70 Gbps 4-PAM and 56 Gbps 8-PAM Using an 850 nm VCSEL

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Krzysztof Szczerba, Petter Westbergh, Magnus Karlsson, Peter A. Andrekson, Anders Larsson.

***Abstract*—We present 56 Gbps unequalized 8-PAM real-time transmission over 50 m of MMF and 70 Gbps 4-PAM operation with offline equalization. FEC would be necessary in both schemes. The bit-rates are given taking into account the FEC overhead needed to reach a BER below 10^{-12}. The experiments were performed with an 850 nm VCSEL with 20 GHz bandwidth and a 22 GHz photoreceiver. We show that 8-PAM modulation requires channel bandwidth comparable with the symbol rate to avoid excessive ISI penalty and that equalization of 4-PAM signal can significantly improve achievable bit-rates.**

***Index Terms*—fiber-optical communications, IM/DD, 8-PAM, 4-PAM, OOK, VCSEL, MMF, short-range, 850 nm, data communications, interconnects, intersymbol interference, equalization.**

I. Introduction

NOWADAYS short-range optical communications relies on low-cost vertical cavity surface emitting lasers (VCSELs), multimode fiber (MMF) and direct detection in the receiver. On-off keying (OOK) is the only modulation used in commercial links today. The main reasons for this are cost efficiency and low power consumption of such simple systems, both of which are very important in typical datacom applications. The link throughput was historically increased by both faster lasers and parallelization. Today commercial optical interconnects with single-lane data rate of 25 Gbps are becoming available [1]. In typical applications multiple lanes are used in parallel to provide higher bit-rate per transceiver, e.g. as done in Ethernet or Infiniband. These approaches have their limits and recently alternative techniques, such as multilevel pulse amplitude modulation (PAM) and equalization have been explored. Binary 64 Gbps transmission over 57 m of MMF was demonstrated in [2] using a high-speed VCSEL and custom equalizer circuits. 60 Gbps operation of 4-PAM without equalization and with bit error rates (BER) measured in real-time down to 10^{-12} was demonstrated in [3]. In [4] pre-distortion of 4-PAM signal to compensate for a VCSEL response was experimentally demonstrated. Off-line equalization and FEC with 4-PAM at 100 Gbps was demonstrated using a directly modulated 1550 nm VCSEL in [5]. The 100 Gbps operation was achieved by use of dual polarization in a single mode fiber, single-polarization data rate was 50 Gbps.

Modulation formats other than PAM were also investigated for short-range links with intensity modulation and direct detection (IM/DD). Simple subcarrier approaches were presented e.g. in [6]–[9]. To improve the sensitivity of the subcarrier modulation schemes three-dimensional subcarrier modulation formats optimized for improved sensitivity were proposed [10]–[12] and demonstrated experimentally [13]–[15]. Discrete multitone (DMT) modulation with multiple orthogonal subcarriers, allowing for adaptive bit loading was investigated for reach extension in VCSEL and MMF based optical links and [16]–[18]. Carrier-less amplitude and phase (CAP) modulation for datacom links was investigated in [19], [20] and extended into a similar concept called orthogonal multipulse modulation (OMM) [21].

In this paper we present results of high-speed data transmission experiments using 4-PAM combined with offline equalization and 8-PAM in real-time without equalization. The experiments were performed using link based on an in-house developed 850 nm VCSEL, an OM4+ MMF and direct detection with a 22 GHz photoreceiver. The two modulation formats were tested at symbol rates of 10, 20 and 25 Gbaud for 8-PAM and 40 Gbaud for 4-PAM. Use of FEC was assumed since it is gaining support in short-range optical communications, e.g. the most recent version of the Infiniband standard [22] includes FEC. The results presented here in detail were described briefly in [23].

II. Theoretical performance

A. Sensitivity

Derivations of the generalized exact expression for the BER of an M-PAM modulation in an IM/DD system can be found in [24], [25]. Here, we are providing a simplified, closed form expression for the BER from [24]. The simplified BER expression is

$$BER \approx \frac{M-1}{M}\frac{d_{\text{avg}}}{\log_2 M}\,\text{erfc}\left(\frac{I_{\text{avg}}}{(M-1)\sigma\sqrt{2}}\right) \qquad (1)$$

where M is the number of levels, I_{avg} is the average photocurrent, σ is the noise variance and d_{avg} is the average Hamming distance between the labels of adjacent symbols [24]. When Gray labeling is used, $d_{avg} = 1$. If natural labeling is used it is [24]

$$d_{avg} = 2 - \frac{\log_2(M)}{M-1}. \qquad (2)$$

The BER expression (1) is valid under assumption of additive white Gaussian noise (AWGN) being the dominating noise type. This is the case when the system is dominated by the thermal noise and without patterning effects. Example theoretical BER values calculated for 4- and 8-PAM are illustrated in Fig. 1. Both natural and Gray labeling are presented. For

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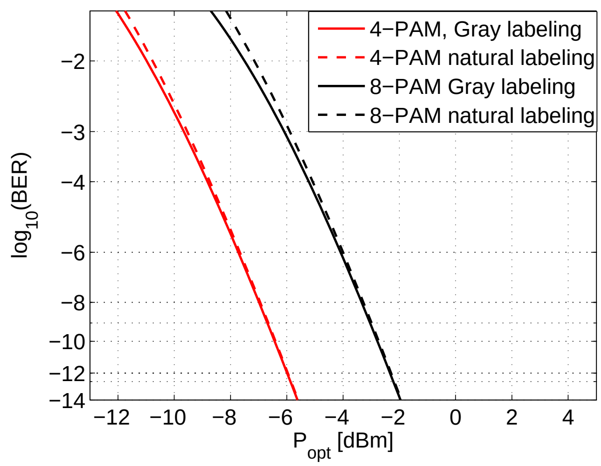


Fig. 1. Theoretical BER for 4- and 8-PAM with a 22 GHz noise bandwidth.

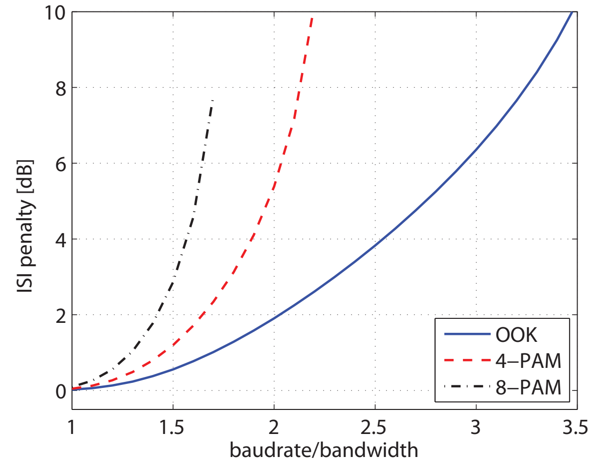


Fig. 2. ISI penalty for OOK, 4-PAM and 8-PAM as a function of symbol rate to bandwidth ratio.

simplicity it is assumed that the noise bandwidth is 22 GHz, the same as the bandwidth of the photoreceiver used in the experimental setup. The noise figure of the amplifier in the photoreceiver is 5 dB and the photoreceiver output is coupled into a 50 Ohm load at room temperature. The RIN of the laser was assumed to be -145 dB/Hz, which is typical for our in-house developed VCSELs. The calculated BER plotted in Fig. 1 shows that at high BER a 0.5 dB difference in sensitivity is expected between the Gray and natural labeling. This difference decreases at low BER to negligible values. In practice, use of natural labeling can simplify experimental setups for M-PAM experiments.

B. Intersymbol Interference

The power penalty due to the intersymbol interference (ISI) is well understood for OOK systems. The basic ISI penalty calculation methods, have been outlined in [26]. The ISI power penalty, generalized to any PAM format and expressed in dB is

$$P_{ISI} = 10\log_{10}\left(\frac{1}{1-E_m}\right), \tag{3}$$

where E_m is the worst case eye closure. In general, for an M-PAM format the eye closure will be

$$E_{m,MPAM} = \frac{M}{2}1.425\exp\left(-1.28\left(\frac{T}{T_C}\right)^2\right). \tag{4}$$

The bit period is denoted as T and the channel 10% – 90% rise-time is denoted as T_C. This ISI penalty calculation method, which is valid under assumption of a Gaussian channel response and rectangular input pulse, was given in [27]. It is used in the IEEE 802.3 link budget spreadsheet [28]. Methods for calculation of the 10% – 90% rise-time for a given system are described in [27].

The expected theoretical ISI penalty for OOK, 4-PAM and 8-PAM are presented in Fig. 2. The ISI penalties are plotted as a function of the ratio of the symbol rate to the channel 3 dB bandwidth. It is clear that the ISI penalties increase faster with increasing number of levels. In fact, OOK can be used with just 2 dB ISI penalty with symbol-rate (and thus bit-rate) twice as high as the channel bandwidth. At the same symbol-rate to bandwidth ratio, 4-PAM is expected to experience 6 dB penalty. In case of 8-PAM the channel bandwidth should be almost equal to the symbol-rate to avoid significant ISI penalties.

We should emphasize that the above theoretical ISI penalties account only for the vertical eye opening, without taking into account the effects of ISI on the horizontal eye opening which will further degrade the sensitivity. The assumption of a Gaussian frequency response of the channel is also not necessarily accurate in all cases. In fact, the frequency response of the VCSELs itself may vary depending on the bias conditions. The Gaussian frequency response is a useful approximation in this situation.

C. Theoretical performance of Reed-Solomon FEC

The Reed-Solomon (RS) error correcting codes were introduced in 1960 [29]. They are one of the most popular classes of FEC, used in many communication technologies. Their hardware implementations have been studied and optimized over many years. The performance of RS FEC was investigated e.g. in [30].

The RS code with n-symbol-long codewords, each symbol of length q bits, can correct $t = (n-k)/2$ symbol errors, where k is the number of payload data symbols. The code rate is k/n. The error rate at the output of the decoder can be calculated as a function of the bit-error rate at the input of the decoder. Assuming that the error probability at the input BER_{in} is independent over all the bits in the FEC codeword, the incoming symbol error rate is

$$SER_{\mathrm{in}} = 1-(1-BER_{\mathrm{in}})^q. \tag{5}$$

Under assumptions that the decoder corrects all errors up to t symbols and detects all errors above t symbols, the symbol error rate at the output of the decoder is [31, Ch. 8]

$$SER_{\mathrm{out}} = \frac{1}{n}\sum_{i=t+1}^{n} i\binom{n}{i} SER_{\mathrm{in}}^{i}(1-SER_{\mathrm{in}})^{n-i}. \tag{6}$$

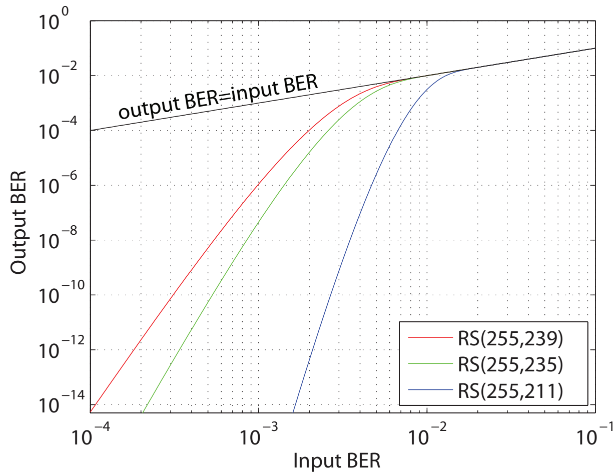


Fig. 3. Theoretical approximate performance of Reed Solomon FEC with eight-bit symbols, 255 symbols long codewords and 239, 235, and 211 data symbols.

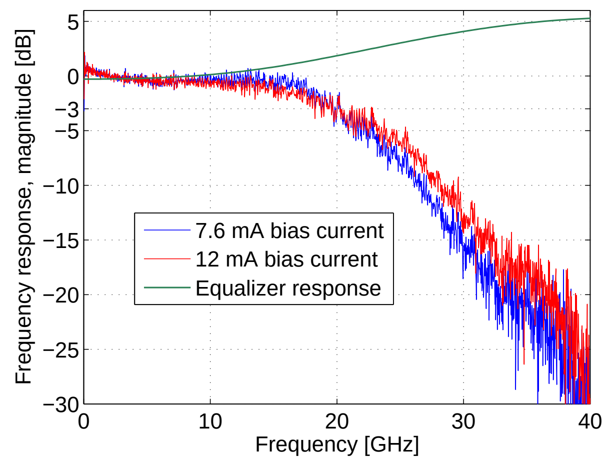


Fig. 4. Magnitude of the frequency response of the VCSEL used in the experiment at 7.6 mA and 12 mA bias current.

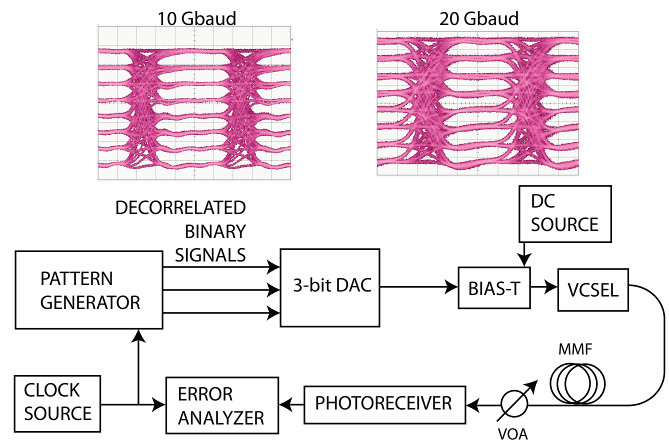


Fig. 5. Experimental setup used for 8-PAM experiments.

The output bit-error rate can can be calculated from the output symbol error rate by inverting the step in (5),

$$BER_{\text{out}} \approx 1 - (1 - SER_{\text{out}})^{1/q}. \tag{7}$$

The last step is an approximation and an exact approach was given in [30]. Because of the high BER in case of the 40 Gbaud 4-PAM reported in this work RS codes with better error correcting capability than RS(255,239) were taken into consideration: RS(255,235) and RS(255,211). The approximate theoretical performance of these codes is shown in Fig. 3. The codes considered here use 8-bit long symbol, therefore the codeword length is 2040 bits. This is comparable to e.g. what is used in the current Infiniband standard, where a shortened cyclic code of length 2112 bits and with 2080 payload bits [22]. A downside of FEC is the introduced latency, mainly in the decoder. In practice, latency requirements have to be evaluated and taken into account in system design.

III. Experimental setup

Two different test setup configurations were used to generate the 8- and 4-PAM signals. The setups used the same laser, photoreceiver and fiber in both cases. The laser was an in-house developed VCSEL, operating at the wavelength of 850 nm, with 8 µm oxide aperture diameter and around 20 GHz modulation bandwidth. The laser was similar to the one reported in [32], where 47 Gbps OOK transmission was demonstrated. The VCSEL output was launched into an OM4+ MMF using a lens package. The coupling efficiency was around 50% and the launch power into the MMF was around 5 dBm. The photoreceiver was a commercially available model with 22 GHz bandwidth and an integrated low-gain linear inverting transimpedance amplifier (TIA). A linear TIA is a requirement for multilevel modulation.

The test setups differed by how the electrical signal was generated and how the BER was evaluated. In the 8-PAM configuration, shown in Fig. 5, a high speed 3-bit digital to analog converter (DAC) was used. The DAC was of the same type as used previously [33], the highest supported symbol rate was 32 Gbaud. It was driven by three decorrelated binary streams, each using a PRBS pattern with length of $2^9 - 1$ bits. Eye diagrams of the electrical 8-PAM signal at 10 and 20 Gbaud are shown as insets to Fig. 5. The signal, with an amplitude of 842 mV, was fed through a bias-T to the VCSEL, biased at 7.6 mA. The frequency response of the VCSEL is shown in Fig. 4. The BER was measured in real-time using an ordinary error analyzer designed for OOK, similarly to our previous work [33] on 8-PAM. The overall BER was derived from the error rate (ER) on the middle threshold. Under assumption that the ERs at all decision threshold are the same, the overall BER is equal to $(11/3)$ER [33].

In the 4-PAM test setup (Fig. 6) the electrical driving signal was generated by coupling together two decorrelated binary signals, each being a PRBS pattern of length $2^7 - 1$. The test pattern was short, but it was deemed acceptable, since in practical systems run-length coding (like the 64b/66b [22]) is used to aid DC balancing and clock and data recovery. Moreover, the probability of the longest run length is typically low. The binary signals were amplified and

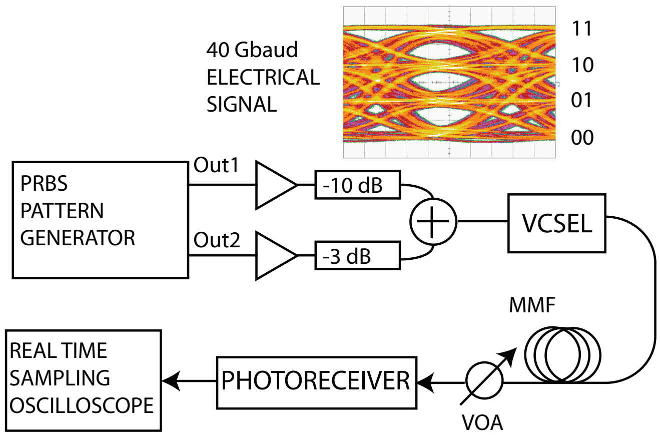


Fig. 6. Experimental setup used for 4-PAM experiments. The amplitude of the binary outputs of the pattern generator was adjustable and used to optimize the electrical 4-PAM signal quality.

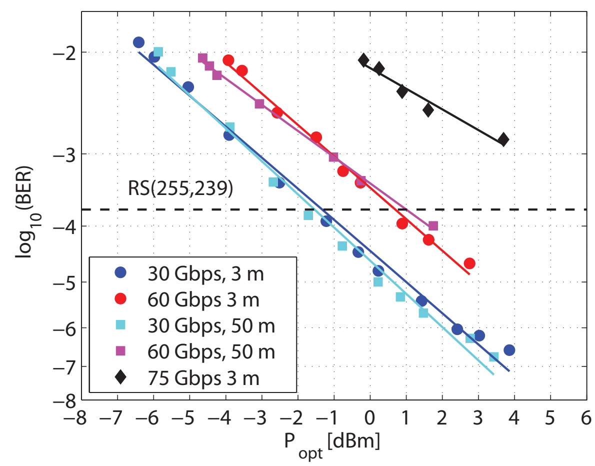
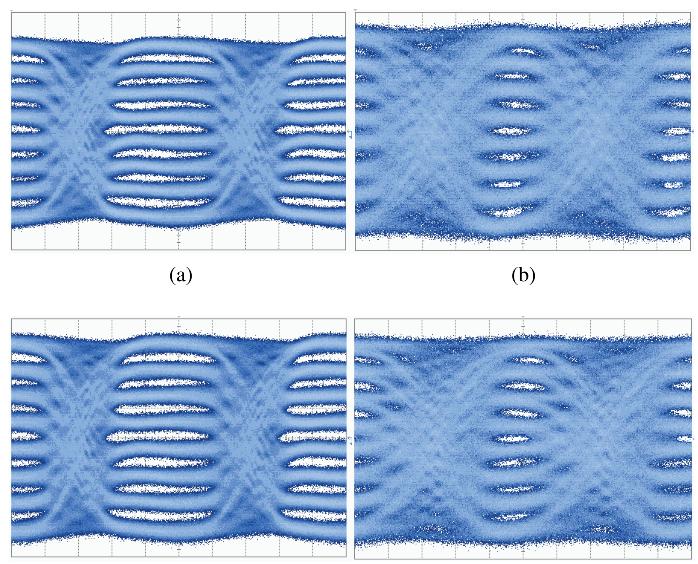


Fig. 7. Experimental BER results achieved with 8-PAM with the FEC threshold for reference.



(a) (b)

(c) (d)

Fig. 8. ye diagrams of 8-PAM at after 3m of fiber at 10 Gbaud (a), 20 Gbaud (b) and after 50 m of fiber at 10 Gbaud (c) and 20 Gbaud (d).

adjusted in amplitude so that one of the signals had half of the amplitude of the other and combined in a high bandwidth resistive 3 dB coupler. An eye diagram of the electrical signal is shown in the inset to Fig. 6. The VCSEL was biased at 12 mA and the electrical signal amplitude was around 1.2 V. The reason for the higher bias current was to accommodate the increased drive signal amplitude. The frequency response of the VCSEL at this bias current is also shown in Fig. 4. A 3 m OM4+ patchcord was used for transmission. The signal from the photoreceiver was acquired using a real-time sampling oscilloscope with 33 GHz bandwidth and 100 Gsps sampling rate and processed offline. The offline processing included a static equalizer designed to emphasize the signal frequencies in the 20–40 GHz range to compensate for the VCSEL frequency response roll-off. The frequency response of the filter is included in the Fig. 4. The equalizer response was a result of a trade-off between reduction of the ISI and amplification of noise in the high frequency part of the response. The equalizer was a linear phase finite impulse response (FIR) filter with length corresponding to 5 symbols for signal at 40 Gbaud. The purpose of the equalizer was to emulate a static analog equalizer which could be placed in the receiver, but which we did not have capacity to implement in hardware, therefore the equalizer was vastly oversampled. In a system with a digital equalizer 5 symbol spaced taps would be sufficient to reach a similar level of performance.

Because of the differences between the test setups, a direct comparison of the 8-PAM and 4-PAM results is not possible. The purpose of the experiments was rather to expand the envelope of what can be achieved using multilevel modulation in 850 nm based links.

IV. Experimental results

A. Results with 8-PAM

The BER results from the 8-PAM experiments are shown in Fig. 7. The tested bit rates were 30, 60 and 75 Gbps over a 3 m OM4-type MMF patchcord and 30 and 60 Gbps over 50 m of OM4-type MMF. There is about 2.5 dB sensitivity penalty between the 30 Gbps and 60 Gbps, which is mainly due to increased baudrate and partly due to the ISI. The effects of ISI for 30 Gbps and 60 Gbps cases can be seen in the eye diagrams shown in Fig. 8. As the bit rate is increased, the eyes start to close, both horizontally and vertically, which is reflected in the BER results. The ISI penalty in case of 8-PAM increases quickly with the ratio of the symbol rate to the channel bandwidth. The channel had around 20 GHz bandwidth, limited equally by the photoreceiver and the laser, consequently 60 Gbps (corresponding to 20 Gbaud) was the highest bit-rate for which the BER goes below the FEC threshold.

Because the lowest measured BER was well above 10^{-12}, use of FEC would be necessary. For example, a popular RS(255,239) would suffice. For this code the input BER threshold to reach 10^{-12} at the output is 1.8×10^{-4}. The code

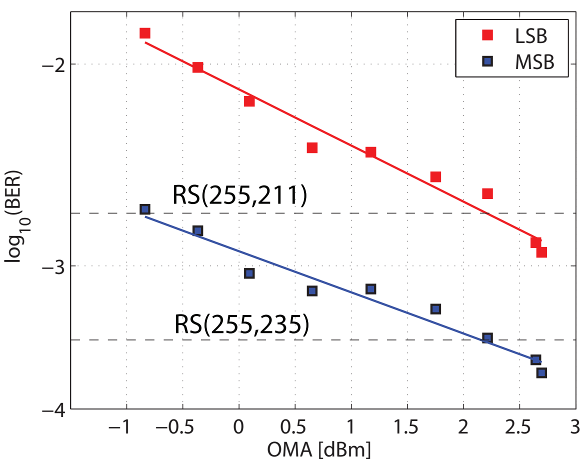


Fig. 9. Experimental BER versus optical modulation amplitude (OMA) for 40 Gbaud 4-PAM experiments, separated between the MSB and the LSB with respective FEC thresholds.

has 7% overhead and therefore the effective transmission bit rates with this code, for the 30 Gbps and 60 Gbps would be, respectively, 28.1 Gbps and 56.2 Gbps. In the 75 Gbps case the measured BER does not fall below the FEC threshold of the RS(255,239).

B. Results with 4-PAM

The BER results obtained with 4-PAM at 80 Gbps are shown in Fig. 9. An eye diagram of the captured signal before equalization is shown in Fig. 10a. The symbol rate of 40 Gbaud was twice the channel approximate -3 dB bandwidth. From Fig. 2 the ISI penalty should be 6 dB, but at this point the eyes are almost closed and with added noise it is impossible to recover the transmitted symbols without equalization. The symbol levels cannot be distinguished in the signal before equalization. An eye diagram of the signal after off-line equalization is shown in Fig. 10b. The signal quality is improved enough to distinguish the signal levels. The BER results are presented separately for the most-significant bit (MSB) and the least-significant bit (LSB). Because of the bit-to-symbol labeling used, the error probability on the MSB is lower than on the LSB. The labeling is illustrated in the inset to Fig. 6. The value of the MSB (which is the leftmost bit in Fig. 6) changes only between the middle two levels, the value of the LSB changes between every adjacent pair of levels and therefore every symbol error results in an LSB error. Because of the different BER values on the LSB and MSB, they should be coded with different FECs, similarly to what was done in [5]. The MSB and LSB can be coded for example with RS (255,235) and RS (255,211). The RS (255,235) code has 8% overhead and requires input BER of less than 3.3×10^{-4} to reach output BER below 10^{-12}. The RS (255,211) code has 17% overhead and the input BER threshold is 2×10^{-3}. With the FEC overhead the effective data rate would be 70 Gbps.

The factor which limited the BER performance the most was the receiver performance. The photoreceiver had a bandwidth of only 22 GHz and conversion gain of only 80 V/W. For comparison, the receiver used in 4-PAM experiments reported in [34] had a conversion gain of 400 V/W. It had 12 GHz bandwidth but it allowed for 30 Gbps 4-PAM transmission over 200 m of OM4 MMF without use of any equalization or FEC. This shows that receiver performance is crucial, and it is unfortunately a bottleneck. An improved receiver, would result in lower pre-FEC BER, this in turn would allow for FEC with lower overhead or with the same FEC for improved link margins. Finally, the equalizer could be also adjusted to provide more gain in the 30 to 40 GHz range and further reduce the ISI.

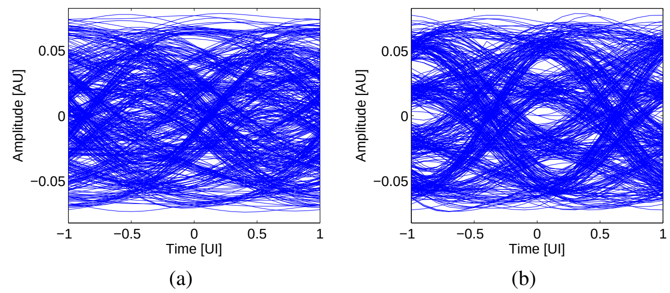


Fig. 10. Recovered 4-PAM eye diagrams before (a) and after (b) equalization.

V. Conclusion

High-speed operation of 8-PAM and 4-PAM was demonstrated in a short-link built with an 850 nm VCSEL. It is clear that an increased number of levels reduces the robustness to ISI and in case of 8-PAM, without equalization, the required system bandwidth is equal to the symbol rate. Equalization of 8-PAM was, however, not studied but it would likely enable increased data rates. In case of equalized 4-PAM it was possible to increase the symbol rate to twice the system bandwidth. This shows that a combination of multilevel modulation, equalization and FEC may be a promising approach in the future. If a photoreceiver with larger bandwidth and higher gain integrated amplifier was available, the data rates could be increased.

The RS FEC is a popular choice in many applications. However, in short-range optical communications energy consumption, latency and complexity are important considerations. The RS codes are good at correcting burst errors, because they correct entire symbols which are a few bits long. This is not necessary in typical short-range interconnects. Potentially, binary codes with simpler decoders, and therefore lower power consumption could be used. Latency requirements preclude use of very long codes.

In general, the performance of the FEC, equalizers and choice of modulation formats has to be optimized together to minimize the energy consumption per bit of the entire system.

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