Experimental study of 112 Gb/s short reach transmission employing PAM formats and SiP intensity modulator at 1.3 μm

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Abstract: We present a Silicon Photonic (SiP) intensity modulator operating at 1.3 μm with pulse amplitude modulation formats for short reach transmission employing a digital to analog converter for the RF signal generator, enabling pulse shaping and precompensation of the transmitter's frequency response. Details of the SiP Mach-Zehnder interferometer are presented. We study the system performance at various bit rates, PAM orders and propagation distances. To the best of our knowledge, we report the first demonstration of a 112 Gb/s transmission over 10 km of SMF fiber operating below pre-FEC BER threshold of 3.8 × 10−3 employing PAM-8 at 37.4 Gbaud using a fully packaged SiP modulator. An analytical model for the Q-factor metric applicable for multilevel PAM-N signaling is derived and accurately experimentally verified in the case of Gaussian noise limited detection. System performance is experimentally investigated and it is demonstrated that PAM order selection can be optimally chosen as a function of the desired throughput. We demonstrate the ability of the proposed transmitter to exhibit software-defined transmission for short reach applications by selecting PAM order, symbol rate and pulse shape.

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References and links


1. Introduction

The need for faster and inexpensive short-reach optics for data centers is growing rapidly due to the incessant spread of cloud services offered by data centers. For an equal bit rate, modulating a single wavelength from a single laser provides a solution that is more cost effective than a hardware multiplexed counterpart which involves the use of multiple wavelengths or fiber lanes to achieve the total desired bit rate. In addition, the single wavelength approach is the more scalable solution to cope with the ever-increasing capacity demand. Although 100 Gb/s transmission of Ethernet frames over SMF has been standardized as 100GBASE-LR4 by using $4 \times 25$ Gb/s in a WDM setting employing four colors in the 1300 nm band [1], it is widely agreed that achieving 100 Gb/s on a single wavelength using inexpensive and power efficient components is a mandatory building block in order to realize the envisioned 400 Gb/s bit rate based on $4 \times 100$ Gb/s [2].

Recently, 100 Gb/s short-reach transmission experiments have been reported using various advanced modulation formats, such as 4-level pulse amplitude modulation (PAM-4).
with polarization division multiplexing [3], half-cycle Nyquist subcarrier modulated 16-ary quadrature amplitude modulation (16-QAM) with polarization division multiplexing [4], multi-band carrierless amplitude phase modulation (multi-CAP) [5], discrete multi-tone (DMT) [6] and a performance comparison between some of these formats has been also reported in [7]. More recently, 400 Gb/s was also experimentally demonstrated as $4 \times 100$ Gb/s using DMT and 4 directly modulated lasers (DMLs) over 30 km in [8], and using multi-CAP modulation and 4 integrated electroabsorption-modulation laser (EML) over 40 km in [9], both in the O-band.

In addition, Silicon photonics (SiP) is rapidly emerging as an inexpensive solution for a variety of applications, including coherent links [10] and short reach (<10 km) optical interconnects [11]. Silicon is a useful host material for fabricating the external modulator because of its low cost, high yield, large scale integration, and CMOS compatibility, enabling the integration of modulators with the RF drivers [12]. Recently in [13], a dual-drive SiP modulator operating at 30 GHz bandwidth near 1300 nm was demonstrated and used at a 50 Gb/s rate (OOK). More recently in [14], we reported a single-drive series push-pull SiP modulator at 1310 nm based on our work in [15] and presented some preliminary transmission results using PAM-4 modulation at 107 Gb/s. In fact, externally modulated lasers used in conjunction with SiP modulators in [14] not only provide a larger bandwidth than directly modulated lasers (DML), they also add the potential of chirp-free transmission.

In this paper, we present a detailed system-level experimental study carried out using the series push-pull SiP modulator working at 1310 nm that is presented in [14]. A bit rate of 112 Gb/s is achieved using eight-level pulse amplitude modulation (PAM-8) at 37.4 Gbaud after 10 km of SMF fiber below the hard decision forward error correction (FEC) threshold of $3.8 \times 10^{-3}$, to provide a final bit error rate under $10^{-15}$. To the best of our knowledge, this is the first demonstration of error-free amplitude modulated, 100 Gb/s transmission using a SiP modulator. We also present exhaustive system performance tests where we vary multiple parameters for multi-level PAM-N formats, namely the PAM order, the bit rate, the distance, the pulse shaping roll-off factor, and the received signal power. The DSP algorithms applied at the transmitter and receiver, allowing the high throughput and also a fully software-defined short reach link, are presented. Finally, a model for the $Q$-factor applicable for multilevel PAM signaling is presented and accurately validated experimentally against the BER performance metric.

The remainder of the manuscript is organized as follows. Section 2 presents the packaged silicon photonic intensity modulator designed to operate at 1310 nm. Section 3 details the experimental setup employed to demonstrate the use of the SiP modulator for short reach transmission using PAM-N modulation. Section 4 presents the digital signal processing applied at the transmitter and receiver for optimum transmission performance of PAM formats. In Section 5, we present a new model for $Q$-factor computation applicable to multilevel PAM-N formats, followed by comprehensive experimental system results. Section 6 concludes the manuscript.

2. Fully packaged SiP modulator at 1310 nm

The 1310 nm SiP modulator uses carrier depletion in pn junctions. It is based on a series push-pull configuration [16–18] and has been fabricated at IME [19] within the framework of the Canadian SiEPIC program [20]. The schematic and a photograph of the device are shown in Fig. 1 and Fig. 2, respectively. This kind of modulator has already been demonstrated in silicon at 1550 nm [16, 17]. The light is coupled into and out of the chip using surface grating couplers and guided in 220 (height) $\times$ 400 nm (width) strip waveguides. Multimode interference (MMI) couplers are used to split and recombine the light in a Mach-Zehnder configuration. In the active portion of the modulator, the light is guided over two rib waveguides (slab height of 90 nm) used to support the pn junctions. The series push-pull configuration is obtained by connecting in series the two pn junctions, which are driven by a single modulating signal. This series combination reduces the capacitance and increases the speed. Of course, in this configuration, the resistance of the series combination is increased as
well. In the center portion of the modulator, the two junctions are thus in contact, through a common n-doped region. A DC bias voltage is applied to this common n-doped region to ensure that the pn junctions are operated in reverse bias at all time. This configuration results in a push-pull action minimizing the chirp, although this is not a critical issue at the selected wavelength of operation.

The series push-pull configuration was proposed by Walker and first demonstrated in GaAs [18], and, later, in InP [21, 22]. In both GaAs and InP, the two diodes are vertically grown onto a n-type common substrate while in SiP, the pn junctions in series are readily formed onto the silicon substrate, without any epitaxial regrowth, requiring only rib waveguide patterning and doping. Compared to dual-drive modulators, series push-pull modulators require a single driver and no bias-tee (unless the former is used at null bias [13]). For high-speed operation, a RF traveling-wave coplanar stripline electrode is used, which is carefully designed to have an index close to that of the optical group index of the rib waveguide of 3.8, once loaded with the pn junction structure. The pn junctions are further segmented into 36 sections, for a modulator length of approximately 6 mm. The biasing scheme required by the series push-pull modulator also avoids DC current to be supplied to the line-matching termination as opposed to dual-drive modulators. In our design, the line termination is integrated on-chip (see Fig. 2). A value of 50Ω has been measured for the load. A small path mismatch of 150 μm is introduced in the interferometer, leading to a 2.6 nm Free-Spectral-Range, to ease the characterization of the modulator. Two thermal phase shifters (TPS) are also incorporated within the interferometer to adjust the operation point at quadrature.

Fig. 1. Schematic of the series push-pull Mach-Zehnder modulator.

Fig. 2. Photograph of the 6 mm long series push-pull modulator. The portion of the layout supporting the grating couplers is not shown.
The modulator optical transmission spectrum is shown in Fig. 3(a) for different sets of DC voltages applied to the two pn junctions (V_{d1} and V_{d2}). These voltage drops across the junctions are produced by applying properly adjusted DC sources, as illustrated in Fig. 3(b). One of these DC sources, V_{bias}, is applied to the bias electrode of the modulator (i.e. biasing the n-doped common region); the other DC source, V_{RF input}, is applied to the RF input electrode of the modulator. One can see that the optical spectrum of the modulator displays a good extinction ratio of more than 20 dB, indicating a good balance of the power from each arm after recombination by the output MMI coupler of the Mach-Zehnder interferometer (MZI). The set of curves shown in Fig. 3(a) reproduces the push-pull operation around a bias voltage of 4 V ((V_{d1} + V_{d2})/2 = 4 V) [21]. The blue curve shows the optical spectrum when the two diodes are polarized with 4 V. The purple and light blue curves display the result of decreasing the reverse voltage on one junction (V_{d1}) while increasing the reverse voltage on the other (V_{d2}). The red and green curves are obtained when the role of the two diodes are reversed. The analysis of these results allows for a determination of V_{π} equal to 6.5 V at a DC bias voltage of 4 V. For this nominal bias voltage, V_{π} varied from 5.7 V to 6.5 V with a mean of 5.9 V on seven chips characterized. By repeating these measurements for different sets of voltages, the DC V_{π} can be extracted as a function of the bias voltage. The result is shown in Fig. 3(c). It can be noted that a different procedure was used in [17] to obtain V_{π}. In that case,
a single voltage was applied to the pn junctions in series. Using this alternative method, the phase shift obtained is that shown in Fig. 3(d), from which a $V_{\pi}$ of about 4 V is deduced. It is important to note that this alternative method provides a $V_{\pi}$ value corresponding to an operation around a bias voltage of 0 V and does not permit a determination of $V_{\pi}$ at larger bias voltages.

The modulator optical loss is 4.6 dB (at maximum transmission) at a bias of 4 V, excluding the 9 dB fiber coupling loss. Figure 3(e) presents the loss dependency to bias voltage. The characterization of the thermal phase shifters, shown in Fig. 3(f), results in an efficiency of 50 mW/$\pi$.

The small-signal frequency response of the modulator with the on-chip load wire bonded is presented in Fig. 4(a) (blue curve). The curve is normalized to the response at the reference frequency of 1.5 GHz. The bias voltage is 4 V and the modulator is operated at the quadrature point. A bandwidth slightly above 20 GHz is obtained. The corresponding electrical reflection ($S_{11}$) spectrum is shown in Fig. 4(b) (blue curve), better than −10 dB over a 32 GHz range.

After the initial characterization discussed above, the chip was fully packaged as shown in Fig. 5. Optical fibers were attached to the chip using a method outlined in [15], without a noticeable insertion loss penalty. The input fiber is a polarization maintaining fiber. The chip is mounted in a package with access to the modulator electrode through a GPPO connector and also to the thermal phase shifters and DC bias connection points. The frequency response and $S_{11}$ characteristics of the packaged modulator are slightly degraded, as illustrated by the red curves in Figs. 4(a) and 4(b), respectively.

Fig. 4. (a) EO frequency response of the modulator chip (blue) and of the packaged modulator (red). (b) $S_{11}$ of the modulator chip and of the packaged modulator.

Fig. 5. Picture of the fully packaged SiP modulator.

3. Experimental setup

We present in Fig. 6 the experimental setup employed to demonstrate the use of the SiP modulator presented in the previous section for short reach transmission using PAM-N modulation. A 20 mW 1310 nm DFB laser with a measured linewidth of 440 kHz is
modulated by the SiP intensity modulator. The RF signals driving the modulator are first generated from a Digital to Analog Converter (DAC). This AC-coupled 8-bit DAC runs at 70 Gsamples/s and feeds the differential input of an Inphi IN3214SZ linear driver. The amplified RF signal is then applied to the RF input of the SiP modulator which was constantly operated at the quadrature point of the optical transfer function by appropriately tuning the thermal phase shifters. The modulated signal is launched into various lengths of single mode fiber: either 0, 2, 10 or 20 km of Corning SMF-28e fiber. An AC-coupled Picometrix PT-40D PIN + TIA receiver is employed before an Agilent DSO-X 96204Q real-time oscilloscope serving as an 8-bit Analog to Digital Converter (ADC) sampling at 80 GSamples/s.

The DAC has a 3 dB bandwidth of 15 GHz and a maximum output differential swing of 1.2 V pp. The linear driver has a gain of 17 dB, maximal single-ended output swing of 6 V pp, and a 3 dB bandwidth of 34 GHz. The 3 spools of different fiber lengths exhibit a loss of 0.34 dB/km at 1310 nm. The PIN + TIA receiver has a 3 dB bandwidth of 35 GHz and a conversion gain of 1.3 V/mW, and the real-time scope has a 3 dB bandwidth of 33 GHz.

In the forthcoming sections, we compare the performance of different PAM orders over variable fiber lengths in order to reach a 100G Ethernet transport rate. To achieve this information transfer rate, the transmitter needs to provide 112 Gb/s. From the OTU4 standard, a client payload of 99.5328 Gb/s is transmitted at a line rate of $\frac{255}{227} \times 99.5328 = 111.80997$ Gb/s [23], including a 6.7% overhead for the forward error correction (FEC). In this manuscript, we assume FEC encoding and decoding at the transmitter and receiver, respectively. FEC is required in currently deployed 100Gb/s metro and long-haul fiber optic transmission systems because it enables a significant reduction of the acceptable raw BER at the expense of signal overhead. The 6.7% FEC overhead in the OTU4 standard produces a net coding gain of 9.19 dB at a corrected BER of $10^{-15}$, translating to a pre-FEC BER threshold of $3.8 \times 10^{-3}$. Measured BER values below this threshold are regarded as ‘error-free’ in the context of optical transport networks [24, 25].

In this work we focus on a comparison of PAM-4 and PAM-8 as a 100G candidate, but also collect PAM-2 (On-Off Keying) and PAM-16 data for completeness. The 8-bit DAC, with its maximum memory of $2^{18}$ samples supports the generation of these differing PAM orders at varying symbol rates.

### 4. Digital signal processing at the transmitter and receiver sides

As the system employs a DAC and an ADC, digital signal processing (DSP) can be applied at both the transmitter and the receiver. This section describes the different DSP schemes applied at both ends of the transmission system, from symbol generation to symbol detection, in order to best generate and recover the desired waveforms.

#### 4.1 Transmitter DSP for PAM-\(N\) formats

We first introduce the required processing to apply at the transmitter on the multilevel data symbols. There are 4 processes to apply on the data symbols before the 8-bit waveform quantization: 3 linear processes and one nonlinear. The processes are applied sequentially on the symbols generated, where we assume a fixed sampling rate of 70 GSa/s. The symbols for PAM-\(N\) format come from an alphabet of integers from $-N+1$ to $N-1$ in steps of 2. The first
process, for a desired symbol rate of $R_b$ in Gbaud, is to upsample the symbol stream from 1 sample per symbol (SPS) to $70/R_b$ SPS. This upsampling is readily done in the frequency domain. Secondly, the desired pulse shaping filter is applied. This pulse shaping filter $h_{RRC}[n]$ is typically a root-raised cosine (RRC) filter. From the RRC pulse theory, the frequency (in GHz) at which the pulse shaping filter's power transfer is reduced by half is $\frac{1}{2}R_b$, and the maximum spectral content is located at $\frac{1}{2}R_b(1 + \alpha)$, where $\alpha$ is called the roll-off factor of the pulse shaping filter [26]. Consequently, for a desired symbol rate of $R_b$, the roll-off factor can only be within the range $0<\alpha<\min(1, \frac{70}{R_b}-1)$ in order to avoid aliasing and respect the Nyquist sampling theorem [27]. Thirdly, the inherent nonlinear raised cosine power transfer function of the SiP Mach-Zehnder intensity modulator is compensated. This process equalizes the spacing between the optical power levels after modulation by applying an arcsin function to the waveform. This nonlinear transform will be cancelled and linearized after the MZI with direct detection. Finally, the last step is to pre-compensate the frequency response of all the components at the transmitter that precede the SiP modulator. The concatenated linear analog response of the DAC, RF driver, and SiP modulator can all be compensated using one lumped inverse response function, $h_{inv}[n]$. It is important to mention that the main frequency constraining component on the transmitter side is the DAC with a 3 dB bandwidth of only 15 GHz. From the Nyquist sampling theory, the content out of the DAC can be manipulated anywhere form DC to 35 GHz, when sampling at 70 GSa/s. It can be difficult and unnecessary to fully equalize the frequency response all the way to 35 GHz, as 1) this frequency exhibits very low output swing and 2) the pulse shaping filter can be designed such that its maximum frequency $\frac{1}{2}R_b(1 + \alpha)$ is less than 35 GHz. Naturally, one drawback of this analog response compensation is a reduction of the output peak-to-peak voltage, leading to smaller driving signals for the SiP modulator, itself leading to reduced level separation of the optical PAM-$N$ signal. Figure 7(a) shows the sequence of DSP operations applied at the transmitter.

4.2 Receiver DSP for PAM-$N$ formats

At the receiver, the required signal processing to be performed off-line is straightforward, where the only major process is linear filtering. The processes are detailed as follows. First, we resample from the fixed ADC rate of 80 GSa/s to twice the symbol rate $2R_b$. Secondly, we apply the matched filter, defined at 2 SPS. The stream of samples is then filtered by a short linear FIR filter $h_{RX}$, with constant coefficients stored in the receiver. During first system startup, the coefficients of this receiver filter can be unknown, in which case all coefficients are set to “0” except the central tap which is set to “1”. The optimum coefficients that will later remain constant and be stored can be obtained in a blind decision-directed scheme using a short period of a few hundred symbols. Once the coefficients are determined, $h_{RX}$ is subsequently linearly applied to the input samples. The receiver also has to implement a digital clock recovery algorithm to recover the transmitter clock and apply symbol decision at proper sampling instants. Clock recovery at the receiver is used to compensate any sampling phase and frequency offset that may exist between the transmitter and receiver clocks. As both clocks are free running, a digital clock recovery is needed for residual clock offset removal. For this task, we employ the blind feedforward timing error estimator in [28] which operates
at 2 samples/symbol and is transparent to modulation format, operating for any PAM-N format. The timing error estimate provided by the estimator, updated on a block-by-block basis, is then used to control a piecewise parabolic interpolator in a recursive fashion similar to the one explained in [29]. The final output symbols are used to compute the desired performance metric, including the mean signal to noise ratio, the mean bit error rate or the equivalent Q-factor. Figure 7(b) shows the sequential DSP blocks applied at the receiver.

5. Performance and results

In this section, we present the metrics employed to assess the performance of the system. We first start by defining a Q-factor for multilevel PAM signal, followed by system performance tests. We use both this Q-factor and the BER from error counting metrics to assess performance.

5.1 Q-factor for PAM-N formats

System performance is often reported using the Q-factor metric, where the noise distribution is assumed to follow a Gaussian distribution. The metric is well suited for binary On-Off signaling, i.e. PAM-2, where the received sampled value fluctuates from symbol to symbol around an average value of μ₁ or μ₂ depending on whether the bit corresponds to 0 or 1 in the binary symbol stream, respectively, with respective noise variance for each level of σ₁² and σ₂². For this format, the BER is found to be [30]

\[
\text{BER}_Q = \frac{1}{2} \left( \frac{1}{2} \text{erfc} \left( \frac{\mu - t^h}{\sigma \sqrt{2}} \right) + \frac{1}{2} \text{erfc} \left( \frac{t^h - \mu}{\sigma \sqrt{2}} \right) \right).
\]  

(1)

where the optimum decision threshold \(t^h\) is

\[
t^h = \frac{\mu_2 \sigma_2^2 - \mu_1 \sigma_1^2 + \sigma_2 \sigma_1 \sqrt{(\mu_2 - \mu_1)^2 + 2(\sigma_2^2 - \sigma_1^2)(\ln(\sigma_2/\sigma_1))}}{\sigma_2^2 - \sigma_1^2}.
\]  

(2)

and where \text{erfc} is the complementary error function. With the assumption \((\mu_2 - \mu_1)^2 \gg (\sigma_2^2 - \sigma_1^2)\) in Eq. (2), \(t^h\) simplifies to \(t^h = (\mu_1 \sigma_2 + \mu_2 \sigma_1)/(\sigma_1 + \sigma_2)\) and the argument of each \text{erfc} of Eq. (1) become identical: \((\mu_2 - t^h)/\sigma_2 = (t^h - \mu_1)/\sigma_1\equiv Q\). This Q can be rewritten as \(Q = (\mu_2 - \mu_1)/(\sigma_2 + \sigma_1)\) and is also called Q-factor. By inverting Eq. (1), one can compute the Q-factor from the knowledge of the BER with.

\[
Q = \sqrt{2} \text{erfc}^{-1} \left(2\text{BER}_Q\right).
\]  

(3)
It is important to point out that $BER_Q$ in Eq. (1) with the subscript $Q$ is used to denote the theoretical BER when two levels $\mu_1$ and $\mu_2$ are corrupted by Gaussian noise with variances of $\sigma^2_1$ and $\sigma^2_2$, respectively. This $BER_Q$ is minimized when $P^b$ equals Eq. (2), at which point one can compute the $Q$-factor using Eq. (3). The reason we introduce the subscript $Q$ in $BER_Q$ is to distinguish this BER from the BER calculated from bit error counting denoted by $BER$. In the forthcoming sections, we will use two performance metrics namely $BER$ and $Q$-factor. The first metric, $BER$, represents the BER obtained from counting the number of bit errors. This counting is performed offline after all receiver DSP processes and with knowledge of the transmitted bit stream. The second metric, $Q$-factor, is evaluated using Eq. (3) with a newly defined $BER_Q$ that considers the statistics $\mu$ and $\sigma^2$ of all levels of the received multilevel PAM-$N$ format, instead of just the 2 levels in Eq. (1), as will be explained in detail hereafter. Finally, the $Q$-factors reported are always cast in dB scale, i.e. $10 \times \log_{10}(Q)$.

Consequently, we define here a $BER_Q$ model that considers the statistics $\mu$ and $\sigma^2$ of all levels of a multi-level format. Moreover, instead of the single optimum threshold $P^b$ given in Eq. (2) for ON-OFF keying, $N$–1 optimum thresholds have to be defined for a PAM-$N$ format, where each optimum threshold uses the statistics $\mu$ and $\sigma$ of its two closest levels. Using a similar approach as that for PAM-2 in Eq. (1), the analytical $BER_Q$ for PAM-$N$ formats is defined as

$$BER_Q = \frac{1}{\log_2(N)} \sum_{i=1}^{N} p(I_i) [P(I_{i-1} | I_i) + P(I_{i+1} | I_i)].$$

(4)

where $N$ is the PAM order, $I_i$ is symbol level $i$ out of $N$ possible levels, $p(I_i)$ is the probability of transmitting symbol level $i$, and finally $P(I_{i-1} | I_i)$ or $P(I_{i+1} | I_i)$ are the probabilities of deciding $I_{i-1}$ or $I_{i+1}$ when $I_i$ was transmitted, respectively. $1 / \log_2(N)$ is the number of erroneous bits when a wrong symbol decision is made, assuming Gray coding and that erroneous symbol decisions are made to either one of the two closest neighboring symbols. Or course, for the first and last symbol levels $I_1$ and $I_N$, erroneous symbol level decision can only be made on one side and $P(I_0 | I_1) = P(I_N | I_1) = 0$. As $p(I_i)$ is equiprobable for all levels $i$, $p(I_i) = 1 / N \forall i$. From the theory relating $P(I_{i \pm 1} | I_i)$ to the error function $erfc$ one can show that

$$BER_Q = \frac{1}{N \log_2(N)} \sum_{i=1}^{N} \left[ \frac{\mu_i - I_{i,low}^b}{\sigma_i \sqrt{2}} \right] + \left[ \frac{\mu_i - I_{i,high}^b}{\sigma_i \sqrt{2}} \right]$$

(5)

where $\mu_i$ is the average value of the received symbols that were transmitted as $I_i$, and $\sigma^2_i$ is their variance. Both $\mu_i$ and $\sigma^2_i$ are computed and obtained at the receiver before the hard decision. The quantities $I_{i,low}^b$ and $I_{i,high}^b$ are the optimum decision level thresholds between $I_{i-1}$ and $I_i$, and between $I_{i+1}$ and $I_i$, respectively. From this analytical $BER_Q$ obtained using solely the first 2 statistical moments of all levels, $\mu_i$ and $\sigma^2_i$, we compute the equivalent $Q$-factor using Eq. (3). This gives the equivalent $Q$-factor for multilevel PAM signaling from analytical $BER_Q$ derivation.

5.2 System performance

In this subsection, we present the system-level performance of the short-reach optical link that was depicted in Fig. 6. The system performance is assessed qualitatively with the aid of eye diagrams and quantitatively in terms of three performance metrics: $Q$-factor according to its definition in subsection 5.1, $BER$ from error counting, and signal-to-noise ratio (SNR) which is defined, for a PAM-$N$ signal, as the ratio of the average signal power to the average power of the noise riding on top of the signal. For a PAM-$N$ signal of possible values $-N+1$ to $N-1$ by jumps of 2, the mean signal power is $(N^2-1)/3$. For the average BER computation, a total
of $10^7$ bits were captured for each trace. Therefore, we assume the minimum BER that is reported with confidence is at around $100/10^7 = 10^{-3}$.

We first present in Fig. 8 a few sample eye diagrams for PAM orders of 2, 4, 8 and 16 at both low and high symbol rate, after 10 km of propagation. Eye diagrams are a good qualitative way to observe the quality of the received signal. The eyes are obtained after the receiver DSP presented in Fig. 7(b) is applied, that is after resampling the ADC output at $2R_{B}$ and applying the matched filter and linear receiver filter $h_{Rx}$ with clock recovery. To obtain multiple points within a symbol duration in the eye diagrams, we upsample the 2 SPS signal.

Fig. 8. Eye diagrams, after 10 km, for (a) PAM-2 at 30 Gb/s, (b) PAM-2 at 60 Gb/s, (c) PAM-4 at 60 Gb/s, (d) PAM-4 at 112 Gb/s, (e) PAM-8 at 60 Gb/s, (f) PAM-8 at 112 Gb/s, (g) PAM-16 at 50 Gb/s and (h) PAM-16 at 60 Gb/s. All eyes obtained after receiver DSP.
obtained after \( h_{Rx} \), without creating spectral content, by padding zeros in the frequency domain.

Above each eye diagram presented in Fig. 8, we give the PAM order, the bit rate, the \( BER \) (from error counting), the \( Q \)-factor and the roll-off factor of the Tx and Rx pulse shaping and matched filters, respectively. Figures 8(a) and 8(b) show PAM-2 eyes at 30 and 60 Gb/s, respectively, Figs. 8(c) and 8(d) show PAM-4 eyes at 60 and 112 Gb/s, Figs. 8(e) and 8(f) show PAM-8 eyes also at 60 and 112 Gb/s, respectively, and finally Figs. 8(g) and 8(h) show PAM-16 eyes at 50 and 60 Gb/s, respectively. It is interesting to compare the eyes for 30 Gb/s PAM-2 and 60 Gb/s PAM-4, as they are generated at the same symbol rate. One can see that PAM-2 at 30 Gb/s provides a signal of very high quality, with a BER that is most probably already below the desired error rate of \( 10^{-15} \): the limited memory length of the ADC prevents accurate BER measurements from error counting at such high signal quality. Keeping the same signaling rate but doubling the number of generated levels from 2 to 4 still gives a rather good signal quality, but with an increase in the BER. It is also interesting to observe the eyes for PAM-16: even if eyes look very similar, an increase in bit rate from 50 Gb/s to 60 Gb/s doubles the BER from an error-free rate of \( 2.88 \times 10^{-15} \) to a BER falling above the BER FEC threshold. This shows the limited qualitative information provided by eye diagrams.

For a fixed baud rate, jumping from PAM-\( N_k \) to a higher PAM-\( N_i \) naturally decreases the \( Q \)-factor by
\[
5 \log_{10} \left( \frac{N_i^2 - 1}{N_k^2 - 1} \right) + \text{excess loss [dB]},
\]
where the mean signal power is assumed equal for both PAM formats, and where the standard deviation of the noise around each level is assumed constant notwithstanding the size of the input alphabet (\( N_k \) or \( N_i \)). From PAM-2 to PAM-4, \( N_k = 2 \) and \( N_i = 4 \), and the \( Q \)-factor drops by at least \( 5 \times \log_{10}(4^2-1)/(2^2-1)) = 3.5 \text{ dB} \). This decrease arises due to neighboring levels being closer to each other as the PAM order increases. In our case, \( Q \) diminished by 3.94 \text{ dB}, showing a small excess loss of signal quality of 0.44 dB by going from 2 to 4 level waveform generation at 30 Gbaud.

In the following, we test the accuracy of the \( BER_Q \) model of Eq. (5) presented in Section 5.1 and its equivalent \( Q \)-factor. The accuracy of the \( Q \)-factor, related to the \( BER_Q \) via Eq. (3), is tested by comparing the \( BER_Q \) of Eq. (5) with the \( BER \) obtained by error counting. Figure 9 shows good agreement between \( BER_Q \) and \( BER \). The figure also demonstrates the accuracy of the Gaussian model, by showing the normalized histogram of the received symbols (blue curves), overlaid with the aggregate probability density functions (PDF) of each level \( i \) of the multilevel signal assuming Gaussian noise statistics. The accumulated PDF shown by the red curves in Fig. 9 is \( \sum_{i=1}^{N} \mathcal{N}(\mu_i, \sigma_i^2) \), where \( \mathcal{N}(\mu, \sigma^2) \) is a Gaussian distribution of mean \( \mu \) and variance \( \sigma^2 \). Results are shown for PAM-8 at 112 Gb/s after propagation of 10 and 20 km.

![Histogram of received symbols and probability density function (PDF) of PAM-8 after (a) 10 km, giving a BER of \( 3.6 \times 10^{-3} \) and after (b) 20 km, giving a BER of \( 1.56 \times 10^{-2} \).](image)
At both distances, BER from error counting and BERQ from Eq. (5) show very good agreement. After 20 km, agreement is also high because the receiver is solely Gaussian noise limited. To show that the model fits any PAM order, Fig. 10 shows its accuracy for PAM-16 at 50 Gb/s, after 10 km.

![Fig. 10. Probability distribution function of PAM-16 at 50 Gb/s, after 10 km. Blue: received symbol distribution. Red: PDF computed from of mean $\mu$ and variance $\sigma^2$ of each 16 level.](image)

The match between both performance metrics, BER, and BERQ, is clear, demonstrating the strong accuracy of the BERQ model of Eq. (5) compared to the BER from error counting for multilevel PAM-N formats. Faithful matching of the histogram curves of PAM-8 and PAM-16 signals with the accumulated PDF curve further validates the model. The accuracy is high for received signals with low Q-factors (high BER) as shown in the previous Figs. 9 and 10 where BERs were greater than $10^{-3}$. Low Q’s are observed for two cases: 1) for low received signal power (after 20 km of propagation) where the receiver is Gaussian thermal noise limited and 2) for high bit rates, where the transmitter’s symbol rate is large, requiring heavy DAC compensation and consequently generating a waveform of smaller swing, of lesser signal-to-noise ratio, and moreover where the receiver integrates more inband noise from larger baud rate.

The match between the BER and the BERQ (Q-factor) diverges for two cases: for received signals of high Q (low BER), and for decreasing PAM orders. The former is explained by the following. High Q signals at the receiver are attainable when both a high Q signal is generated at the transmitter and when sufficient power is presented to the receiver. For signals of high enough power, the highest PAM level of a PAM-N format typically has a noise distribution that does not exactly match a Gaussian distribution. The cause is twofold. First, the two edge levels generated at the transmitter can be readily compressed by the raised cosine power transfer function of the Mach-Zehnder intensity modulator, while the central levels remain in the linear transfer region. Second, and most importantly, the p-i-n + TIA receiver can start saturating when receiving the highest PAM level of a higher power signal. Those cumulative effects at the transmitter and receiver will modify the additive white Gaussian noise (AWGN) assumption employed in the Q-factor computation. This effect builds up as PAM order decreases, from 16 to 8, 4 and 2, as the ratio of non–AWGN-only levels over AWGN levels increases, explaining the divergence of the two metrics as the PAM order decreases. The non–AWGN-only distribution of the highest level is naturally more clearly observed for signals of high quality, when the non-AWGN portion of the distribution is not buried under other AWGN noise sources, like thermal noise at low received signal power. To demonstrate this, we show in Fig. 11 the histogram of received symbols and the equivalent aggregate Gaussian PDF for (a) PAM-2 and (b) PAM-4 signals of high Q. In Fig. 11(a) we see how both inner tails of the histogram spread out slightly more than what is predicted by a Gaussian-only distribution. The excess spreading is moderately larger for the higher level. In Fig. 11(b), for high Q PAM-4, we clearly observe the non–AWGN-only
distribution of the higher level (level from alphabet label ‘3’) and its larger histogram inner tail, significantly increasing the BER from error counting from the predicted BERQ of a AWGN-only model. At such low BERs, a slightly larger tail does significantly impact the BER.

Figure 12 shows the system performance as we vary the bit rate from 25 Gb/s to 120 Gb/s for different PAM orders of 2, 4, 8 and 16, over different propagation distances of 0, 2, 10 and 20 km. For each bit rate of each PAM order, the roll-off factor of the pulse shaping filter is optimized to give the best system performance. We first present the transmission performance results using the Q-factor and the BER metrics in Figs. 12(b) and 12(c), respectively. Those results exhibit important characteristics for PAM order selection. First, for a fixed PAM order, it is observed that increasing the bit rate degrades the BER and the Q-factor. This is a normal behavior as a bit rate increase translates into increasing the signaling rate (baud rate), itself increasing the bandwidth of the signal, where the larger matched filter at the receiver will integrate more inband noise. Integrating more inband noise power while maintaining the average signal power degrades the BER and Q-factor. A more revealing observation drawn from both Figs. 12(b) and 12(c) is the following: looking at a fixed distance, we observe that as the bit rate increases, the highest Q and the lowest BER are attained using increasingly higher PAM-orders, where transitions happen at specific bit rates. From the Q curves in Fig. 12(b), bit rates below 53 Gb/s have the highest Q using PAM-2, bit rates between 53 Gb/s and 109 Gb/s have the highest Q using PAM-4, and bit rates above 109 Gb/s exhibit higher Q using PAM-8. The same trend is observed using the BER metric as shown in Fig. 12(c), at different transitional bit rates. For instance, the transitional bit rate at which PAM-8 gives a smaller BER than PAM-4 as the bit rate is increased is 104 Gb/s compared to 109 Gb/s when the Q-factor metric is employed, showing a good match of the transitional bitrate using two very different performance metrics. At lower bit rates, the transition where PAM-2 gives a smaller BER than PAM-4 as the bit rate is reduced occurs at a bit rate that exhibits a BER that cannot be accurately measured from the bit error counting method due to the limited memory of the receiver. This PAM-2 to –4 transitional bit rate would also be smaller using the BER technique than the one obtained using the Q-factor metric. Notwithstanding this off-line measurement limitation, the BER slopes of PAM-2 and PAM-4 in Fig. 12(c) clearly indicate that as the bit rate is lowered a transition point will be reached at which PAM-2 will outperform PAM-4. This greater deviation of the transitional bit rate between PAM-2 and –4 for the two metrics can be explained from Fig. 12(d), where we investigate the relation between each y-axis variable in Figs. 12(b) and 12(c), i.e. where we plot the error counting BER as a function of the Q-factor.

The waterfall curves in Fig. 12(d) corroborate the histograms, PDFs and remarks drawn from Figs. 9–11. Figure 12(d) shows that the two metrics of Q-factor and BER match very well for signals of low Q and signals of high PAM order, and start diverging as the signal...
quality is increased and as PAM order is decreased. As the transitional bit rate between PAM-2 and 4 occurs at both high $Q$ and low PAM order, the two metrics exhibit a different transitional value.

For completion, SNR versus bit rate curves are also added to Fig. 12 in (a) as a third performance metric in order to compare it with the two main metrics of $BER$ and $Q$. An important observation is that for any desired bit rate, the SNR is always greater at larger PAM orders, when monitored after a fixed distance. This contradicts the observations of the $BER$ and $Q$ metrics and shows the inaptitude of the SNR metric for multilevel PAM signaling to act as a representative qualitative figure of merit, like the $Q$-factor proves to be. This being said, we can explain this SNR behavior by the lower required baud rate at any desired bit rate as the PAM order increases. Smaller baud rate translates to a further accurate waveform generation with more samples per symbols, and a matched filter of smaller bandwidth. A second observation from the SNR curves shows that as the bit rate increases, the SNR of PAM-4 decreases faster than that of PAM-8. The rational is closely related to the previous observation. For an increase in bit rate of 30 Gb/s, PAM-4 requires an increase in signaling rate of 15 Gsymbols/s, where PAM-8 only requires an increase by 10 Gsymbols/s. Rapid increase of signaling rate translates into rapid decrease of the signal to noise ratio of the transmitter waveform generated. This trend rapidly accumulates as the bit rate is increased from 50 Gb/s to 120 Gb/s.

![Fig. 12. (a) SNR, (b) $Q$-factor [dB] and (c) BER for PAM orders 2, 4, 8 and 16, after propagation distances of 0, 2, 10 and 20 km, for varying bitrates. Figure (d) shows the BER from error counting against $Q$($BER_{th}$). Black solid curve in (d) represents the theoretical $BER_{th}(Q)$ relation of Eq. (3). Hashed line in (c) and (d) is $BER = 3.8 \times 10^{-3}$ threshold.

In light of our observations in Fig. 12, it is interesting to look back at the eye diagrams of Fig. 8. First, the eyes of Fig. 8 are represented by bold stars in Figs. 12(b) and 12(c). The optimum PAM order selection as a function of bit rate observed in the performance curves of Fig. 12 is in fact visually well depicted in the eyes of Fig. 8. For low bit rates in the region of
30 Gb/s, it is clear that the PAM-2 format offers better performance than PAM-4: the eye is wide open, levels are largely separated and the raw BER is potentially already below $10^{-15}$. However, for a target BER in the region of 60 Gb/s, Figs. (8(b), 8(c), and 8(e)) show that PAM-4 provides the cleanest signal at the receiver, with the most well defined, distinctly separated levels. Indeed, both the error-counting BER and the $Q$-factor are better when using PAM-4 at 60 Gb/s compared to when using PAM-2 or PAM-8. On the other side, if the desired bit rate is in the 112 Gb/s region, Figs. (8(d) and 8(f)) show that the BER obtained using PAM-4 can be halved by reducing the signaling rate by a third of its value and using 8 levels instead of 4. The $Q$-factor also improves at 112 Gb/s with PAM-8. Trends in Fig. 12(c) show that the BER improvement using PAM-8 over PAM-4 further increases with increasing bit rates starting at around 104 Gb/s.

In the following Fig. 13 we present the system performance as we vary the signal power presented to the $p$-$i$-$n$ + TIA receiver, for different PAM orders. To collect this data the experimental test bed of Fig. 6 is slightly modified: the system is in back-to-back (0 km) and the receiver is preceded by a variable optical attenuator (VOA), followed by 99/1 coupler with characterized port ratio and a power meter on the 1% port. For PAM-4 and PAM-8, the bit rate is set to 107 Gb/s and for PAM-2, 60 Gb/s. Both the $BER$ and $Q$-factor metrics are presented to monitor the performance, respectively in Figs. 13(a) and 13(b). The figure shows that the optimum received signal power is dependent on the PAM order. For PAM-8, the received signal power yielding the lowest BER is around $–4.3$ dBm. For PAM-4 and PAM-2 the maximum received power was lower than the optimum operational point, allowing room for improved performance. The coupling loss of the grating couplers of the integrated SiP modulator provides the bulk fiber to fiber loss of the modulator. Better light coupling into the optical chip would significantly improve the available power to present to the $p$-$i$-$n$ + TIA receiver and allow operation at optimum Rx power for both lower PAM orders. Nonetheless, looking at the trends around the highest received signal power of approximately $–3$ dBm for PAM-4 and PAM-2, we observe that PAM-4 is optimum just above $–3$ dBm, and PAM-2 requires even higher power before it reaches optimum operation. This optimum power reduction with increasing PAM order shows the receiver linearity requirement as we increase the number of signaling levels. A received optical power that is too large will compress the higher levels of PAM-$N$: a compression that is more detrimental as $N$ increases.

Figure 13(a) shows both the $BER$ obtained from error counting, as well as the equivalent $BER_Q$ from the computed $Q$-factor of Fig. 13(b). For low signal power, where the receiver operates fully linearly and is limited by Gaussian thermal noise, both BERs match perfectly, meaning that in this regime, the AWGN assumption of Eq. (5) is met. From the initial slope of Fig. 13(b), an increase of 1 dB in received power is linearly translated to a $Q$ increase of 0.76 dB. As the power is further increased, solely by changing the VOA attenuation, the receiver starts behaving non-linearly and the dB-for-dB relation vanishes. The higher PAM levels start compressing down and the Gaussian noise statistics are modified. The $Q$-factor from the computed mean $\mu_i$ and variance $\sigma_i^2$ of all levels give an equivalent $BER_Q$ that is much smaller than the real BER from error-counting. This is because the AWGN assumption on all PAM levels does not hold well at increasing powers, where the highest levels are compressed by the $p$-$i$-$n$ + TIA and exhibit non-AWGN noise distribution, as mentioned in the discussion of Fig. 11. Computing the first two statistical moments of all levels and applying the Gaussian model to each level fails to provide an accurate BER estimate. A sum of two different noise processes, with one being power dependent, would better model the noise statistics at the receiver and provide a better BER estimation. Figures 13(a) and 13(b) show a vertical line at dBm, representing the received optical signal after 10 km of SMF-28e+. 
In the following Fig. 14, we present the signal quality, by showing the BER and the $Q$-factor, for PAM orders of 2, 4 and 8 at different baud rates, versus the roll-off factor of the root raised cosine pulse shaping filter. Results are all captured after 10 km of propagation. BERs are obtained from error counting, each trace containing $10 \times 10^6$ bit of information. Consequently, the BER’s log-scale is deliberately limited to $10^{-6}$ for the lower bound. For all PAM orders, we observe that both $Q$ and BER have an optimum roll-off factor that decreases with increasing baud rates. It is important to reiterate here that the DAC’s sampling rate is always fixed at 70 GSamples/s. This first limits the range of roll-off factors as a function of the baud rate (section 4.1) and second, changes the number of SPS as the baud rate varies.

Smaller SPS translates to closer spectral images and less accurate analog waveform generation, to the extent that at 1 SPS, the DAC acts as a multilevel PPG with no pulse shaping capabilities. Moreover, at an equal roll-off factor, increasing the baud rate by 5 Gbaud means increasing the maximum spectral content of the signal that needs to be generated by the DAC by $\frac{5}{2}(1 + \alpha)$ GHz. From this, the experimental decrease of $\alpha$ becomes intuitive. Three co-acting effects worsen the quality of the generated signal with larger spectral content. The first effect is the reduced output swing. The analog response of all the RF components at the transmitter side rapidly decreases in amplitude with frequency. The response can still be equalized after the addition of the extra bandwidth to generate the desired pulse shape, but this is done at the expense of reduced swing out of the DAC. As all the PAM-$N$ levels have to fit in this reduced electrical swing, the electrical signal to noise ratio of the generated RF waveform is inevitably reduced. This will get translated to the optical waveform generation, and observed at the receiver. The second effect worsening the quality of the generated signal after a baud rate increase is the effective number of bits (ENOB) of the DAC, which decreases with frequency. Consequently, the signal quality within the additional bandwidth, required for the increased baud rate, will inherently be of worse quality. Thirdly, increasing the baud rate widens the full width at half max (FWHM) of the receiver matched filter, which translates into integrating more noise power with the signal as the matched filter is applied. Consequently, even if equal signal power was assumed after a baud rate increment, the SNR at the receiver would intrinsically worsen. The sum of those three effects explains 1) the worse performance at a fix roll-off factor and PAM order for increasing baud rates, 2) the reduced optimum roll-off factor at a fixed PAM order for increasing baud rates, and 3) the reduced performance for increasing PAM order at a fixed baud rate and fixed roll-off factor.
Fig. 14. System performance assessment for varying PAM orders, at different baud rates, for varying roll-off factors. (a) $Q$ for PAM-2, (b) BER for PAM-2, (c) $Q$ for PAM-4, (d) BER for PAM-4, (e) $Q$ for PAM-8, (f) BER for PAM-8. Hashed black line is BER of $3.8 \times 10^{-3}$ or the $Q$ equivalent.

Finally, for low baud rates of 25, 30 and 35 Gbaud where 2 SPS or more are employed to accurately generate the desired waveform, one can observe that the optimal roll-off factor decreases for increasing PAM order. It is noteworthy to reiterate that all data of different bit rates and PAM orders presented in Figs. 12 and 13 employ the optimum roll-off factor presented in Fig. 14.
The following Fig. 15 shows (a) the BER and (b) the Q-factor for PAM-4 and PAM-8 when running at 112 Gb/s, over varying distances of 0, 2, 10 and 20 km. We can observe that PAM-8 is able to yield a BER that is below $3.8 \times 10^{-3}$ after 10 km. At a greater distance of 20 km, the received signal power is too low and the BER increases with decreasing signal power. The received power after 20 km was −9 dBm. For 0 and 2 km, the received signal power had to be slightly attenuated in order to operate at the optimum powers, as shown in Fig. 13. PAM-4, however, can never deliver the 112 Gb/s throughput while providing a BER below the $3.8 \times 10^{-3}$ FEC threshold at any distance. This format also shoots to smaller Q’s and higher BERs as the received signal power decreases from 10 km to 20 km of SMF-28. Figure 15 presents both metric performance, however, only the BER from error counting defines whether or not a modulation format performs below the BER FEC threshold after a specific distance, consequently confirming or denying its error free operation.

Fig. 15. (a) Q-factor and (b) BER for PAM-4 and PAM-8, at 112 Gb/s, at varying distances.

6. Conclusion

The performance of a SiP intensity modulator at 1310 nm for short reach transmission using PAM modulation format is experimentally studied. Transmission of 112 Gb/s over 10 km of SMF fiber below the pre-FEC BER of $3.8 \times 10^{-3}$ using PAM-8 is experimentally validated. A large parameter space was swept for both the transmitter and receiver, namely the PAM order, baud rate and bit rate, pulse shaping roll-off factor, propagation distance and received signal power. We show that to optimize the system performance and minimize the BER, there exists an optimum PAM order at a given desired bit rate, with PAM of an increasingly higher order being preferable as the bit rate is increased. We also present a model for the Q-factor for multilevel PAM signaling. We experimentally demonstrate the accuracy of the model by first comparing its equivalent BER to the BER from error counting, and second by comparing the aggregate PDF of the model with the histogram of the receiver data. The two approaches show an accurate match in the case of Gaussian noise limited transmission.