High speed analog-to-digital conversion with silicon photonics

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High Speed Analog-to-Digital Conversion with Silicon Photonics

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ABSTRACT

Sampling rates of high-performance electronic analog-to-digital converters (ADC) are fundamentally limited by the timing jitter of the electronic clock. This limit is overcome in photonic ADC’s by taking advantage of the ultra-low timing jitter of femtosecond lasers. We have developed designs and strategies for a photonic ADC that is capable of 40 GSa/s at a resolution of 8 bits. This system requires a femtosecond laser with a repetition rate of 2 GHz and timing jitter less than 20 fs. In addition to a femtosecond laser this system calls for the integration of a number of photonic components including: a broadband modulator, optical filter banks, and photodetectors. Using silicon-on-insulator (SOI) as the platform we have fabricated these individual components. The silicon optical modulator is based on a Mach-Zehnder interferometer architecture and achieves a $V_{\pi L}$ of 2 Vcm. The filter banks comprise 40 second-order microring-resonator filters with a channel spacing of 80 GHz. For the photodetectors we are exploring ion-bombarded silicon waveguide detectors and germanium films epitaxially grown on silicon utilizing a process that minimizes the defect density.

Keywords: Electronic photonic integrated circuits, silicon photonics, high index contrast, optical sampling, optical analog-to-digital conversion

1. INTRODUCTION

Rapid progress in CMOS technology combined with advances in parallel computing architectures has made Teraflop digital processors a reality. However, the wealth of new system capabilities offered by such processors cannot be fully exploited due to the limited performance of electronic analog-to-digital converters (ADCs). Performance of ADCs at high sampling rates is fundamentally limited by the timing jitter of the electronic clocking circuits [1] and the jitter performance of electronic oscillators, which is currently around 100 fs for state-of-the-art on-chip electronic oscillators. Fig. 1 illustrates a microwave signal with amplitude $V_0$ and period $T_0$ and the fundamental relationship between signal resolution $\Delta V/V_0$, expressed in number of bits $N$ and timing jitter $\Delta t$ involved in the sampling process. For a 20 GHz analog signal, the resolution in number of bits translates into the timing jitter values shown in Table 1 [1]. Current
electronics can barely achieve 70 fs jitter. Therefore, high resolution, > 6 bit, sampling of microwave signals with greater than 20 GHz bandwidth is beyond current electronic capabilities.

In this paper, we demonstrate photonic technologies, such as femtosecond lasers, optical integration and multiplexing enabling the miniaturization of high repetition rate mode-locked lasers, electro-optic conversion, filter and detector technologies that can be combined to overcome the electronic bottleneck in ADC [2,3,4,5]. Femtosecond lasers provide a stream of sampling pulses with much reduced timing jitter, when compared to integrated microwave oscillators, approaching attosecond jitter levels over milliseconds of measurement time [6]. The approach taken here towards a photonic ADC is based on wavelength division multiplexing to parallelize and demultiplex the stream of sampling pulses into lower rate channels, that can then be digitized with conventional electronic ADCs [5].

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<th>Targeted A/D resolution</th>
<th>Required sampling jitter</th>
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<td>10-bit</td>
<td>4 fs</td>
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<tr>
<td>8-bit</td>
<td>18 fs</td>
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<td>6-bit</td>
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Although microwave sources with much better timing jitter performance than 70 fs exist, those sources are typically bulky, because of the necessary high quality microwave cavities involved, which makes them difficult to integrate in a sampling system. In addition, there is the difficulty of preserving such low jitter during the sampling process itself. Figs. 2(a) and (b) show the behavior of a microwave signal and optical pulse train, respectively, emitted from an ensemble of such sources, where the phase of the microwave signal or the pulse position in a mode-locked laser undergo a random walk due to the fundamental noise sources in the generation process.

In the steady state of a high-quality oscillator, both electrical and optical cases based on a high-Q resonator, the gain element compensates for the losses of the cavity including output coupling. The resonator losses can be characterized by a cavity decay time \( \tau_{\text{cav}} \). For a microwave oscillator, the dissipation-fluctuation theorem requires that the loss-channels of the cavity couple a noise energy equivalent to \( kT \) into the cavity within one cavity decay time, otherwise the cavity itself cannot maintain thermal equilibrium when isolated. An ideal gain element at least doubles that noise, while compensating for the signal decay due to the resonator losses. Assuming that this noise energy is due to random white noise coupling into the cavity, one may easily derive that this addition of random noise leads to a random walk in phase of the signal with a linear increase in the timing jitter variance given by Eq. (1). Thus the timing jitter is related to the oscillation period and the relative noise added to the intracavity energy stored in the resonator mode, \( W_{\text{mode}} \), per cavity decay time. In other words, if one would introduce as much noise into the cavity as there is stored energy within a cavity decay time, the phase of the signal could be changed as much as \( 2\pi \) within that time.
Therefore, it should not be surprising, that the corresponding timing jitter of an optical pulse train from a mode-locked laser, which generates a short optical pulse of duration $\tau$ and pulse energy $W_{\text{pulse}}$ circulating within the optical cavity (see Fig. 2(b)) obeys Eq.(2) [6,7]

$$\frac{d}{dt} < \Delta t^2_{\text{RF}} > = \frac{\tau^2}{4\pi^2} - \frac{1}{W_{\text{mode}}} \frac{kT}{\tau_{\text{cav}}}$$  \hspace{1cm} (1)

$$\frac{d}{dt} < \Delta t^2_{\text{ML}} > = \frac{\pi^2 \tau^2}{6} \frac{2}{W_{\text{pulse}}} \frac{\hbar \omega}{\tau_{\text{cav}}}$$  \hspace{1cm} (2)

Of course, real microwave oscillators or mode-locked lasers may have additional noise sources or nonlinear dynamics that may further increase the timing jitter indicated in Eqs. (1) and (2), but the fundamental scaling should be preserved.

Due to the operation of the laser at optical frequencies, the noise energy added within a cavity decay time is $\hbar \omega$, quantum noise, rather than thermal noise scaling with $kT$. This makes optical amplifiers and sources much more noisy than their microwave counterparts, because at typical optical frequencies, $\omega$, corresponding to 1.55 $\mu$m, $\hbar \omega \sim 50kT$.

However, for a laser generating $\tau=100$ fs pulses, the timing jitter of the optical source (Eq. (2)) is scaled down by a factor of 1000 when compared with a 10 GHz microwave oscillator having a period $T_0 = 100$ ps, which easily overcompensates for the increased noise by more than two orders of magnitude assuming all other parameters stay constant and similar.

It is this favorable fundamental scaling of the timing jitter of femtosecond lasers with pulse width and cavity finesse and the possibility of integrating such lasers with superior timing jitter on a chip, that makes photonic ADCs a viable concept to pursue.

The layout of the photonic ADC pursued at MIT is shown in Fig. 3 and is based on wavelength-demultiplexing to downscale the overall sampling rate to parallel channels operating at lower rates [5]. A low-jitter pulse train with repetition rate $f_R$ generated by a mode-locked laser passes through a dispersive fiber, such that there is continuous coverage of the time axis. This chirped pulse train is modulated by a Mach-Zehnder (MZ) modulator whose RF driving voltage $V(t)$ is the signal to be sampled. The modulator effectively imprints the time dependence of $V(t)$ onto the optical spectrum. The optical signal is demultiplexed into $N$ channels by a filter bank, so that every pulse is split into $N$ sub-pulses. Each sub-pulse is detected by a photodetector and digitized by an electronic ADC taking one sample per pulse. This sample represents the RF signal at time moment $t_n = \tau(\omega_n)$, where $\omega_n$ is the filter center frequency. $N$ samples $V_{\text{ADC}}(t_n)$ are obtained, which are spaced uniformly across the repetition period $T_R$. This approach allows improvement of the sampling rate over what is available in electronic ADCs by a factor of $N$. By using both outputs of the MZ modulator we can also linearize its transfer function which is otherwise sinusoidal and factor out pulse-to-pulse amplitude fluctuations [4].

![Fig. 3. High-speed, high resolution optical sampling system. An integrated low-jitter femtosecond laser, currently planned on a separate chip, with repetition rate $f_R=1/T_R$ is emitting a stream of 100-200 fs pulses that is spectrally dispersed in regular single-mode fiber. Dispersion is chosen such that the chirped pulses cover the time interval between pulses with a smooth spectrum. The optical](image)
spectrum is limited by a bandpass filter. The RF-waveform to be sampled is imprinted on the chirped pulse stream via a dual-port silicon-based electro-optic modulator. The differential optical output is channelized via a dual-WDM-filter bank with precisely-tuned center frequencies that map each optical frequency component to certain sampling time slots. The signals from each channel correspond to time interleaved sample sequences are then separately digitized in low-rate high-resolution ADC’s. The silicon photonics chip may comprise the silicon optical modulator, filter banks, detector arrays, the low-rate electronic ADCs and feedback circuitry that is necessary to stabilize the optical filter bank.

2. INTEGRATED FEMTOSECOND LASERS

The proposed photonic ADC concept hinges on a compact, if possible, integrated femtosecond laser with low timing jitter. The approach towards such a device taken here is based on a planar light wave circuit using an Er-doped waveguide as the amplifier and a III-V semiconductor saturable Bragg reflector (SBR) as the passive mode-locker, see Fig 4(a). The reason for choosing Er-doped waveguide technology is two-fold. The long upper state lifetime of Erbium, 10 ms, ensures that relative intensity noise of the laser, which may couple to timing jitter, is only relevant in the low frequency range < 100 kHz. If necessary, fluctuations in that frequency range can be easily combated by feedback electronics. Second, the host material is glass, where two-photon absorption is absent in contrast to semiconductor materials, where two-photon absorption and other nonlinear optical effects strongly impact femtosecond pulse formation at useful average power levels.

Early attempts at passively mode-locked lasers using an erbium-doped waveguide as a gain medium and an SBR as a mode-locking element were limited to lower repetition rate and picosecond pulse operation [8,9]. Femtosecond operation of such lasers [10] was only achieved by employing nonlinear polarization rotation modelocking, which is difficult to integrate especially at high repetition rates. In addition, in all of these cases the cavity comprised free-space optics in between the SBR and the erbium-doped waveguide. Here, we demonstrate a 394 MHz, self-starting, passively mode-locked femtosecond laser based on planar silica waveguide technology. The laser generates 440 fs pulses with an average output power of 1.2 mW for a pump power of 400 mW [11]. The low efficiency is due to currently high internal losses, which can be minimized in the future.

The schematic laser layout is depicted in Fig. 4(a). The laser cavity consists of a 5 cm section of erbium doped alumino-silicate waveguide with a group-velocity dispersion of 30 fs²/mm. A 20 cm section of phosphorous-doped silica waveguide with a dispersion of -25 fs²/mm is used to obtain a net anomalous intracavity dispersion to enable soliton mode-locking [12]. The Er-doped section and the P-doped silica-glass section have an effective mode area of 10 μm² and 40 μm², respectively. A loop mirror is used at one end to provide 10% output coupling, while the other end is butt-coupled to an external SBR. The SBR is a commercial device with 14% modulation depth, a 2 ps recovery time, and a saturation fluence of 25 μJ/cm². Pump power is provided by an external 980 nm laser diode coupled into the waveguide chip. The laser was operated with 400 mW of cavity-coupled pump power; the intracavity signal power was measured to be 12 mW, corresponding to a 30 pJ intracavity pulse energy. The output pulses with an average power of 1.2 mW are then amplified to 18 mW using an EDFA (980 nm pump, 350 mA), detected using a 10 GHz photodiode, and measured with a 500 MHz sampling scope and a signal source analyzer (Agilent E5052).

![Fig. 4. a) Integrated laser layout. Inset on the left depicts the SBR’s reflection and dispersion spectra. b) Picture of the laser setup.](image-url)
Fig. 5. Measurement traces at 400 MHz: a) normalized optical spectrum before and after amplification, b) RF spectrum (3 GHz span, 10 MHz resolution), d) RF spectrum (100 kHz span, 500 Hz resolution), d) 10 second persistence trace, e) background free autocorrelation trace and f) single side band (SSB) phase noise of the first harmonic of the laser.

Fig. 5 depicts the measurement results. The persistence trace, see Fig. 5(d) shows excellent signal stability, while the RF spectrum in (c) indicates a side-mode suppression ratio of 80 dB. The 8.4 nm FWHM optical bandwidth before amplification implies 300 fs duration transform-limited pulses. After amplification, the optical bandwidth decreases to 7.4 nm, corresponding to 340 fs. The autocorrelation measurement yielded a pulse duration of 440 fs. The difference is attributed to incomplete compensation of the chirp added by the erbium-doped fiber amplifier. The laser was self-starting; as the pump power is increased, the laser first operates in a mode-locked Q-switching state before transitioning to a continuous-wave soliton mode-locked state at a pump power of 160 mW. Fig. 5(f) shows the phase noise of the first harmonic (394 MHz) of the laser. The timing jitter integrated from 20 MHz progressively down to 1 kHz is also shown. The timing jitter integrated from 10 kHz to 20 MHz is 24 fs, which is close to the noise sensitivity limit of the Agilent Agilent E5052 signal analyzer. Currently work is in progress to further shorten the pulse duration to the 100-200 fs range, to post amplify and interleave the pulse train to achieve the necessary output power levels and 2 GHz repetition rate necessary for the planned photonic ADC. We also expect that the timing jitter can be further reduced to the few-femtosecond or even sub-femtosecond range [13].

3. SILICON ELECTRO-OPTIC MODULATORS

Another important component of the sampling system is a high-speed silicon optical modulator that transfers the electronic signal into the optical domain. Fig. 6(a) illustrates the Mach-Zehnder interferometer device architecture used in these experiments. The interferometer arms contain diode sections (Fig. 6(b)) whose index can be changed by electronic carrier injection and the resulting plasma effect [14]. The output light intensity is modulated by the phase shifts induced in each arm of the interferometer by changes in carrier density. This type of modulator can be operated either as a forward biased PIN diode or as a reverse biased PN diode. In forward biased operation the device is more sensitive, but under reverse bias the modulator exhibits greater bandwidth [15]. The ADC requires the bandwidth of reverse bias operation, and we concentrate on those results here.

The modulators were fabricated with a standard CMOS-compatible lithography process. The sidewalls of the waveguide are moderately doped to a concentration of $10^{19} \text{cm}^{-3}$, and the center of the waveguide is doped n-type to a lighter concentration of $2\times10^{17} \text{cm}^{-3}$. These doping concentrations were chosen to produce a diode that would simultaneously cause a reasonable phase shift without excessive optical loss [16]. When operated in reverse bias, a
depletion region forms at the junction at one side of the waveguide. As the reverse bias voltage is increased, this depletion region becomes larger and extends into the center of the waveguide. A highly doped \((10^{19} \text{ cm}^{-3})\), 50-nm-thick silicon layer connects the waveguide to electrical contacts located 1 \(\mu\)m from the waveguide. Even higher doping concentrations of \(10^{21} \text{ cm}^{-3}\) are used under the metal contacts to assure good ohmic contacts.

By employing a relatively short interaction length of 0.5 mm (the travel time of the light is only 5 ps), a simple lumped-element electrode design can be used to achieve speeds of about 30 GHz. The modulator can be operated in a push-pull configuration by driving the center electrode (as shown), or a single arm can be driven. An additional thermal phase shifter (not shown) is fabricated on one arm, to allow the relative optical phase of the two arms of the modulator to be adjusted independently of the diodes. An adiabatic 4-port 3 dB coupler [17] is used to interfere the two arms at the output of the Mach-Zehnder interferometer leading to two complementary modulator output ports. To provide efficient fiber-to-chip coupling, a silicon inverse-taper mode converter combined with a lower index silicon oxynitride waveguide [18] was used.

To model the carrier concentration in the waveguide, numerical simulations were performed by solving Poisson and carrier continuity equations. The modeled change in carrier concentration determines a change in the refractive index of silicon which can then be used to determine the change in the modal index, \(N_{\text{eff}}\). The change in the modal index can then be used to determine the phase shift for a given diode length.

The DC response of the modulator was tested by varying the DC bias voltage on arm #1 while the other arm remained at the same bias level. Fig. 7 shows the simulated and measured normalized light intensity as a function of the applied voltage, for the two complementary outputs. The measured and simulated results are in good agreement with each other. To get a better measure of \(V_{\pi L}\), a longer device, with 5 mm arm lengths, was also measured. The output of that modulator (which was fabricated with only one output channel) is also shown in Fig. 7, and gives a \(V_{\pi L}\) of just over 4 Vcm. This is comparable to what had been achieved earlier with a structure requiring more complicated fabrication [19]. Operating the modulator in a push-pull configuration has also been demonstrated, and cuts the effective \(V_{\pi L}\) for the device in half to 2 Vcm, as expected.

Fig. 8 shows the small signal frequency response of the modulator after correcting for frequency dependent cable losses. A 3-dB cut-off frequency of 26 GHz is reached in a 0.5 mm device, which agrees with the estimate calculated above.
shorter device with 0.25 mm long phase shifters was also measured, and exhibits between 30-40 GHz of bandwidth. At high frequencies, the electrode can no longer be considered a lumped-element device, leading to irregularities such as the secondary peak at 40 GHz in the 0.5 mm device. The sensitivity of the device was measured by operating the 0.5 mm device in a push-pull configuration. In this configuration, 46 mW of RF power achieves ~22% modulation depth, (where the modulation depth is defined in [20]), for a 26 GHz sinusoid. In an optically sampled ADC, this modulation depth is sufficient to achieve a signal to noise ratio of 47 dB [20]. This is only 6 dB less than the theoretical maximum of 53 dB achieved with a 50% modulation depth under otherwise similar conditions.

![Normalized Intensity vs Applied Voltage](image1)

**Fig. 7.** Measured and simulated normalized light transmission as a function of applied voltage.

![Frequency Response](image2)

**Fig. 8.** Measured frequency response of the modulator.

### 4. HIGH INDEX CONTRAST FILTER BANK

After the incoming electric signal has been imprinted onto the femtosecond-laser pulse by the MZ modulator, the signal traveling out of each output arm must be demultiplexed. Filter banks consisting of microring resonators, utilizing high-index contrast materials, are ideal for this function. High-index-contrast microring resonators have been shown to achieve the large free-spectral range (FSR) and low loss required for these filter banks, while maintaining wavelength scale dimensions required for chip-level integration [21,22]. The main challenge in the fabrication of these filter banks is achieving the dimensional precision, on the tens of picometer scale, required to control the resonant frequency spacing between each filter of the filter bank. The proposed ADC design requires the output from each arm of the MZ modulator to pass through identical twenty-channel filter banks comprised of second-order microring resonator filters. The microring resonators are designed to have 2 THz FSR and 25 GHz bandwidth, allowing for a channel spacing of 80 GHz with less than 30 dB of crosstalk.

Previously, we reported the results of a twenty-channel dual filter bank that used silicon-rich silicon nitride (SiN$_x$), $n=2.2$ at 1550 nm, as the core material (Fig. 9) [23]. The resonant frequency spacing of this filter bank was controlled by changing both the ring radius and average waveguide width of each microring resonator by the appropriate amount. This required changing the average waveguide width on a scale smaller than the grid size of the scanning-electron-beam lithography (SEBL) tool, with a precision of tens of picometers. This was achieved by modulating the electron-beam dose used to write each filter to change the average waveguide width on a scale much finer then the SEBL grid size [23]. The average channel spacing of the filter bank was measured to be 83 GHz with a standard deviation of 8 GHz. This 3 GHz error in average spacing corresponds to dimensional precision of the average waveguide width of 75 pm, much smaller than the 6 nm step size of the SEBL address grid.

Looking at the frequency response of the filter bank in Fig. 9 it is seen that there is a rather large drop loss (~12 dB on average) and the bandwidth of the filters is much larger than the target 25 GHz. The two reasons for this are that there is an average frequency mismatch of 29 GHz between the two rings of the same filter and a higher than expected...
propagation loss in the SiN_x. The frequency mismatch causes both increased drop loss and widening of the filters bandwidth. This can be easily corrected through postfabrication thermal trimming using integrated microheaters. Fig. 10 demonstrates how a microheater can be used to reduce this mismatch to less than 1 GHz in a two-channel filter bank. After thermal trimming the frequency response for the two filters shown are nearly identical. Likewise, if both rings of the filter are heated simultaneously it is possible to shift the resonant frequency response of the filter, with an efficiency of 80μW/GHz/filter, to correct for any frequency spacing errors [24]. As shown in Fig. 10 even after correcting for the frequency mismatch there is still ~8 dB of loss in each filter. This loss is due to material absorption in the SiN_x core. The SiN_x used is a non-stoichiometric material optimized for very low stress. The optical properties of this material, including material absorption, varies from batch to batch. One way to solve this problem is to switch to a SOI platform, the single crystal silicon layer used for the filters will not vary in its properties and is known to have low optical absorption.

Fig. 9. Measured frequency response for a dual twenty-channel filter bank with all non-filter related losses factored out.

Fig. 10. Transmission response of a two channel filter bank before and after thermal trimming.

To make the change from SiN_x to Si both the filter design and fabrication process must be altered. The Si filter design chosen uses a very low aspect ratio, 105 X 600 nm and 105 X 500 nm, for the ring and bus waveguide, respectively, instead of the typical Si single mode waveguide dimensions of ~220 X ~500 nm. This waveguide design reduces the filter’s frequency sensitivity to width to 40 GHz/nm, from the high sensitivity of ~100 GHz/nm. This makes the Si
filter design only slightly more sensitive to changes in width than the SiN\textsubscript{x} design despite the much higher index contrast. In addition to the changes in design, the basic fabrication process has to be changed because the process for the SiN\textsubscript{x} filters uses a Ni hardmask, formed via a PMMA lift-off process, during the reactive-ion etching (RIE) step. It is found that using metal hardmasks during RIE of Si can lead to a silicide formation on the waveguide sidewalls, causing excess propagation loss [25]. The new process uses hydrogen silsequioxane (HSQ), a negative electron-beam resist, as both the resist and the hardmask. HSQ transforms into a SiO\textsubscript{2} like hardmask after being exposed by an electron-beam and developed in TMAH. The HSQ pattern is then transferred into the Si using RIE with HBr gas, which has a very high selectivity between SiO\textsubscript{2} and Si. After etching, the HSQ is stripped with a dilute HF dip. The devices are then cleaned with an RCA clean and then overclad with 1 \textmu m of HSQ that is annealed at 400\textdegree C for 1 hr in an oxygen atmosphere to turn it into SiO\textsubscript{2} (Fig. 11(a)) [26]. Titanium resistive heaters are then fabricated above the HSQ using contact lithography and a lift-off method. The propagation loss of these Si waveguides is measured to be 2.5 dB/cm over the wavelength range of 1520-1610 nm.

This low loss in the Si waveguides makes it possible to fabricate the small bandwidth, high FSR filters that are required for the filter bank while maintaining minimal drop loss. Fig. 11(b) shows the fabricated Si filter design before being overclad with HSQ. The measured and fitted transmission response for this filter design is shown in Fig. 11(c). The drop loss of this filter is less than 0.5 dB, much lower than the 8 dB that was measured in the SiN\textsubscript{x} filters. The fitted bandwidth is measured to be 21 GHz, only slightly smaller than the designed bandwidth of 25 GHz. The frequency mismatch between rings in this filter is fitted to be less than 0.25 GHz and the crosstalk between neighboring channels 80 GHz away is less than -30 dB. This filter response meets all the requirements for what is needed for the ADC’s twenty-channel filter bank.

Since the change to Si required changing electron-beam resists from a positive resist, PMMA, to a negative resist, HSQ, empirical experiments were performed to find the new frequency dependence on electron-beam dose. These experiments were performed by writing both a reference filter and a calibration filter that share the same through port for each variation of dose. The dose used for the reference filter is kept constant while the dose used for the calibration filters is varied. The frequency dependency on dose is then found by comparing the relative frequency shift between each reference and calibration filter [23]. From these experiments it is found that the frequency dependence increases from 14 GHz/% for the SiN\textsubscript{x} filters to 66 GHz/% for the Si filters. This increase is more than can be explained by the difference in frequency sensitivity for the two designs. The extra sensitivity is most likely explained by the difference in contrast of the two e-beam resists. It is also seen that the standard deviation of the fitted frequency calibration data of the Si filters is approximately a factor of 10 larger than for the SiN\textsubscript{x} filters. This is not a big concern since the thermal trimming efficiency of the Si filter is about a factor of 10 more efficient. This means the total power needed to correct for frequency errors in a Si filter bank should be about the same as for the SiN\textsubscript{x} filter banks.
As mentioned earlier the proposed ADC needs two identical filter banks, one for each output arm of the MZ modulator. One way to insure that the two filter banks are identical is to use the same filter bank for both arms. This can be done since one ring resonator supports two optical modes with orthogonal propagation directions, clockwise and counterclockwise. As long as the sidewall roughness of the ring waveguide is low enough there will be no significant coupling between the two modes [27]. To check the possibility of using this method a single ring filter was fabricated with the layout shown in Fig. 12(a). The frequency response for the propagating modes as well as the crosstalk of this filter was measured (Fig. 12(b)). The transmission responses of both modes are exactly the same as is expected. Looking at the crosstalk at first it seems that there is a large amount of crosstalk between the two counter-propagating modes. Looking more carefully it is seen that this “crosstalk” is actually the sum of both a through and drop response. There are two ways that this can happen either the power that is not dropped by the filters bounces off the end-facet and then travels back towards the filter and is dropped or the power that is dropped bounces of the end-facet travels back to the filters and hits a through response. Both of these possibilities can be eliminated by reducing the reflections at the end-facets. To put an upper limit on the crosstalk in the filter the drop response is subtracted from the crosstalk giving a through response. The dip at the resonant frequency of the through response is below -30 dB showing that the crosstalk is at least less than -30 dB. This shows that it is possible to use one filter bank (Fig. 12(c)) to demultiplex both arms of the MZ modulator as long as great care is taken to prevent any reflections in the system.

5. SILICON AND GERMANIUM INFRARED PHOTODIODES

Both germanium and silicon photodiodes are being pursued for this system. The Ge diode technology is well established, but requires careful attention to process integration issues when used in a silicon CMOS process. In pursuit of easier process integration, we have developed a novel all-silicon infrared diode device.

In 1959 Fan [28] reported that radiation damaged Si would produce a photocurrent when illuminated with sub bandgap radiation. Knight [29,30] and others [31,32,33] have used this result to make waveguide optical detectors. Unlike most other detectors formed in Ge or InGaAs where the optical absorption is > 1000 dB cm⁻¹, the Si waveguide photodiodes used for the ADC applications discussed here exhibit only ~20 dB cm⁻¹ absorption. A fact that needs special attention. Here, we summarize the recently achieved device performance; the fabrication, optical absorption, and quantum efficiency are given in more detail elsewhere [33,34].
The waveguide photodiodes are of similar structure as shown in Fig. 6, the RC time constant for a 50 Ω load is ~60 ps cm⁻¹, while the optical transient time determined by the group velocity of light is 130 ps cm⁻¹. Thus the optical absorption of the waveguide detectors directly impacts the tradeoff between the optical response and the electrical bandwidth due to the necessary device length. As the waveguide length is increased a larger fraction of the incoming light is absorbed and converted into an electrical signal. However, the increased propagation time of the light through the longer device reduces the electrical bandwidth. Fig. 13 shows the trade off between the fraction of light absorbed in the detector and the transient time limited bandwidth for several absorption coefficients. What fraction of the absorbed light actually generates a photocurrent or is lost to some other mechanism does not effect this bandwidth absorbed light tradeoff. A traveling wave photodiode structure would have a larger bandwidth, but only half of the photo response. To absorb half of the incoming light a waveguide-photodiode-length of 1.5 mm is required, which limits the bandwidth to 22 GHz. Although only half of the incoming light is absorbed in the diode, the internal photoresponse of ~ 0.4 to 10 A W⁻¹ implies an external quantum efficiency of 0.2 to 5 A W⁻¹ is achieved. Even if the waveguide length is increased to 5 mm, which will absorb 90% of the incoming light, the transit-time bandwidth is ~ 10 GHz, an acceptable value for our ADC application.

![Fig. 13. Calculated maximum 3 dB RF power bandwidth as limited by light propagation time through the waveguide photodetector as a function of the fraction of the light absorbed for several optical absorption coefficients. An optical group velocity of 7.5x10⁹ cm s⁻¹ was assumed for these calculated results.](image)

To determine the intrinsic photodiode bandwidth, the frequency response of a 100-μm-long waveguide photodiode was measured. This short diode absorbs only 4.5% of the incoming light but the 100 μm length implies a propagation-time-limited bandwidth of > 200 GHz. The frequency response of the diode was determined using a measurement system containing a LiNbO₃ Mach-Zehnder optical modulator with a bandwidth of ~34 GHz and a commercial InGaAs photodiode with a bandwidth > 50 GHz, used as a reference. The response of the measurement system with the Si diode and the standard diode are shown in Fig. 14. The normalized photodiode response was obtained by taking the ratio between the measured response of the Si and the InGaAs-reference detectors and is shown in Fig. 15. This procedure eliminates the response of the modulator, RF losses in the cables, and the bias tee. However the response of the RF probes used to contact the Si waveguide photodiode cannot be compensated for. The Si-diode response agrees with the...
InGaAs-reference within a few dB up to 30 GHz and stays close even up to 40 and 50 GHz. The calculated RC-time-constant bandwidth of the contact pad capacitance and the 50 Ω input impedance of the network analyzer is ~ 100 GHz. Also a ~90 GHz bandwidth was calculated considering only the transit time of holes and electrons at their saturated velocity of 1x10^7 cm s⁻¹. The RF power to the optical modulator was increased with operating frequency to compensate for reduced response. Measurements were made at different RF power levels to ensure that the modulator and photodiodes were operating in their linear region.

One challenge of a wideband analog system is that one must ensure that the second-order distortions are small, not just the third-order. This is because, by definition, a wideband system has enough bandwidth for octaves of the measured signal to be present in the measurement bandwidth which will corrupt the signal estimation. We measured the nonlinear response of the silicon waveguide ion-bombarded photodiodes using a three-tone measurement setup [35] in which three independent DFB lasers were modulated at f₁ (380 MHz), f₂ (620 MHz), and f₃ (2.000003 GHz). A useful feature of this technique is that, by judicious choice of frequencies, the second-order nonlinear distortion at f₁ + f₂ can be placed very close to the third-order nonlinear distortion f₃-(f₁+f₂) and a narrowband RF filter will pass the distortion components to the measurement system while removing the strong fundamental components—this is very challenging using only a two-tone measurement setup. More importantly, this technique does not rely on any frequency components of type N·f ±M·f where N and M are integers > 1.

Fig. 16 shows the power of the fundamental compared to the second-and third-order distortion products as a function of photocurrent for two different bias voltages. The second and third harmonic distortion increased with increasing bias voltage, and the second harmonic distortion is observed to be more severe than the third harmonic.

In addition to Si photodetectors, germanium photodiodes are under investigation. Blanket Ge-on-Si films have a relatively high threading dislocation density due to the 4% lattice mismatch between Ge and Si. To reduce the threading dislocation density, Ge can be grown selectively in oxide openings. For the test devices in this work, the starting material was 6” p+ Si wafers with patterned silicon dioxide films. Using an Applied Materials low-pressure chemical vapor deposition epitaxial reactor, the germanium was grown to overfill the openings to ensure that the oxide openings were completely filled. A chemical-mechanical polishing step was then performed to planarize the Ge film. The samples were then implanted with phosphorus and subjected to a rapid thermal anneal at 600°C for 5 s to activate the implant. Contact vias were etched and metal contacts were deposited to form vertical pin diodes. A cross-sectional schematic of the device is shown in the inset of Fig. 18.

DFB lasers were modulated at f₁ (380 MHz), f₂ (620 MHz), and f₃ (2.000003 GHz). A useful feature of this technique is that, by judicious choice of frequencies, the second-order nonlinear distortion at f₁ + f₂ can be placed very close to the third-order nonlinear distortion f₃-(f₁+f₂) and a narrowband RF filter will pass the distortion components to the measurement system while removing the strong fundamental components—this is very challenging using only a two-tone measurement setup. More importantly, this technique does not rely on any frequency components of type N·f ±M·f where N and M are integers > 1.

Fig. 15. Measured frequency response of a 100-μm-long waveguide Si photodiode using the difference between a high bandwidth standard diode and the Si diode. The data for both diodes individually is shown in Fig. 14. The increase in response between 40 and 50 GHz is believed to result from RF probes used to contact the Si diode, which are rated to 40 GHz. More details of this measurement are given elsewhere [34].
Fig. 17 shows plan view transmission electron microscopy (TEM) images for various sized Ge features. For the as-grown Ge layers, the threading dislocation density decreases with device width from $5.5 \times 10^8 \text{ cm}^{-2}$ to $2.6 \times 10^8 \text{ cm}^{-2}$ to $1.7 \times 10^8 \text{ cm}^{-2}$ for 5, 1, and 0.65 \text{ μm}-wide devices respectively. This effect is consistent with previous results on aspect ratio trapping [36] and a small area effect [37, 38]. Fig. 16 also illustrates that a cyclic anneal reduces the dislocation density. Analysis of multiple plan view TEM areas indicates that the dislocation density was reduced from $1.7 \times 10^8 \text{ cm}^{-2}$ to $2.6 \times 10^7 \text{ cm}^{-2}$ in the 0.65 \text{ μm}-wide islands by cyclic annealing.

Fewer threading dislocations are expected to improve the yield and reduce the dark current of these devices. Fig. 18 shows current versus voltage characteristics of vertical \textit{pin} diodes fabricated in 0.35 x 10 μm germanium islands for both an as-grown Ge film and a film cyclically annealed between 830°C and 450°C. These are representative characteristics of rectifying devices. After processing, it was found that about 50% of the devices displayed rectifying behavior and the others were shorted. The impact of material quality and device processing on yield requires further investigation. The dark current for both devices is ~0.5 nA at -1 V. While the cyclic anneal reduced the threading dislocation density in these devices, this did not have a significant effect on the dark current. This indicates that the leakage in these smaller devices is dominated by another mechanism and more work is needed to fully characterize it. The cyclic anneal did have an effect on the on current, however. The device that received the anneal has ~2x higher on current at 1 V than the device without the anneal. This is most likely due to a reduction in the series resistance.
6. CONCLUSIONS

We have presented the system architecture and fabricated key elements of a photonic ADC. The fabricated devices include: a 394 MHz waveguide femtosecond laser with 24 fs timing jitter, a Si modulator capable of modulating speeds faster than 25 GHz, high performance Si-microring filters for a demultiplexing filter bank, silicon and germanium photodiodes. Using such devices for the proposed ADC system allows higher sampling speeds than current pure electronic systems due to the low timing jitter of the femtosecond laser. The size-scale, materials, and processing methods for these devices make them suitable for integration on an electronic-photonic silicon-on-insulator platform.

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