

# A Hybrid CMOS Photonic RF Channelizer: System Design and Analysis

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**Abstract**—Wireless perception requires processing of broadband spectrum by Artificial Intelligence (AI) pipelines. However, direct sampling of other electronics-only solutions is unsuitable for simultaneous processing of low-RF to FR3 band (24 GHz). We propose combining CMOS electronic circuits with silicon-based photonic integrated circuits (PICs) to enable novel broadband RF photonic channelizers. This work presents a system-level design and analysis of a channelizer using on-PIC optical filter banks and optical-domain demodulation.

**Index Terms**—Analog optical link, Photonic Integrated Circuit (PIC), RF Photonics, RF-to-optical modulator, Silicon Photonics.

## I. INTRODUCTION

THE demand for wideband operation in future wireless communications, the Internet of Things (IoT), imaging, sensing, and electronic warfare necessitates advanced spectrum management through integrated electronics. AI-driven distributed systems further require low-cost, power-efficient solutions for processing wideband signals.

Direct sampling of the entire spectrum of interest from DC to 24GHz and channelization in the digital domain is the architecturally simplest approach. However, this will require a Nyquist rate analog-to-digital converter (ADC) with over 48 GHz sampling rate. Such ADCs, realized using time-interleaved architectures, are prohibitively expensive and power-intensive with their effective number of bits (ENOB) limited by the sampling jitter to around 6-bits. In principle, an analog filter bank could be used to decompose a wideband signal into multiple narrower sub-bands, which could then be downconverted to the baseband using a bank of RF mixers. Alternatively, mixer-bank-based [1] or frequency-folded [2] architecture can be employed. However, these electronic architectures are limited to below 1GHz bandwidth.

Besides limited bandwidth, the key limitation of traditional analog approaches is that active IC-based signal processing introduces significant noise and linearity challenges. Our approach explores recent breakthroughs in photonic integrated circuit (PIC) technology to overcome the inherent bandwidth, noise, and linearity constraints of conventional on-chip analog electronic sub-banding techniques for wideband spectrum.

The advent of Silicon Photonic (SiP) integrated circuits and their fabrication on multi-project wafer (MPW) platforms is enabling the integration of complete optoelectronic systems within a single chip-scale package. RF photonics, a well-

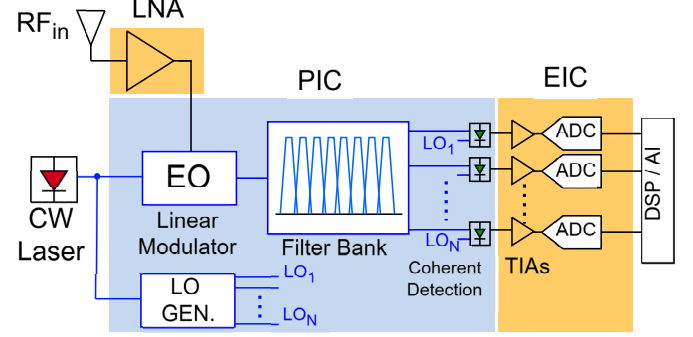


Fig. 1. RF Photonic channelizer architecture.

established field, processes RF signals using discrete lasers, modulators, and photodetectors before transmission over optical fiber. Beyond the inherent advantages in size, weight, power, and cost (**SWaPc**), novel PICs offer solutions to longstanding limitations of discrete optoelectronic systems.

This article presents the system-level design of an integrated RF photonic channelizer for broadband spectrum perception. The remainder of the manuscript is structured as follows: Section II details the overall architecture and circuit components. Section III provides noise analysis and simulation results outlining fundamental trade-offs in the on-PIC link budget, followed by the conclusion.

## II. RF PHOTONIC CHANNELIZER ARCHITECTURE

This work investigates an RF photonic channelization from near DC to the edge of the FR3 band (24GHz), and down-conversion architecture as shown in **Fig. 1**. Most signal processing is performed in the optical domain on a PIC. In this coherent processing architecture, the continuous-wave (CW) laser at 1550nm is split into two paths: one for modulation and another for optical local oscillator (LO) generation. The input RF spectrum is amplified using a broadband low-noise amplifier (LNA). The RF modulation of the laser is performed using a lithium niobate electro-optic (EO) modulator for high linearity.

An optical filter bank then processes the modulated optical spectrum realized using arrayed waveguide grating (AWG) with  $N=8$  channels of  $B=3$  GHz spacing, ranging from DC to 24 GHz. The sub-banded channels from the PIC realization of a wideband filter bank can be sub-sampled using an ADC with  $\geq B$  sampling rate for achieving wideband functionality through multiple sub-bands in the digital signal processor

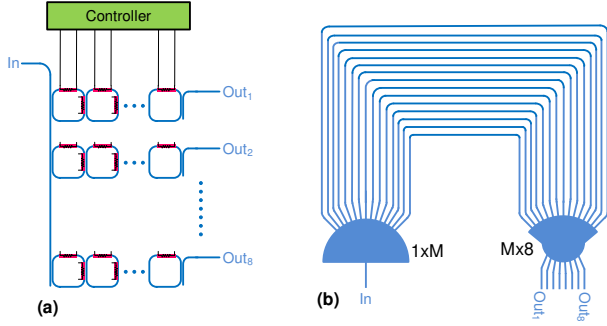


Fig. 2. (a) Channelizer based on CROW filters, (b) Arrayed Waveguide Grating (AWG). Replace this by your new plot from Lumerical Interconnect.

(DSP). However, the curse of sub-sampling is that it suffers from noise aliasing effect, *i.e.*, where out-of-band noise is aliased into the band of interest). Moreover, this would require the transimpedance amplifiers (TIA) with sufficient bandwidth to process the entire  $N \times B$  bandwidth, with a resulting reduction in optical sensitivity due to broadband thermal noise ( $kTNB$ ).

Optical-domain downconversion is an alternative approach, where coherent detection is performed for each channel. However, this approach would require  $N$  optical LOs to be used with the  $90^\circ$  optical hybrid. Finally, the downconverted signals are in the form of photocurrents, which are then processed by linear TIAs. The advantage of this approach is that the linear TIAs need to process only  $B$  bandwidth with lower in-band noise ( $kTB$ ) and significantly reduced power consumption.

#### A. RF to Optical Modulation

In the proposed system, the input RF signal is first routed through a wideband and linear LNA with a bandwidth of 24 GHz. Even though some RF photonic systems avoid the LNA in the interest of linearity, the LNA improves the link noise figure (NF) while compensating for the inherent photonic losses, ensuring that the signal maintains a sufficient signal-to-noise ratio (SNR) for subsequent processing stages. Following amplification, the signal is directed to a Mach-Zehnder Modulator, which converts the RF signal into the optical domain. However, Dual-Parallel Mach-Zehnder Modulator (DP-MZM) is proven to have better linearity over the standard MZM, as it cancels even order distortion and with low third-order intermodulation distortion (IM3) resulting in a desirable spur-free dynamic range (SFDR)  $> 100 \text{ dB}^{2/3}$  across the entire range of spectrum of interest [3] [4]. Notably, the DP-MZM supports single-sideband (SSB) modulation.

#### B. Optical Filter Bank

The broadband RF-modulated optical signal is processed by a photonic filter bank as seen in Fig. 1. Optical filters can be designed on a PIC with high selectivity and  $> 20 \text{ dB}$  out-of-band suppression. Since passive optical circuits are inherently linear and broadband, they far surpass the filtering achieved in electronic realization. The photonic filter bank can be realized using a cascaded Mach-Zehnder interferometer (MZI) or ring-based filters. Ring-based filters can use coupled-resonator optical waveguides (CROW), shown in Fig. 2(a), where resonators are

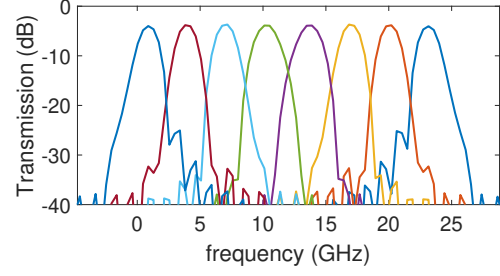


Fig. 3. Simulated AWG spectrum for  $N=8$  channels with separation of 3GHz.

coupled through waveguides to synthesize sharp optical filter responses [5]. While CROWs are compact and on-PIC filters with automatic tuning have been demonstrated [6], an 8-channel filter bank will need  $\geq 24$  rings. Tuning coupling and resonances of each ring, especially with on-chip thermal crosstalk, will lead to a complex controller design. MZI-based channelizers are attractive but exhibit lower out-of-band rejection [7].

In our architecture, we propose to use arrayed waveguide grating (AWG) shown in Fig. 2(b), which require a smaller number of tuning elements and exhibit lower loss. The AWG splits the input into  $M$  waveguides with progressively increasing group delay of  $\Delta L$ . These waveguides are combined with a free-space  $M \times N$  star coupler, resulting in the  $N=8$  equally spaced outputs. The frequency spacing is given by

$$\Delta f_{\min} = \frac{c}{n_g N \Delta L} \quad (1)$$

where  $n_g$  is the group index of the waveguides. Table I compares recent PIC-based AWGs for RF photonic applications. For GHz frequency spacing, the PIC area becomes very large with associated losses and cost [8]. Recent work used a low-loss silicon nitride (SiN) process, associated with a large bend radius ( $> 100 \mu\text{m}$ ) leading to a large PIC area [9]. In our architecture, we are using Si waveguides in an ultra-low-loss SiP process with  $< 1 \text{ dB/cm}$  loss. The simulated spectral response is shown in Fig. 3. This allows tighter bends ( $10 \mu\text{m}$ ) with compact serpentine routing of the waveguides leading to a manageable estimated PIC area of  $4 \text{ mm} \times 3 \text{ mm}$ .

TABLE I  
COMPARISON OF SiP AWGs FOR RF PHOTONIC APPLICATIONS.

Ref.	Channel Spacing	Num of Channels	Num of WGs	Technology	Area
[8]	50 GHz 10 GHz 1 GHz	11	35	$1 \mu\text{m}$ Si WG	$4 \text{ mm} \times 4 \text{ mm}$ $11 \text{ mm} \times 4 \text{ mm}$ $9 \text{ mm} \times 12 \text{ mm}$
[9]	3.9 GHz	10	33	Low-loss SiN WG	$12.5 \text{ mm} \times 19.5 \text{ mm}$
This work	3 GHz	8	16	Low-loss $0.5 \mu\text{m}$ Si WG	$4 \text{ mm} \times 3 \text{ mm}$ (Estimated)

#### C. Optical Coherent Downconversion

Coherent optical detection beats the optical signal,  $P_s$ , with the optical local oscillator (LO) signal,  $P_{LO}$ , resulting in an output current [10]

$$I_{out}(t) = 2\rho \sqrt{P_s(t) \cdot P_{LO}} \quad (2)$$

Coherent downconversion is performed using a  $180^\circ$  coupler and a pair of photodiodes as depicted in Fig. 4 (a) [11].

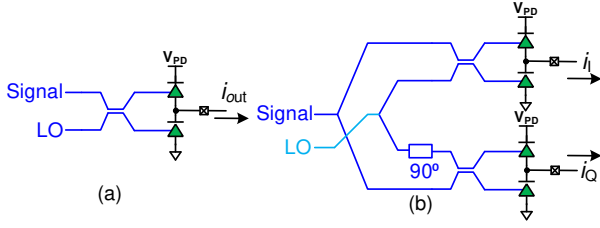


Fig. 4. Balanced Detection Implementations. (a) In-phase Balanced Detection. (b) I/Q Balanced Detection.

However, a small phase mismatch between LO and Signal leads to random scaling of the downconverted signal. Using the balanced detection scheme, seen in **Fig. 4** (b), results in in-phase and quadrature (I/Q) components,  $\rho$  being the detector responsivity:

$$I_I(t) = 2\rho\sqrt{P_S(t) \cdot P_{LO}} \cdot \cos(\theta_{sig}(t) - \theta_{LO}) \quad (3)$$

$$I_Q(t) = 2\rho\sqrt{P_S(t) \cdot P_{LO}} \cdot \sin(\theta_{sig}(t) - \theta_{LO}) \quad (4)$$

Digital Signal Processing (DSP) techniques are used to recover both magnitude and phase data from the I/Q signals [12]. Alternatively, analog domain squaring and summing can be used to recover the envelope as  $I_{env} = \sqrt{I_I^2 + I_Q^2} = 2\rho\sqrt{P_S(t) \cdot P_{LO}}$  [13]. I/Q Balanced Detection is used in the proposed system.

#### D. Linear TIAs

A linear transimpedance amplifier (TIA) with high linearity and low noise is crucial for accurately converting photodetector currents to the voltage domain. **Fig. 5** shows a linear TIA circuit realized in 65nm CMOS technology. A resistive feedback CMOS inverter topology is used for linearity followed by linear programmable gain stages and output buffers to interface with off-chip ADCs [14]. A DC offset cancellation (DCOC) loop is needed to set the output common-mode voltage and cancel any mismatches of the PDs. Preliminary simulations shown in **Fig. 6** predict a TIA gain of  $R_T=63$  dB $\Omega$  for a  $B=3$ GHz bandwidth with an input referred noise of  $i_{n,TIA} = 3.1$  pA/ $\sqrt{Hz}$ .

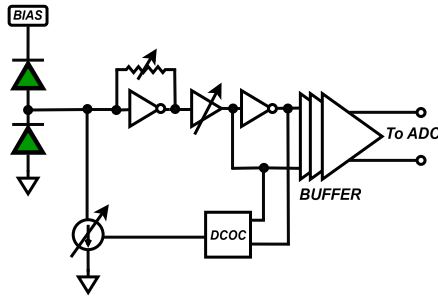


Fig. 5. A linear TIA circuit to process the downconverted optical signals.

### III. SYSTEM-LEVEL NOISE ANALYSIS

In this section, we estimate the noise contributions from the individual blocks in the channelizer and analyze their impact on the overall performance.

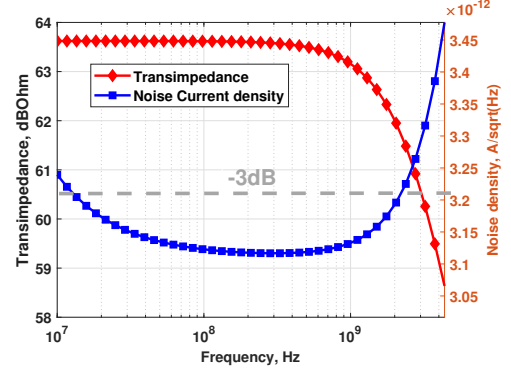


Fig. 6. Simulated TIA gain and noise response.

#### A. Noise sources

##### 1) MZM noise

The MZM is impedance-matched with the RF source. Neglecting the electrode loss, the MZM noise contribution in each channel is estimated as  $kTB$  [15].

##### 2) Photodetector noise

Photodetectors (PDs) generate shot noise and thermal noise in the RF photonic front end. Shot noise is dependent on the DC component in the output current,  $I_D$ , and is given by  $P_{shot} = i_{sn}^2 B = 2qI_D B$ . The thermal noise contribution is negligible [15], [16].

##### 3) Laser Relative Intensity Noise

The laser's Relative Intensity Noise (RIN) is due to the random emission of the stimulated photons. The effect of RIN is observed in the photodetector output and depends on  $I_D^2$ . The RIN noise contribution is given as [15], [16]

$$P_{RIN} = i_{rin}^2 B = 10^{(\frac{RIN}{10})} I_D^2 B \quad (5)$$

##### 4) TIA noise

The input referred noise from the TIA,  $P_{n,TIA} = i_{n,TIA}^2 B$ , includes the thermal and flicker noise contributions [17].

#### B. Balanced Detection Noise Analysis

**Fig. 7** depicts the noise model for the entire channelizer. The noise in a balanced detector is estimated differently from an intensity-modulated link. The shot noise and RIN noise depend on the DC component of the subtracted output currents. If the same laser source is used as desired, the RIN noise in both the PDs are closely correlated. Therefore the RIN noise currents will be subtracted. Since the shot noise in the two PDs is uncorrelated, the currents will be added [18].

Since the DC current in both the PDs is due to the LO and signal power ( $P_s = \frac{T_{ff} P_L}{L_c}$ ), the shot and RIN noise powers can be written as [19]

$$\begin{aligned} P_{RIN} &= 10^{(\frac{RIN}{10})} P_{LO}^2 (\rho_1 \kappa - \rho_2 (1 - \kappa))^2 B \\ &= 10^{(\frac{RIN}{10})} P_{LO}^2 [\rho \Delta \kappa + \Delta \rho (1 - \Delta \kappa)]^2 \cdot B \rightarrow 0 \end{aligned} \quad (6)$$

$$\begin{aligned} P_{shot} &= 2qP_{LO} (\rho_1 \kappa + \rho_2 (1 - \kappa)) B + \frac{\kappa P_s B}{2} \kappa P_s / 2 \\ &\approx 2qP_{LO} \rho B + \frac{\kappa P_s B}{2} \end{aligned} \quad (7)$$

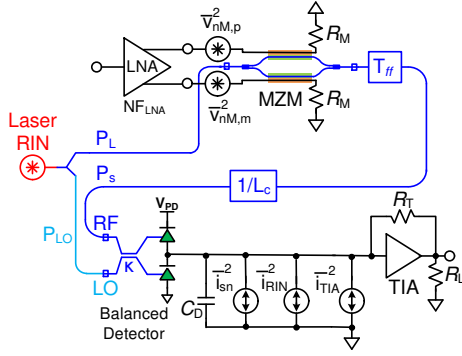


Fig. 7. Small-signal noise model for the RF Channelizer system.

Here,  $\rho_{1,2}$  are the PD responsivity and  $\kappa$  is the splitting ratio of the 3dB coupler. The total noise at the TIA input is:

$$P_{n,tot} = 2qP_{LO}\rho B + \frac{\kappa P_s B}{2} + P_{RIN} + i_{n,TIA}^2 B \quad (8)$$

We estimate the signal-to-noise ratio (SNR) of the balanced detection as:

$$SNR = \frac{4\rho^2 \cdot P_s(t) \cdot P_{LO}}{P_{n,tot}} \quad (9)$$

With ideal coupler ratio of  $\kappa = 0.5$  and matched PDs, the RIN noise will diminish. However, the SNR will degrade with any coupler and/or PD mismatch and can be expressed as function of the common-mode rejection ratio (CMRR) of the balanced detection downconverter [20].

### C. On-PIC Link Gain

The MZM's RF gain is given by [21]

$$G_M = \frac{s_{mz}^2}{R_s} = \left( \frac{\pi T_{ff} P_L R_s}{2V_\pi} \right)^2 \frac{1}{R_s} \quad (10)$$

where  $T_{ff}$  captures the insertion loss,  $P_L$  is the input laser power,  $V_\pi$  is the voltage needed for  $\pi$  phase shift in the MZM, and  $R_s$  is the reference impedance (i.e. 50Ω).  $G_c = \frac{1}{L_c}$  is the channelizer insertion loss. The combined detector and TIA gain, assuming  $R_L = R_s = 50\Omega$  load, are estimated as

$$G_{det} = 4\kappa^2 \rho^2 P_{LO} \frac{R_T^2}{R_s} \approx \rho^2 P_{LO} \frac{R_T^2}{R_s} \quad (11)$$

The overall on-PIC link gain is given by  $G = G_M \cdot G_c \cdot G_{det}$  and expressed as

$$G = \left( \frac{\pi T_{ff} P_L \rho \sqrt{P_{LO}}}{2V_\pi L_c} \cdot R_T \right)^2 \quad (12)$$

Evidently, the link gain is improved by using higher laser and LO powers, high performance MZM (i.e. lower  $V_{pi}$  and insertion loss), and lower AWG insertion loss.

### D. On-PIC Link Noise Figure

The noise figure (NF) and noise factor (F) are given by [15]

$$NF = \log_{10}(F) = \log_{10} \left( 1 + \frac{N_a}{GN_i} \right) \quad (13)$$

Where  $N_a$  is the added noise power at the link output,  $N_i$  is the noise at the input, and  $G$  is the link power gain. The input noise power is due to the source resistance and is given by  $N_i = kTB$ . The added noise,  $N_a$ , components include the MZM thermal noise contribution of  $G \cdot kTB$ , and  $P_{n,tot}$  components from Eq. 8. We first calculate the link NF without the LNA:

$$F_{noLNA} = 1 + \frac{kTG + (i_{rin}^2 + i_{sn}^2 + i_{n,TIA}^2) \cdot R_T^2 / R_s}{G \cdot kT} \approx 2 + \frac{(i_{sn}^2 + i_{n,TIA}^2) \cdot R_T^2 / R_s}{G \cdot kT} \quad (14)$$

With an LNA in the front, the NF is

$$F = F_{LNA} + \frac{F_{noLNA} - 1}{G_{LNA}} \approx F_{LNA} + \frac{(i_{sn}^2 + i_{n,TIA}^2) \cdot R_T^2 / R_s}{G_{LNA} \cdot G \cdot kT} \quad (15)$$

### E. Simulation Results and Discussion

Fig. 8 plots Eqs. 14 & 15 with respect to the link power gain,  $G$  for link with and without LNA. The link parameters are  $P_L=20\text{dBm}$ ,  $P_{LO}=20\text{dBm}$ ,  $V_\pi=2\text{V}$ ,  $RIN=-165\text{dB/Hz}$ , MZM loss of 2dB, AWG loss of 3dB,  $R_s=50\Omega$ ,  $R_T = 63\text{dB}\Omega$ ,  $\rho=1\frac{\text{A}}{\text{W}}$ . The front LNA has 30dB gain and an NF of 3 dB.

As expected, the LNA improves the overall gain and NF by compensating for the optical losses and detector noise. For gain,  $G \gg 1$ , the NF asymptotically approaches the LNA gain. The overall link NF can be further improved by using higher input laser and/or LO laser powers to increase the MZM and detector gains respectively.

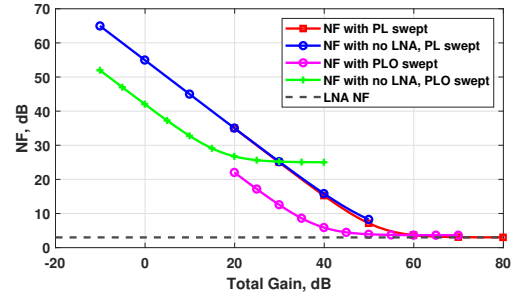


Fig. 8. NF versus gain for the on-PIC RF photonic link. The laser or LO powers are swept from 0 to 30dBm, while keeping the other at 20 dBm. Plots are shown with and without the front LNA.

## IV. CONCLUSION

In this work, we presented a noise and link analysis of a PIC-based RF photonic channelizer and demodulator. The analysis presented in this work informs the overall channelizer PIC and TIA co-design. The input and LO laser powers can be increased to compensate for optical losses, while a linear LNA helps improve overall NF. Future work entails PIC and CMOS chip tape-outs, copackaging, and testing.

## ACKNOWLEDGMENT

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