}^2 = \frac{i_0^2}{4\omega_p}\left[1 - e^{-4\pi \cdot \omega_p \cdot f_n^2 \cdot \tau_R^2}\right] sinc\left(\frac{\omega}{\omega_p}\right) \approx \frac{4}{c^2}\alpha \cdot A_r \cdot R_{PD}^2 \cdot P_L \cdot P_t \cdot f_n^2 \quad . \quad (8)$$

Equation (5) shows that under the following condition, the contribution of the photodiode shot noise will be negligible:

$$\frac{2}{c^2}\alpha \cdot A_r \cdot R_{PD} \cdot P_t \cdot f_n^2 > q_e \quad . \quad (9)$$

Both sides of this inequality are independent of the target distance. For $\alpha = 10\%$, $A_r \approx 1\ \text{in}^2$, and $R_{PD} = 0.8$, and with the laser used in this work, the right-hand side of this inequality will evaluate to approximately 1.8×10^{-18} C which is more than 10 times larger than the electron charge q_e. Hence, in this

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work, the laser phase noise will always be the limiting factor for the performance of the lidar.

Fig. 6 shows the value of the SNR versus the range R calculated numerically using (4) and (5). Using these equations and considering the effect of post-processing steps (e.g., sampling, filtering, and frequency-measurements based on zero-crossings of the beat signal), the ranging error versus the range is also calculated and plotted in Fig. 7. It is important to notice that unlike predictions generated by the assumption of a white frequency-noise spectrum [17], the RMS value of the range error versus the range increases at a rate much higher than $\sqrt{R}$. This result can be attributed in part to the colored spectrum of the laser frequency noise that must be considered; the result is further attributed to an artifact of the particular post-processing steps employed in this work that are optimized to increase the precision of the range measurement for close distance objects.

III. Electro-Optical PLL (EO-PLL) Principle

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In the previous section, the tunable laser was identified as one of the essential elements in an FMCW lidar. One of the important issues when employing a tunable laser in an FMCW lidar is the linearity of its frequency-tuning characteristic. Fig. 8 shows the modulation nonlinearity of the DBR laser used in this work. As can be seen from the figure, the modulation slope of the laser can vary up to 9 %. In addition to the systematic variation that can be reduced by calibration, there is also random drift in this characteristic. Measurement results indicate that this drift causes a 65-μm ranging error for a target placed at a distance of 5 cm. One way to suppress both the nonlinearity and the drift in the tuning characteristic is to use an electro-optical phase-locked loop (EO-PLL) [15]. In this section, the operation principle of an EO-PLL is described.

A. *Linear Frequency Modulation of the Laser Using Feedback*

The basic architecture of an EO-PLL is shown in Fig. 9. The operation of this feedback mechanism can be more easily understood by first considering the sub-blocks enclosed in the dashed box, which include an integrator, the tunable laser (TL), and an optical element called a Mach-Zehnder interferometer (MZI). Assuming that a constant voltage is applied to the input of the integrator, its output would be a voltage ramp. Thus, the laser frequency that is controlled by this voltage will also increase linearly. With linear frequency modulation, the light will then propagate into the MZI. This element splits its input light into two waveguides with unequal length and recombines them into a single output waveguide. As with the FMCW ranging measurement, the delay between the two light beams with linear frequency modulation creates a beat tone that can be detected by the photodiode. Because the differential delay of the MZI is fixed, any variation in the MZI beat frequency is proportional to the deviation of the laser frequency modulation's slope γ. The EO-PLL, similar to an electronic PLL, locks the beat frequency to a clean electronic local oscillator (LO) using a phase/frequency detector (PFD) followed by a loop filter. The modulation slope γ will then be fixed to the following value:

$$\gamma = \frac{f_{LO}}{\tau_{MZI}} \qquad (12)$$

where f_{LO} is the frequency of the electronic LO, and τ_{MZI} is the differential delay of the MZI.

B. Switching Between Up and Down Ramps

The linear frequency modulation of the laser cannot continue indefinitely. The frequency of the DBR laser used in this work can be tuned up to approximately 120 GHz, and after that, it should either be reset or the modulation slope should be reversed to a down-ramp. In this work, the sign of the input signal to the integrator is periodically inverted to switch the modulation between up- and down-ramps, as illustrated in Fig. 9. The resulting waveform of the laser frequency is shown in Fig. 9(b).

IV. EO-PLL Design

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In this section a more rigorous analysis of the performance of the EO-PLL is presented, including its linearity, gain-bandwidth product, noise, etc.

A. Loop Gain and Stability

As with electronic PLLs, loop-gain is the main characteristic of the EO-PLL that defines its dynamics. As previously mentioned, the slope of the laser frequency modulation should be reversed periodically. This switching process can cause a large disturbance in the instantaneous frequency of the equivalent VCO, and the EO-PLL should have sufficiently large locking range to suppress these errors and acquire the correct modulation slope after switching. For this reason, a charge-pump type-II PLL architecture is employed in this work. A phase-margin of 70° must be maintained for optimum settling behavior. Waveguide connection between the optical elements causes an approximately 20 ns time delay in the loop. To maintain the desired phase margin with this constant delay in the loop, its bandwidth should be lower than approximately 700 kHz. The overall open-loop gain for the EO-PLL is given in Fig. 10.

It is worth mentioning that when designing EO-PLLs with the described architecture, the tuning characteristic of the laser may also be of significance. The tuning characteristic of the DBR laser used in this work has a bandwidth of 60 MHz, which is far larger than the approximately 700 kHz EO-PLL bandwidth, and its effect on the phase margin of the EO-PLL is negligible. For other types of tunable lasers, such as thermally or mechanically tuned ones, the tuning bandwidth can be much smaller than this value, in which case its effect on the EO-PLL dynamics must be considered.

B. Maximum Modulation Speed

A higher modulation speed is desirable for increasing the gain of the system, which, from (1), is proportional to the modulation slope γ, and also for enabling larger throughput. However, by increasing the modulation speed, the error due to nonlinearity will also repeat at a higher rate, and the EO-PLL will be less effective in suppressing the error. Without taking special measures, the nonlinearity and its drift can result in a 65-μm ranging error, which is not acceptable for many applications. For instance, present-day 3D printers can have a depth resolution as good as approximately 20 μm, and to harvest this capability for rapid prototyping, a 3D imager with even better resolution (e.g., sub-10 μm) is required. In addition, in medical applications, such as 3D corneal imaging or 3D vision for robotic microsurgery, a sub-10-μm resolution can enable 3D imaging at the cellular level and provide the maximum possible reliability in such applications. Thus, to have any meaningful impact, the feedback loop should be able to suppress the ranging error from 65 μm to less than 10 μm. As shown in Fig. 10, the frequency of the triangular modulation should be equal to approximately 125 kHz for its nonlinearity to be suppressed by 7 times the 17 dB loop gain, which is equivalent to an approximately 4 μs single-ramp duration. To maintain some extra margin, in the final design, a ramp duration of approximately 5.5 μs is used. Thus,

$$\gamma = \frac{\Delta f_{max}}{T_{ramp}} = \frac{120\ \text{GHz}}{5.5\ \mu\text{s}} = 22\ \text{GHz}/\mu\text{s}\ . \qquad (13)$$

From (12), the value of γ can be set by choosing proper values for τ_{MZI} and f_{LO}. A large value for f_{LO} relaxes the phase-margin requirement of the EO-PLL; however, a fixed γ also requires a large τ_{MZI}. Furthermore, a longer MZI waveguide exhibits more loss and occupies a larger area. In this design, an MZI is selected with a delay of 330 ps, corresponding to an approximately 6.5-cm physical waveguide length on the photonic chip. With this delay, the resulting LO frequency is 7.2 MHz.

C. Gated Ramp-Switching

Switching the direction of the modulation ramp can cause a phase jump in the phase of the MZI beat signal. As illustrated in Fig. 11, the maximum and minimum phase jumps occur when the ramp-switching instant aligns with a zero-crossing or peak of the MZI signal, respectively. The large phase jump caused by ramp-switching at a zero-crossing forces the EO-PLL out of lock and, as shown in Fig. 11(b), immediately after the switching instant, the modulation waveform becomes nonlinear. The normalized error in the modulation slope due to this effect is shown in Fig. 11(c). On these figures, the green curves correspond to the case in which switching occurs close to one of the peaks (maxima or minima) of the MZI beat signal. When switching occurs close to the peak ($\pm 1°$), the switching error becomes negligible; hence, a mechanism to align the switching instant with a peak of the MZI beat signal can suppress this error.

Because the LO and MZI beat signals are phase locked, the transition (rising or falling edge) of the LO signal can be used as a reference to gate the switching signal and delay it until the beat signal peak to minimize the switching error. Such a mechanism is illustrated in Fig. 12. It should be noted that in the ideal case, if there were no extra delay in the loop, the edge of the LO signal would require an exactly 90° phase shift for switching to occur at the peak; however, because other elements of the EO-PLL add some parasitic delay to the signal path, the phase shift in the LO edge should be adjusted so that the overall delay in the loop and the phase shift block sum to 90°. This adjustment is made in a servo loop by observing the error in the period of the beat signal after each switching instant using a time-to-digital converter (TDC) and setting $\Delta\Phi$ to minimize this error.

D. Design of the EO-PLL Circuit Blocks

A detailed block diagram of the EO-PLL with gated ramp-switching is shown in Fig. 13. In this section, the important design considerations for the front-end electronics and the ramp-switching block will be discussed.

1) Front-End Electronics: The photocurrent from the PD_{MZI} is converted to a square voltage to be applied to the charge-pump phase/frequency detector. This task is accomplished using a trans-impedance amplifier (TIA) followed by a voltage limiter. The TIA consists of a low-impedance input stage implemented by the gm-boosted transistor MN_1, and the trans-resistance R_1 amplifies and high-pass filters (M_3 to M_6) the current to remove the low-frequency baseline variation caused by the laser's intensity fluctuation as a side effect of its frequency modulation. The high-frequency noise content is filtered through capacitor C_1 in parallel with R_1. An inverting stage (MP/N_5) matched with the trans-impedance stage (MN/P_4) removes the amplitude envelope before the PFD.

In Section II it was shown that, for a target within the laser coherence length, the laser phase noise can be modeled as noise added to the spectrum of the photocurrent. The auxiliary MZI employed in the EO-PLL is essentially equivalent to a short-range target with a round-trip-delay of $\tau_{MZI} = 330\ \text{ps}$. The amplitude of the photocurrent at the output of the PD_{MZI} is equal to 10 μA; inserting these numbers into (8) results in a current noise density of 80 pA^2/Hz. The input-referred noise from the front-end electronics must be kept below this level. The input-referred current noise from MN_1 is suppressed by the gain-boosting amplifier, OPA. The noise from the transistors MN/P_2 are low-pass filtered through the capacitor C0; hence, the input-referred current noise is dominated by the transistors $MP_{1,3}$ and MN_3 and the resistor R1:

$$g_{m_{n1}} + g_{m_{p1}} + g_{m_{p3}} + \frac{1}{R_1} < 5\ \text{mS} \quad (16)$$

By choosing the $1/f$ noise corner frequency below $1/T_{ramp} = 180\ \text{kHz}$, its effect on the accumulated phase of the MZI and ranging signals within each modulation period will be negligible as well.

2) Ramp-Switching: It was previously mentioned that to reduce the effect of switching error, the ramp-switching signal must be gated with a phase-delayed version of the LO signal. An implementation of this function is illustrated in Fig. 13. The pre-gated ramp-switching signal is generated using a hysteresis comparator. This comparator senses the tuning voltage at its input and is designed to toggle and change the integration direction when its input crosses the allowed tuning boundaries at 1.2 V and 1.4 V. The comparator output is gated and applied to the integrator to change the ramp direction at the next peak of the MZI beat signal.

The gating signal is generated from the LO signal using a PLL followed by a phase-multiplexer. A block diagram of the PLL and phase-multiplexer is depicted in Fig. 15. The PLL employs a four-stage differential VCO with eight output phases. The frequency of the VCO is set to 32 $\times f_{LO}$; hence, within each LO cycle there are 256 equally spaced edges that can be selected by the phase-multiplexer to gate the switching signal.

The phase-multiplexer sets the switching gate by selecting one of the VCO output phases and feeding it to a 5-bit free-running counter that generates an overflow after counting down from its loaded input. A TDC uses the output of the PLL as an 8-phase reference clock and measures the error in the period of the beat signal immediately after each switching instant. The phase-select bits of the phase-multiplexer are then set to minimize the switching error.

V. EO-PLL Verification and Implementation

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One of the challenges to implementing integrated electronic-photonic circuits is their verification. There are well-established and reliable simulation tools for electronic circuits, and photonic circuits can be modeled and simulated using similar tools (e.g., modeling in Verilog-A and simulation with Spectre). The time-scale for accurate simulation of the electronic circuits can be milliseconds; however the central frequency of the optical fields for the 1530 nm wavelength is approximately 200 THz. The time-domain simulation of a large-scale network with such a high-frequency signal for a time duration of milliseconds can take days or even weeks, depending on the complexity of the network. For this reason, the photonic signals and elements are modeled and simulated in amplitude and phase/frequency domains. Imperfections such as laser tuning characteristics, phase noise, the frequency dependence of the laser's intensity, and MZI loss are also included in the models.

All of the electronic and photonic circuits used in the architecture of the EO-PLL shown in Fig. 13 are integrated on a chip-scale platform, except for the tunable laser. The electronic circuits are fabricated in a 0.18 μm CMOS, and the MZI and photodiode are implemented on a silicon-photonic chip. Photographs of the chips are shown in Fig. 16. Together, the two chips occupy 3×3 mm^2 and are pitch-matched for stack integration with through-silicon-vias (TSVs). A photograph of the integrated stack and a tilted SEM of the integrated stack diced at the position of the TSVs are shown in Fig. 16(b) and 16(c), respectively.

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VI. Experimental Results

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This section presents the measurement results for the EO-PLL's performance in linearizing the laser frequency modulation, as well as its application for ranging and 3D micro-imaging.

A. EO-PLL Performance

The performance of the EO-PLL is quantified by measuring the error on the modulation slope γ. The output frequency of the MZI beat signal is proportional to γ and is measured to determine its value. For this purpose, the square-wave voltage at the TIA output on the CMOS chip is recorded, and its cycle-to-cycle frequency is extracted through post-processing. This experiment is performed with three different laser frequency modulation methods: applying a triangular waveform from a signal generator (open-loop), using the proposed EO-PLL (closed-loop) without activating the gated ramp-switching feature, and using the proposed EO-PLL with gated ramp-switching. The plot of the MZI beat frequency versus a single modulation ramp for all of these experiments is shown in Fig. 17. It can be seen from the figure that when using the EO-PLL with gated ramp-switching, the measured error in the MZI beat frequency is limited by the contribution of the laser's phase noise, and the error due to modulation nonlinearity and drift is negligible.

B. Ranging and 3D Micro-Imaging Applications

Improved precision of the laser frequency modulation slope enables ranging with precision close to the theoretical curve derived in Section II. Fig. 20 shows the ranging setup. The laser light power is split into two paths: one path is directed to the on-chip MZI to monitor and regulate the frequency modulation slope with the EO-PLL circuitry, and the other path is used for ranging. This latter part is itself split into two branches. The light from one branch passes through the circulator (port-1 to -2) and is emitted to the target. The reflected light from the target is collected back at the circulator (port-2 to -3) and is combined with the local source light to generate the interference signal at the photodiode output. The frequency of this signal is proportional to the difference between the target distance and the length of the local branch, $2R - c \cdot \tau_0$. This frequency is extracted and normalized to the MZI beat frequency in post-processing steps to determine the range:

$$\frac{f_{MZI}}{l_{MZI}} = \frac{f_R}{2R - c \cdot \tau_0} \Longrightarrow R = \frac{f_R}{f_{MZI}} \cdot l_{MZI} + \frac{c \cdot \tau_0}{2} \qquad (17)$$

With this ratio-metric measurement, the optical length difference of the on-chip MZI branches can be used as the length unit for the target range measurement. Furthermore, because the slow drift of f_{LO} has a proportional effect on f_R and f_{MZI}, the ratio-metric measurement can relax the requirements for the close-to-carrier phase noise of the local oscillator.

It must be noted that the effect of the laser's phase noise on the photocurrent spectrum and ranging error is a function of the path-length difference, $2R - c \cdot \tau_0$ and will be at a minimum when the path-length difference value is equal to zero. In the setup shown in Fig. 20, τ_0 is set to 3.3 ns; hence, the ranging error due to the laser's phase noise would be minimum at a range baseline of approximately 50 cm.

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The ranging error is measured both for open-loop laser modulation and with EO-PLL. A target with a return loss of approximately 5 dB is incrementally moved away from the range baseline, and at each point, 14 ranging measurements are taken, each with the duration of a single modulation ramp (5.5 μs). The standard deviation of these measurements versus distance from the range baseline is plotted in Fig. 19. The precision of the measurement with open-loop modulation is limited by the error on the modulation slope γ. When using the EO-PLL, the error in the modulation slope is suppressed to a negligible level and the standard deviations of the measurements are reasonably close to the expected theoretical curve.

The same setup shown in Fig. 20 is used to create a 3D image of an object. For this purpose, the XYZ stage is stepped in the XY plane, and the distance to each point of the object's surface is recorded in a point cloud matrix. Fig. 20(b) shows a photograph and 3D image of a miniature gear acquired with this technique. The gear is placed 40 cm from the imaging lens. With 180 kP/s throughput, the precision is 11 μm, which appears as a roughness on the object's surface; the actual roughness on the object's surface is less than 1 μm.

VII. Conclusion

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The integrated EO-PLL presented in this work enables 3D imaging with micrometer-level precision in a chip-scale platform. In addition to numerous applications in manufacturing industries, this work enables further applications of 3D micro-imaging in a wider range of fields, such as miniature 3D imagers for robotic microsurgery devices and corneal imaging for contact lens fitting in medical fields and high-fidelity 3D copy machines for rapid prototyping.

The miniaturization of the technology also has favorable effects on the performance of the system. By reducing the physical length of the electrical wires and optical fibers in the feedback path, the dynamics of the EO-PLL can be improved to track and suppress the higher frequency errors. Furthermore, close integration of the electronic circuits with photonic devices enables sophisticated control mechanisms such as the proposed gated-switching technique.

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Appendix: Photocurrent Spectrum in FMCW Lidar with Colored Phase–Frequency Noise Spectrum

In general, spontaneous emission is the dominant source of the frequency noise in fixed-wavelength lasers and can be modeled with a white spectrum for the laser frequency. In tunable lasers, the frequency noise has an additional component from the tuning process, which shapes the spectrum following the tuning transfer function. Tuning processes based on carrier injection or the thermal properties of the laser cavity (thermal expansion or thermo-optic effects) exhibit first-order transfer functions. The bandwidth of the tuning transfer function for thermally-tuned lasers is usually limited from the kHz to MHz range, whereas with carrier injection, multiple tens of MHz is achievable. In MEMS-tunable lasers, with one dominant mechanical resonance mode for the MEMS mirror, the tuning transfer function has a second-order shape with a bandwidth defined by the mechanical design of the mirror.

In this analysis, the spectrum of the photocurrent in an FMCW lidar employing a tunable laser is derived. It is assumed that the laser tuning has a first-order transfer function with a bandwidth of ω_p. The frequency noise of the laser is assumed to have a DC level of f_n^2 Hz^2/Hz, dominated by the tuning noise. For the

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purposes of this analysis, the effect of the spontaneous emission is considered negligible.

As the first step, assuming a linear frequency modulation for the electric field at the output of the tunable laser, this electric field can be written in the following form:

$$e(t) = e_0 \cdot \cos\left(\omega_0 t + \frac{\gamma t^2}{2} + \varphi_n(t)\right), \qquad (18)$$

where e_0 is the amplitude of the electric field, ω_0 is the central frequency of the laser, γ is the modulation slope, and $\varphi_n(t)$ is the laser's phase noise at time t. For a round-trip delay of τ_R to the target, the waveform of the photocurrent at the coherent receiver will have the following form:

$$i(t) = i_0 \cdot \cos(\varphi_0 + \gamma\tau_R t + \Delta\varphi_n(\tau_R, t)), \qquad (19)$$

where φ_0 is a constant phase component equal to $\omega_0\tau_R + \gamma\frac{\tau_R^2}{2}$, $\gamma\tau_R$ is the frequency of the beat signal, and $\Delta\varphi_n(\tau_R, t)$ is the phase noise difference between the two interfering light fields with a time separation of τ_R:

$$\Delta\varphi_n(\tau_R, t) = \varphi_n(t + \tau_R) - \varphi_n(t). \qquad (20)$$

In [17], it has been shown that the autocorrelation of the photocurrent in (19) can be written in the following form:

$$R_i(u) = \frac{i_0^2}{2}\cos(\gamma\tau_R t)\, e^{\frac{-\sigma_\theta^2(\tau_R,u)}{2}}, \qquad (21)$$

where the term $\sigma_\theta^2(\tau_R, u)$ is the contribution of the laser's phase noise φ_n to the autocorrelation of the photocurrent and can be written as:

$$\sigma_\theta^2(\tau_R, u) = 2\sigma_{\Delta\varphi_n}^2(\tau_R) + 2\sigma_{\Delta\varphi_n}^2(u) - \sigma_{\Delta\varphi_n}^2(u + \tau_R) - \sigma_{\Delta\varphi_n}^2(u - \tau_R), \qquad (22)$$

with $\sigma_{\Delta\varphi_n}^2(a)$ defined as $\langle|\varphi_n(t + a) - \varphi_n(t)|^2\rangle$. In [17], the value of $\sigma_{\Delta\varphi_n}^2(a)$ was replaced with $\Delta\omega|a|$, which is the result of assuming a white frequency noise spectrum for the laser with a linewidth of $\Delta\omega$. In this work, $\sigma_{\Delta\varphi_n}^2(a)$ will be calculated without this assumption and the results from (22) and (21) will be used to find the power spectral density of the photocurrent. The frequency noise spectrum shown in

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Fig. 5 can be written in the following form:

$$S_{f_n}(\omega) = \frac{f_n^2}{1+(\omega/\omega_p)^2} \quad [\mathrm{Hz}^2/\mathrm{Hz}]. \qquad (23)$$

The phase noise spectrum can then be written as:

$$S_{\varphi_n}(\omega) = \frac{1}{\omega^2} \cdot \frac{4\pi^2 \cdot f_n^2}{1+(\omega/\omega_p)^2} \quad [\mathrm{rad}^2/\mathrm{Hz}]. \qquad (24)$$

The transfer function from the laser phase to $\Delta\varphi_n(a)$ is the subtraction of the two copies of φn by delay a, which can be written in the form of $1 - \exp(-j\omega a)$. By applying this transfer function to the laser phase noise spectrum given in (22), the spectrum of $\Delta\varphi n(a)$ can be written in the following form:

$$S_{\Delta\varphi_n(a)}(\omega) = \left|1 - e^{-j\omega a}\right|^2 \cdot \frac{1}{\omega^2} \cdot \frac{4\pi^2 \cdot f_n^2}{1+(\omega/\omega_p)^2} = \mathrm{sinc}^2(\omega a/2) \cdot \frac{1}{\omega^2} \cdot \frac{4\pi^2 \cdot a^2 \cdot f_n^2}{1+(\omega/\omega_p)^2} \; [\mathrm{rad}^2/\mathrm{Hz}]. \quad (25)$$

To find $\sigma^2_{\Delta\varphi_n}(a)$, the spectrum given above should be integrated over all frequencies. The result of this integral is as folllows:

$$\sigma^2_{\Delta\varphi_n}(a) = \frac{4\pi^2 f_n^2}{\omega_p}\left[a \cdot \omega_p - 1 + \cosh(a \cdot \omega_p) - \sinh(a \cdot \omega_p)\right]. \qquad (26)$$

For the laser used in this work, with $f_n^2 = 2.3\times10^6\ \mathrm{Hz}^2/\mathrm{Hz}$ and $\omega_p = 2\pi\times60\ \mathrm{Mrad/s}$, the plot of $\sigma^2_{\Delta\varphi_n}(a)$ is shown in Fig. 21. The graph based on the assumption of a white frequency noise spectrum is also shown on the same figure for comparison. As can be seen on this plot, the effect of a limited frequency noise bandwidth on $\sigma^2_{\Delta\varphi_n}(a)$ is significant, particularly for the values of $|a| \ll 1/\omega_p$.

The result of (26) can now be combined with (21) and (22) to find the autocorrelation function and consequently the spectrum of the photocurrent. While this can be done using the accurate form of (26) followed by numerical techniques, it is intuitive to use a simplified approximation for this equation and find a closed-form solution for the spectral density of the photocurrent. Based on the values of the delay variable a, (26) can be divided into two different regions: for small values of $a \cdot \omega_p$, the hyperbolic functions can be expanded in the first few terms of their Taylor series; and for large values of $a \cdot \omega_p$, it can be shown that the value of $\cosh(a \cdot \omega_p) - \sinh(a \cdot \omega_p)$ approaches zero exponentially. Therefore, (26) can be approximated in the following form:

$$\sigma^2_{\Delta\varphi_n}(a) = \frac{4\pi^2 f_n^2}{\omega_p} \begin{cases} \frac{(|a|\cdot\omega_p)^2}{2} - \frac{(|a|\cdot\omega_p)^3}{6} & |a| \le \omega_p^{-1} \\ |a|\cdot\omega_p - {}^2/_3 & |a| > \omega_p^{-1} \end{cases} \quad (27)$$

where the constant for the large values of $a \cdot \omega_p$ has been adjusted ($1 \rightarrow 2/3$) to maintain continuity at $a \cdot \omega_p = 1$. Using this approximation, and for the values of $\tau_R \le \omega_p^{-1}$, (22) can be evaluated as:

$$\sigma^2_{\Delta\varphi_n}(a) = \frac{4}{\xi_R} \begin{cases} |u| - \frac{\tau_R}{3} & |u| \le \omega_p^{-1} \\ \frac{1}{\omega_p} - \frac{\tau_R}{3} & |a| > \omega_p^{-1} \end{cases} \quad (28)$$

where $\xi_R = (\pi^2 f_n^2 \omega_p^2 \tau_R^2)^{-1}$, defined for algebraic simplicity as a function of the laser parameters and the target distance. This result combined with (21) determines the autocorrelation function. The Fourier transform of the this function around the carrier $(f = \gamma \cdot \tau_R)$ is the spectral density of the photocurrent:

$$S_i(\omega) = \frac{i_0^2}{2}\left\{e^{\frac{-2}{\xi_R\cdot\omega_p}} \cdot \delta(\omega) + \frac{\xi_R}{1+(\xi_R\omega_p/2)^2} \cdot \left[1 - e^{\frac{-2}{\xi_R\cdot\omega_p}} \cdot \left(\cos\left(\frac{\omega}{\omega_p}\right) + \frac{1}{\xi_R\cdot\omega} \cdot \sin\left(\frac{\omega}{\omega_p}\right)\right)\right]\right\}. \quad (29)$$

The first term is a Δ function that contains the signal information, and the second term is the spectral density of the noise. It is worth noting two important points in this equation. First, unlike the analysis based on the assumption of a white frequency spectrum for the laser [17], the spectral density of the photocurrent for $\tau_R \le \omega_p^{-1}$ is heavily dependent on the bandwidth of the laser's frequency noise spectrum, ω_p. For thermally- or mechanically-tuned lasers, in which the tuning bandwidth is smaller than the bandwidth of the DBR laser used in this work, the effect of the colored noise spectrum will be even more pronounced. For MEMS-tunable lasers, the frequency noise spectrum will have a second-order roll-off and may include peaking in its transfer function, which should be carefully considered when evaluating $\sigma^2_{\Delta\varphi_n}(a)$ in (26).

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Second, the amplitude of the signal drops exponentially as a function of τ_R^2, whereas based on the assumption of a white frequency spectrum, the degradation of the signal with increasing target distance would occur as an exponential function of τ_R.

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In practice, because of the limited duration of the modulation ramp, only time-windowed versions of this signal can be measured. To account for this effect, the resulting spectrum should be convolved with a sinc function as shown below:

$$S_i(\omega) = \frac{i_0^2}{2}\left\{e^{\frac{-2}{\xi_R\cdot\omega_p}}\cdot sinc^2\left(\frac{T_{ramp}\cdot\omega}{2}\right) + \frac{\xi_R}{1+\left(\xi_R\omega_p/2\right)^2}\cdot\left[1 - e^{\frac{-2}{\xi_R\cdot\omega_p}}\cdot\left(\cos\left(\frac{\omega}{\omega_p}\right) + \frac{1}{\xi_R\cdot\omega}\cdot\sin\left(\frac{\omega}{\omega_p}\right)\right)\right]\right\}. \quad (30)$$

Typically, in an FMCW lidar with linear frequency modulation, $T_{ramp} \gg \frac{1}{\omega_p}$, and as a result, the function $\mathrm{sinc}^2\left(\frac{T_{ramp}\omega}{2}\right)$ has a much narrower spectral width than the noise and can be considered as a Δ function when being convolved with the noise spectrum. From this reasoning, the effect of windowing on the noise spectrum can be neglected.

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