Experimental Demonstration of Flexible Hybrid Modulation Formats for Future Adaptive Optical Transmission Systems

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Abstract—The exponentially growing demand for high-speed data transmission is one of the main reasons for the ongoing evolution from fixed and static to link-adaptive systems in fiber-optic communications. The need to maximize the flexibility as well as the capacity of the system, while keeping the optical network infrastructure unchanged, leads to modifications of the system terminals, resulting in so-called flexible transceivers. A highly desired property of such a transceiver is a fine data rate granularity, which allows to match the transmission capacity to the link budget. In this paper, we experimentally analyze flexible quadrature-division hybrid modulation formats, namely Flex-PAM, as a promising candidate for flexible transceivers. This method works at every integer number of bits per symbol, and, compared with other candidates proposed in the literature, requires a very simple digital signal processing scheme. We focus in particular on the effects of the effective number of bits in state-of-the-art digital/analog and analog/digital converters and fiber nonlinearities on the performance of Flex-PAM transmission. We analyze the capabilities of flexible modulation in a $5 \times 32$ Gbd wavelength division multiplexing coherent transmission on a dense 37.5 GHz grid and support the results by thorough numerical simulations. We successfully demonstrate the capability of Flex-PAM modulation to handle a wide range of bit-rates ($5 \times 128$ to $5 \times 320$ Gbit/s), a maximum spectral efficiency of 6.8 (bits/s)/Hz, and propose a simple and adaptive digital signal processing scheme to compensate for the major linear and nonlinear impairments of the system.

Index Terms—optical fiber communication, transceivers, digital modulation, digital signal processing, wavelength division multiplexing, coherent transmission

I. INTRODUCTION

Optical networks are undergoing substantial changes, determined by the exponential growth in data traffic and the demand for fast and reliable exchange of information. One aspect is the evolution of fiber capacity, which has been boosted by several orders of magnitude over the last decade, and is expected to grow by 40–60% per year in the future [1, 2]. In the past, a simple on-off keying modulation and a fixed channel-spacing was adequate for bit rates up to 10 Gbit/s and a relatively fixed network. Later-on, higher modulation formats and polarization division multiplexing (PDM) were used to enable bit rates of 100 Gbit/s and beyond. Compared with a fixed-length point-to-point link, an optical network results in a variety of transmission paths, each with its unique link-budget. While changes to the network itself result in high costs, it is advantageous to modify the order of the modulation format in order to match the data rate to the achievable capacity determined by the link budget. The usage of standard modulation formats such as square PDM-quadrature amplitude modulation (QAM) enables only a rough granularity, resulting for instance in four bits per symbol for PDM-4QAM, eight for PDM-16QAM and 12 for PDM-64QAM [3, 4]. The extension of square QAM formats with non-square formats such as PDM-8QAM (6 bits per symbol) and PDM-32QAM (10 bits per symbol) reduces the granularity gap to two, but requires already a more complex digital signal processing (DSP). Now assuming a network path that allows transmission of 11 bits per symbol as a result of its link-budget or optical signal-to-noise ratio (OSNR). Working with the aforementioned granularity of bits per symbol will force a suboptimal transmission, i.e. an unnecessary loss of three bits per symbol for quadratic PDM-QAM and one bit per symbol when including non-quadratic QAM modulation. Therefore, flexible transceivers enabling a finer granularity regarding possible amounts of bits per symbol are of great interest, as they offer a more efficient use of the available network resources. Advanced and fast DSP at both transmitter (Tx) and receiver (Rx) sides is the core of flexible transceivers, allowing the variation of the data rate and its adaptation to the light-path quality. There exist a number of methods for achieving flexibility. Among them are orthogonal-frequency division multiplexing (OFDM) [5], coded modulation [6, 7], rate-adaptive modulation [8], probabilistically shaped QAM modulation [9], four-dimensional optimized modulation formats [10–12], nonlinearity-tolerant four-dimensional 2A8PSK [13], time-division hybrid multiplexing (TDHM) [14, 15] and baud-rate adaptive approaches [16]. Another approach is the hybridization between known modulation formats in the quadrature domain, i.e. the in-phase (I), quadrature (Q), and the two polarization axes X and Y. This approach, which is referred to in the literature as flexible pulse amplitude modulation (Flex-PAM) [3, 17], offers a simple but yet powerful tool for the variation of data rate granularity (bits per symbol). In this paper, we experimentally investigate hybrid modulation.
as a promising candidate for future flexible communication systems. Flex-PAM works at every integer number of bits per symbol and therefore enables a flexible granularity, while keeping the DSP simple. We successfully transmit Flex-PAM over 189.6, 366.7 and 579.3 km in a lab experiment operating with a bits per symbol granularity ranging from four to 12 in a 37.5 GHz grid wavelength division multiplexing (WDM) 32 Gb/s transmission system, enabling maximum net data rates of up to $5 \times 312$ Gbit/s.

The paper is organized as follows: In Section II we summarize and illustrate the principle of hybrid modulation with Flex-PAM and define its lower bound performance in terms of the bit error ratio (BER). The key components of Tx and Rx DSP are discussed in Sections III-A and III-B, respectively. The simulation and experimental setup for the investigation are described in Section IV, followed by an analysis and discussion of the results in Section V. Finally, the paper is concluded in Section VI.

II. FLEXIBLE MODULATION THEORY

Using coherent optical modulation transmission systems, one can modulate the carrier light-wave in four dimensions, namely the in-phase (I) and quadrature (Q) in each of the two polarizations X and Y [18]. The resulting modulation dimensions are denoted in this paper as XI, XQ, YI and YQ, as shown in Fig. 1a. Flexible modulation uses this property to enable variable bit-rates by combining different modulation formats in several or all available modulation dimensions. Using the Flex-PAM technique, flexibility can be achieved by hybridization of simple real pulse amplitude modulation (PAM) modulation formats in the four available modulation dimensions, enabling tunability between all the integer values of modulated bits per symbol. Although Flex-PAM allows a lower granularity than its contenders, e.g. TDHM (which can achieve a non-integer granularity [15]), it offers in return other important advantages [3, 15, 17]: First, one can use four possible slower telecommunication flows for traffic grooming as different incoming tributaries. Depending on the traffic demand, the flows can be assigned to different axes, each carrying an M-PAM modulation with different modulation levels $M$ and consequent rates. Second, and more importantly, the Tx and Rx structures including their DSP for Flex-PAM modulation are less complex compared with the TDHM structures as well as the other approaches mentioned in Section I. In particular, the Tx structure for Flex-PAM is basically the same as the one for standard PDM-QAM, which is based on electro-optical (E/O) conversion by nested Mach-Zehnder modulators (MZMs) and coherent detection at the receiver with its DSP. A general Flex-PAM frame structure can be found in Fig. 1a. The degree of freedom using Flex-PAM is defined by the (standard) modulation formats used in the four possible dimensions. Different parameter choices yield different performances [17]. In theory each modulation axis can hold an individual modulation format with its own power and corresponding signal-to-noise ratio (SNR). Restricting the general Flex-PAM frame composition (I in Fig. 1a) to two nearest-sized PAM formats, namely $M_1 = M$ and $M_2 = 2M$, simplifies the frame composition (II to V in Fig. 1a), while keeping the granularity unchanged [3]. In this example, four different combinations are considered, yielding a total number of $4N$ (II), $4N + 1$ (III), $4N + 2$ (IV) and $4N + 3$ (V) bits per symbol. The BER of M-PAM as a function of the SNR is defined as [3]

$$\text{BER}_{\text{PAM}}(\text{SNR}, M) = \frac{M - 1}{M \log_2(M)} \text{erfc}\left(\frac{3\sqrt{\text{SNR}}}{M^2 - 1}\right). \quad (1)$$

Using (1), the general BER term for Flex-PAM can be determined:

$$\text{BER}_{\text{FPAM}}(\text{SNR}, M) = \frac{1}{N_{\text{bits}}} \sum_{n=0}^{3} \log_2(M(n)) \cdot \text{BER}_{\text{PAM}}(\text{SNR}(n), M(n)), \quad (2)$$

wherein $N_{\text{bits}} = \sum_{n=0}^{3} \log_2(M(n))$ represents the total number of modulated bits per symbol and the two arrays $\text{SNR} = [\text{SNR}_{\text{XI}}, \text{SNR}_{\text{XQ}}, \text{SNR}_{\text{YI}}, \text{SNR}_{\text{YQ}}]$ and $M =$

![Diagram](image-url)

**Fig. 1.** a) A General Flex-PAM frame structure (I) composed of four PAM modulations of orders $M_{\text{XI}}, M_{\text{XQ}}, M_{\text{YI}}$, and $M_{\text{YQ}}$ and their corresponding numbers of bits per symbol according to $N = \log_2(M)$. Note that only the optimum frame compositions by means of the minimum BER strategy are shown (II - IV). b) Simulated received Flex-PAM constellations for an AWGN scenario at a BER of $3.8 \times 10^{-3}$ using the minimum BER strategy regarding the power ratio. In compositions II and III $N = 1$, whereas $N = 2$ for compositions IV and V. All depicted constellations are normalized to an equal mean power of one.
[M_{X1}, M_{XQ}, M_{Y1}, M_{YQ}] indicate the SNR and modulation levels \( M \) of the four modulation dimensions, respectively [17]. In the following, the nomenclature \( F_{\text{PAM}} = [N_{X1}N_{XQ}N_{Y1}N_{YQ}] \) is used to describe each Flex-PAM modulation by its modulation order \( M \) and the corresponding bits per symbol \( N = \log_2(M) \) in each dimension. In Fig. 1b, the combination [1111] (II in Fig. 1a) stands for PDM-quadrature phase-shift keying (QPSK), whereas [1112] (III in Fig. 1a) results in QPSK and 8-QAM modulations in the X and Y polarizations, respectively. In the same manner, other hybrid modulation formats can be generated, as can be seen in the same figure for combinations [2323] (IV in Fig. 1a) and [2333] (V in Fig. 1a). An important optimization parameter in Flex-PAM is the power ratio (PR) between the two used modulation formats [3]. The aforementioned frame structure restriction enables the following four PR strategies. First, two straightforward methods to set the power ratio \( \text{PR}[\text{dB}] = \text{SNR}_1 - \text{SNR}_2 \) are either leaving the used modulation formats at the same power (PR = 0 dB) or using the same Euclidean constellation distance. Second, there are two favorable PR setting strategies between the two used modulation formats \( M_1 = M \) and \( M_2 = 2M \) [3], which outperform the above-mentioned methods: The first controls the SNR of the two modulation formats, such that all sub-streams operate at the same BER. Therefore, the required SNR has to be determined for each modulation format to achieve a target BER, denoted in this paper as \( \text{BER}_{\text{target}} \). Inverting the BER expression in (1) yields

\[
\text{SNR}^1_{\text{req}} = \frac{M^2 - 1}{3} \left( \text{erfc}^{-1} \left( \frac{M \log_2(M)}{M - 1} \cdot \text{BER}_{\text{target}} \right) \right)^2
\]

and

\[
\text{SNR}^2_{\text{req}} = \frac{4M^2 - 1}{3} \left( \text{erfc}^{-1} \left( \frac{2M \log_2(2M)}{2M - 1} \cdot \text{BER}_{\text{target}} \right) \right)^2,
\]

for the two SNR or the PR, respectively [3]. The second and optimal strategy suggests that there is one pair \((\text{SNR}^1_{\text{req}}, \text{SNR}^2_{\text{req}})\) for both modulation formats that achieves a global minimum BER. The two described favorable strategies are compared in Fig. 2, by means of their corresponding BER waterfall curves in an AWGN scenario. As can be seen, the two strategies show nearly equal performance, where the minimum BER approach gives a slight additional advantage of around 0.2 dB (at \( \text{BER}_{\text{target}} = 2.2 \cdot 10^{-2} \)). Unfortunately, this strategy leads to burst errors, which is disadvantageous when employing forward error correction (FEC) techniques. Therefore, the strategy of operating at the same BER for all implemented modulation formats was chosen to be used in this work. A more in-depth analysis of the theoretical operational strategies can be found in [3].

### III. Digital Signal Processing

#### A. Transmitter

The used DSP setup is given in Fig. 3. As bit pattern a \( N_{\text{bit}}\)-ary de-Brujin pseudo-random multilevel bit-sequence (PRMS) of symbol length \( N_{\text{sym}} \) was generated for the channel under test. For every interfering channel, a random \( N_{\text{bit}}\) \( \cdot N_{\text{sym}} \) bit sequence was generated directly. \( N_{\text{sym}} \) for each measured BER was chosen to reach at least 100 errors in every measurement and 1000 errors for the simulative results. The sequence was then mapped to the chosen four-dimensional Flex-PAM modulation scheme according to Fig. 1a. The \( N_{\text{sym}} \) modulation symbols were re-sampled and shaped with a root raised-cosine (RCOS) filter and a roll-off factor \( \alpha = 0.2 \). Linear frequency domain pre-compensation filters were used to compensate for Tx and Rx distortions, e.g. inter-symbol interference (ISI) caused by low-pass (LP) characteristics of the DAC, electrical amplifier, the nested PolMux-MZM at the transmitter side, the coherent PDM PIN-based receiver and the analog-to-digital converter (ADC) at the receiver side. Linear pre-compensation of low-pass filtering effects automatically increases the power of higher frequencies and correspondingly changes the requirements on quantization, but compared to the advantages of linear pre-compensation the impact on the performance is rather low. To obtain the complex frequency response, we transmitted a broadband OFDM signal and estimated the response \( H_{\text{Rx}} \) by calculating the ratio between received and transmitted signal on all four channels after correlation. This procedure also found the skews of transmitter and receiver, which were used to directly compensate for rough timing offsets in ADC and DAC. In addition, the noise contributions (AWGN, quantization noise) were removed using averaging.

A resulting re-sampled magnitude response obtained in the experimental investigation is shown as an inset in Fig. 7.

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1. All results in this paper are given with respect to the two suitable FEC implementations, namely hard decision (HD)-FEC operating at a BER = 3.8 \( \cdot 10^{-3} \) and soft decision (SD)-FEC operating at a BER = 2.2 \( \cdot 10^{-2} \) [19].
B. Receiver

The received and sampled signal was processed off-line. Prior to post-processing, the signal was first re-sampled to have two samples per symbol, and a front-end correction DSP unit compensated for the Tx and Rx imperfections in terms of residual amplitude, phase and timing imbalances. To avoid the residual time variant IQ-imbalance, an adaptive blind-moment estimation algorithm was used for imbalance estimation during front-end correction [20]. At the beginning of the digital signal processing there is a static frequency-domain finite impulse response (FIR) filter to combat the time-invariant chromatic dispersion (CD) [21]. A coarse frequency correction using a blind joint FFT-based algorithm was then applied to remove the frequency uncertainty and the slow drift (over time) between the Tx and Rx continuous wave (CW) lasers, which occurs as a result of intradyne coherent detection [22]. In the next stages, a matched filter (matched with the Tx RCOS pulse-shaping filter [21]) was used to maximize the SNR, followed by an adaptive complex- valued 2 × 2 butterfly equalizer to compensate for polarization mode dispersion (PMD) and residual CD. For adaptation of the butterfly-equalizer taps we used a dual polarization (DP) decision directed (DD) least mean squares (LMS) algorithm (pre-trained by 128 known symbols and the minimum-mean square error (MMSE) criterion for faster convergence) with a sufficient number of equalizer coefficients (not exceeding 12). For final fine frequency and phase recovery, the Blind-Phase-Search (BPS) algorithm with 64 test angles is used [23].

IV. SETUP

A. Simulation

A schematic of the general setup is given in Fig. 4. Our simulations are based on a five channel Nyquist 37.5 GHz WDM setup over an dispersion uncompensated and amplified multi-span link using our own MATLAB routines. At the transmitter side, we implemented five parallel standard DSP schemes as described in Section III-A. The resulting sequences were converted to an analog signal using a sample and hold approach with an up-sampling factor of 32, a Butterworth filter of 4th-order with a cut-off frequency of 13.1 GHz and a quantization unit to emulate specific DAC characteristics. If not specified, we set the transmitter quantization to 5.2 bits according to the given effective number of bits (ENOB) hardware specifications. The signal was transferred to the optical domain using a DP optical in-phase and quadrature modulator (IQ-MOD) with nested MZMs, each with a cosine transfer function, followed by an additional Butterworth filter of 3rd order with a cut-off frequency of 39.8 GHz to emulate its LP properties. An imbalance between I and Q axes as well as X and Y-polarizations was applied according to the experimental setup. The Tx local oscillator (LO) was set to a center wavelength of $\lambda_f = 1550.142$ nm ($f_c = 193.4$ THz) with a line-width of 80 kHz. Before multiplexing, each optical signal was boosted with an ideal Erbium-doped fiber amplifier (EDFA) to a specific launch-power (shown in Fig. 8). The five individual optical data signals were up-sampled to an overall up-sampling factor of 128 and added together to simulate channel multiplexing. To simulate the nonlinear standard single-mode fiber (SSMF) with PMD, we implemented the coupled nonlinear Schrödinger equation (CNLSE) with the Manakov approach [24]. The SSMF was implemented with an attenuation coefficient $\alpha_{\text{fiber}} = 0.2$ dB/km, a dispersion parameter $D_{\text{fiber}} = 17$ ps/(nm-km) and a nonlinearity coefficient $\gamma_{\text{fiber}} = 1.3$ (W · km)$^{-1}$. Each span was composed of an SSMF with a length between 81.3 and 100.2 km according to the lab setup. The accumulated span-loss for each span was compensated by an ideal EDFA. For noise-loading, a variable optical attenuator (VOA) combined with an EDFA (noise-figure of 4.8 dB) was used at the receiver. Prior to detection, the signal was separated and filtered using a 35.5 GHz (3 dB) Gaussian filter of 3rd order. The optical analog signal was down-converted using a coherent-receiver with a 90°-hybrid and a 3 dB bandwidth of 25 GHz and an additional LO operating at a wavelength of $\lambda_f = 1550.142$ nm. The Rx LOs were simulated with a line-width of 5 kHz. The electrical analog signal was then detected and converted into the digital domain using an ADC operating at a sampling rate of 100 GSa/s and a vertical resolution of 5.6 bits (ENOB) unless otherwise specified. We have chosen the simulation parameters according to the hardware specifications conducted...
in the experimental evaluation explained in the next paragraph.

B. Experiment

The sequence pre-generated by an off-line DSP according to Section III-A was re-sampled and fed to four Tektronix AWG70002A DACs working at a sampling frequency of 50 GSa/s, each with a vertical resolution of 8 bits (a specified ENOB of about 5.2 bits). The four independently-working DACs were synchronized with an additional synchronization hub to operate as a single four channel DAC. The generated analog signals ($V_{\text{peak-peak}} = 1\text{ V}$) were amplified by two SHF 807 matched electrical amplifiers for each polarization to reach the required swing voltage of the following 100 GHz DP-IQ-MOD 6M100300 by Oclaro with $V_{\text{in}} = 8.3\text{ V}$ for optical modulation. An external cavity laser (ECL) with a line-width of 80 kHz oscillating at $\lambda = 1550.142\text{ nm}$ was used as Tx LO. Note that the required PRs were sought by simulations, but the DAC output amplitudes were adapted slightly in iterations to match the hardware imperfections and reach equal BERs in all dimensions. The evaluated optimal DAC amplitudes were in a range of 0.4 and 0.5 V ($\pm 10\%$ of $V_{\text{peak-peak}}$), while 0.5 V should give the theoretical optimum. The amplitudes were re-determined several times a day and due to strong temperature fluctuations, this led to slightly new optima every optimization. To generate the interferer channels at a 37.5 GHz channel spacing, the inverted four channels generated by the DAC were delayed electrically and fed to a Tektronix OM5110 Multi-Format Optical Transmitter driven by four polarization multiplexed Agilent ECLs with a line-width of 80 kHz operating at $f_{\text{int,n}} = f_c + n \cdot 37.5\text{ GHz}$ each, where $n \in \{-2, -1, 1, 2\}$ corresponds to the neighboring channels. The center channel and interferer channels were multiplexed and gain flatttened with a Finisar WaveShaper 9000S. We made all the necessary adjustments such that all channels have the same optical power as well as the same bias and driving voltages relative to $V_{\text{in}}$. Subsequently, we integrated two microcontroller based bias control (ABC) units: (1) external ABC by IDphotonics for the center channel and (2) internal OM5110 ABC by Tektronix for the interferer channel. A VOA with power-monitoring was used to vary the launch-power into the fiber for each measurement. The launch powers $P_{\text{fiber}}$ and corresponding OSNRs evaluated for the 189.6 km measurement can be found in Fig. 8. These values are also to be used as a reference to interpolate further values for the rest of the measurement series. The signal was then transmitted over SSMF fiber spans ranging from 81.3 to 100.2 km followed by an EDFAs each, resulting in an overall transmission length $L$ between 0 and 579.3 km (0, 4, 6 spans). The resulting OSNR was measured by integration and normalization with an optical spectrum analyzer (OSA) operating at 0.01 nm bandwidth and 2001 sampling points as depicted in an inset in Fig. 4 [25]. After transmission, the channel under test (center channel) was filtered with an optical filter (a second order Gaussian filter) with an optimized 3 dB bandwidth of 35.5 GHz. The power $P_{\text{Rx}}$ into the u2t coherent (CO) receiver was kept as constant as possible using a VOA (between $-15$ and 5 dBm, depending on the OSNR and the corresponding received power). An autonomous ECL with a line-width of 5 kHz working at $\lambda = 1550.142\text{ nm}$ was used as an optical Rx LO for the center channel under test. After detection, the signal was sampled with a Tektronix DPO70000SX ADC operating at a sampling rate of 100 GSa/s and a vertical resolution of 8 bits (ENOB of about 5.6 bits). The sampled signal was processed off-line according to the Rx DSP schematic given in Section III-B.

V. RESULTS

A. Simulation Performance

The first step of our simulation analysis was to verify the simulation setup and the conducted theoretical BERs in Section II. Therefore, we simulated a B2B optical system including a simplified transmitter with perfect up-sampling and linear E/O conversion, neglecting LP influences and quantization effects. At the receiver we used perfect down-sampling and opto-electrical (O/E) conversion and congruent simplified DSP. The simulated required optical signal to noise ratios (rOSNRs) to reach the HD-and SD-FEC limits of $3.8 \cdot 10^{-3}$ and $2.2 \cdot 10^{-2}$, respectively, are shown in Fig. 5. As can be seen, these results match with the results shown in Fig. 2. Note that the rOSNR values are shifted by 4 dB with respect to the required SNR in Fig. 2, which is the ratio between the symbol rate $R_s = 32\text{ Gbd}$ and the OSNR reference bandwidth of 12.5 GHz.

Since higher-order modulation formats are more influenced by quantization noise than lower ones, the effects of quantization noise due to DAC and ADC imperfections were investigated in detail by adding quantization noise to the simulation setup. The penalty resulting from a limited ENOB of the DAC and ADC is shown in Fig. 6. The results are presented in rOSNR penalty $\Delta \text{rOSNR}$ in dB compared with the optical B2B AWGN reference rOSNR depicted in Fig. 5 for the HD-FEC and SD-FEC limits. The ENOB was varied between eight and four in steps of 0.2 bits, which results in 255 to 15 quantization levels, respectively. Using a quantization with 8 bits almost results in no penalty, but we see a significant

![Fig. 5. Simulation results showing the required OSNR (rOSNR) at the HD and SD-FEC limits vs. F_{PAE} modulations for an optical B2B transmission. The gross bit-rates corresponding to the used modulations are given as labels at the top of the figure.](image-url)
loss in ΔrOSNR starting from 6 bits ongoing. The results are plotted for Flex-PAM modulation reaching from [1111] (blue + circles) to [3333] (orange + stars) according to the same color coding in Fig. 2. The complexity of the modulation formats has different manifestation in the rOSNR development over quantization as expected. In Fig. 6 a clustering of the curves depending on the modulation can be observed. The results for F\textsubscript{PAM} = [1111] (blue + circles) represent the first cluster c1 using only PAM-2 modulation. The second cluster c2 contains all modulation formats that include PAM-2 or PAM-4 modulation, e.g. F\textsubscript{PAM} = [2111] or F\textsubscript{PAM} = [2222]. The last and most disturbed cluster c3 treats all the combinations of PAM-4 and PAM-8 modulation, e.g. F\textsubscript{PAM} = [3222].

The effects of ISI due to LP filtering (caused by hardware restrictions in Tx and Rx) and linear compensation variants are shown in Fig. 7. The figure shows the performance of F\textsubscript{PAM} = [1212] for a system including LP filtering, a XY-skew of 1 ps, an IQ-skew of 3 ps, an amplitude IQ-imbalance of 2%, a phase IQ-imbalance of 4°, 5.2 quantization bits at Tx and 5.6 quantization bits at Rx. The results include a setup without equalization (dashed, blue + squares), with pre-trained LMS equalizer running with a sufficient number of equalizer coefficients (12 taps) (dashed, green + squares) and with a reduced number of LMS coefficients (dashed, red + squares) of two taps. The results with the discussed pre-compensation using frequency domain equalization in addition to the pre-trained LMS working with 12 (purple + circles) and two taps (yellow + circles) are plotted as solid lines. As expected, both setups perform better with pre-compensation, whereas the increase of the equalizer coefficients yields only a marginal improvement. For the rest of the paper, all results include pre-compensation as well as adaptive LMS equalization (EQ) with sufficient number of equalization taps.

While all previous results from Figs. 5-7 deal with specific simplified scenarios, Fig. 8 depicts the final complete simulation results including all discussed impairments, a full DSP with pre-compensation and 12 equalization taps of the Flex-PAM investigation for a transmission over 189.6 km. In that figure theoretical B2B reference performance (solid, black + no markers), simulative (solid, colored + no markers) and experimental (solid, black + circles) BER waterfall curves over an OSNR ranging from 9 to 36 dB are shown in a separate plot for each Flex-PAM format under test. To fully support the complete experimental design in our final simulations, we additionally included the following impairments and their corresponding DSP units for compensation as discussed in Section III: polarization (XY) skew (1 ps), IQ skew (3 ps), IQ imbalance (2% amplitude and 4° phase), Tx and Rx LP characteristics, LO frequency offset (300 MHz) and phase noise (Tx: 80 kHz, Rx: 5 kHz). The impact of quantization and nonlinearities can be seen with higher modulation order, starting at F\textsubscript{PAM} = [2223] regarding the simulation results. To illustrate the effect of quantization with non-linearities, simulations with different transmitters quantization (5.2 bits to 4.6 bits in steps of 0.2) are shown.

### B. Experimental Performance

The conducted thorough simulative investigations were used as reference for the experimental results. The performance of Flex-PAM was determined by means of the OSNR required to reach the two FEC limits. For this, all BER waterfall...
curves were recorded for 0 km (B2B), 189.6 km, 366.7 km and 579.3 km.

Fig. 8 shows in addition to simulation results discussed in Section V-A the Flex-PAM waterfall curves for a transmission over 189.6 km (solid, black + circles). As can be seen, the experimental curves show similar behavior, but have a higher error floor in terms of BER compared to the simulation operating with quantization given by the hardware manufacturer. This increase is fundamentally supported by quantization noise higher than expected in combination with fiber nonlinearities. As explained in Section V-A and shown in Fig. 6 the higher modulation formats are more affected by quantization noise and nonlinearities.

Fig. 9 shows the rOSNR penalty with respect to the B2B AWGN rOSNR for all eight Flex-PAM modulation formats tested in the scope of this investigation. Notice that the rOSNR penalty results are given separately for the HD-FEC (upper row) and SD-FEC (lower row) for 0 (B2B), 189.6, 366.7 and 579.3 km from left to right. The black circles describe the experimental results, while the red rectangles (Tx quantization of 5.6 bits) and blue triangles (Tx quantization of 4.8 bits) represent the simulative results. Please note that missing rOSNR values for higher modulation formats mean that the specific FEC limit could not be reached. The obtained penalties from Fig. 9 added to the corresponding results from Fig. 5 give the rOSNRs for each modulation format evaluated in the lab. A mean experimental implementation penalty of around 1.5 dB can be observed in Figs. 8 and 9. It can be seen that the simulative as well as the experimental results follow the same trend and have a larger deviation from the AWGN theory with increased number of modulated bits per symbol and modulation order \( M \). Even in the B2B case the penalty operating with HD-FEC rises with modulation order, which indicates an additional implementation penalty by ADC and DAC imperfections. This implies a higher DAC and ADC quantization or nonlinearity as expected, when operating at the hardware specification edge at baud-rates of 32 Gbd. While QPSK or [1111], [1112], [1212], [1222] and 16-QAM or [2222] show a relatively slow increasing slope, every modulation format using PAM-8 has a higher penalty increasing with the number of dimensions making use of PAM-8 modulation. The curves with a quantization ENOB...
of 4.8 bits (blue triangles) in Fig. 9 and the simulation curves from Fig. 8 support this conclusion. This result in an overall rising error floor with modulation order and corresponding rOSNR, especially working with a HD-FEC. Working with an SD-FEC we achieved a maximum spectral efficiency (SE) of 6.8, 6.1 and 6.1 (bit/s)/Hz experimentally for 189.6, 366.7 and 579.3 km, respectively. Working with an HD-FEC limit resulted in 6.1, 5.3 and 5.3 (bit/s)/Hz, respectively.

VI. CONCLUSION

In this contribution, we experimentally demonstrated the use of Flex-PAM as a promising candidate for flexible transceivers in link-adaptive fiber-optic systems. The performance of Flex-PAM was tested in a 5 × 32 Gbd WDM transmission using a 37.5 GHz grid over up to 580 km, yielding data rates ranging from 113.6 to 312.2 Gb/s per channel and a maximum achievable net bit-rate of 1.561 Tbit/s. While the implementation penalty for Flex-PAM modulation up to [2222] was relatively low (about 1.5 dB), a rising penalty was observed for modulation formats including PAM-8 modulation. Albeit the used hardware had a stated ENOB of 5.2 bit for the transmitter and 5.6 bit for the receiver, quantization is a big issue, since a vertical resolution inside the range between 5.2 and 5.6 shows a sharp increase in the observed penalty for higher modulation (Figs. 6 and 8). The experimental results indicate an additional error floor for higher modulation formats, which is the result of higher quantization noise than expected. Nevertheless, Flex-PAM showed an outstanding flexibility while operating with a simple full adaptive DSP in an experimental environment capable of handling a remarkable range of bit-rates (128 to 320 Gbit/s) and fiber-lengths.

ACKNOWLEDGMENT

The authors thank Tektronix for the loan of the AWG70002 arbitrary waveform generator (AWG), the optical unit OM5110 and digital oscilloscope DPO70000S.

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