Comparison of Direct-Detection Approaches for High-Speed Datacenter Campus Networks

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Abstract *We experimentally compare virtual carrier based self-coherent detection with Kramers-Kronig reception and chromatic dispersion pre-compensation for 10km C-band transmission at gross bit rates between 120Gb/s and 300Gb/s. The latter scheme shows a better performance up to 230Gb/s although being more demanding for the DAC.*

Introduction

To cover the demand for growing datacenter (DC) traffic, companies are building mega DCs and clusters of DCs, which require optical networks to extend up to 10 km[1]. The transmission rates in next generation Ethernet are expected to be 800 Gb/s or 1.6 Tb/s[2]. To keep the hardware demand low, bit rates beyond 200 Gb/s per lane are of interest. These rates can be reached with high-order PAM formats and advanced receiver digital signal processing (DSP)[3]-[5]. However, the increased transmission distances lead to significant performance degradation in such high-speed intensity-modulation (IM) and direct-detection (DD) systems. These systems are limited by chromatic dispersion (CD) and the resulting power fading effect after direct detection. A straightforward approach to minimize the CD effect is the transmission in O-band, but even in this case, the outer channels in a coarse wavelength division multiplexing (CWDM) system suffer from relevant dispersion effects[6], which can be additionally enhanced by modulator chirp.

Several approaches to overcome dispersion effects in DD systems have been shown[7]-[11]. One option is self-coherent detection based on single-sideband (SSB) transmission, which avoids the power-fading effect[7]-[10]. This approach is often combined with Kramers-Kronig (KK) reception, which overcomes signal-signal beat interference (SSBI) and therefore results in improved performance. Alternatively, electronic dispersion compensation (EDC) can be applied in the transmitter DSP of a DD system, if the necessary field information is available[11].

In this paper, self-coherent detection with KK reception and the transmitter side EDC are compared on the same transmission system at gross data rates between 120 Gb/s and 300 Gb/s. PAM6 transmission with transmitter EDC shows superior performance, if Volterra nonlinear equalization (VNLE) is applied, but the reachable rates are limited by the digital-to-analog converter (DAC) sampling rate. For self-coherent detection with KK reception, a less complex linear equalizer is sufficient and the performance at low optical signal-to-noise ratio (OSNR) is superior.

Considered Approaches

The field information of a signal can be recovered through DD if the transmitted signal is SSB and the carrier-to-signal power ratio (CSPR) is sufficiently high[12]. Several options to generate SSB signals exist[7]-[10]. These include optical SSB filtering[8], insertion of an optical carrier on the edge of the spectrum[9] and up-conversion of the digital signal[10]. A promising option is the virtual carrier approach[7]. As shown in Fig. 1 a), a digital tone is added at one the side of the signal spectrum in the transmitter DSP. The inphase-quadrature (IQ) modulator is driven at the null point so that no additional carrier is generated. This way, an optical SSB signal is generated without using additional optical components. An exemplary case is shown in Fig. 1 b). The full bandwidth of the transmitter components can be utilized for transmission of complex signals and the detection is done using a single photodiode. The phase of the signal can then be recovered by applying the well-known KK relations[13].

A second option to cope with the distortions caused by fiber dispersion is transmitter side EDC. This can be effectively done since in the transmitter DSP, in contrast to the receiver DSP of a DD system, the full field is known. Compared to conventional IM/DD systems, this approach requires a dual-drive Mach-Zehnder modulator

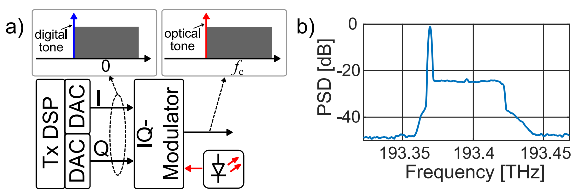

Fig. 1: a) transmitter configuration for SSB signal generation using a virtual carrier and b) exemplary optical spectrum of a resulting signal.

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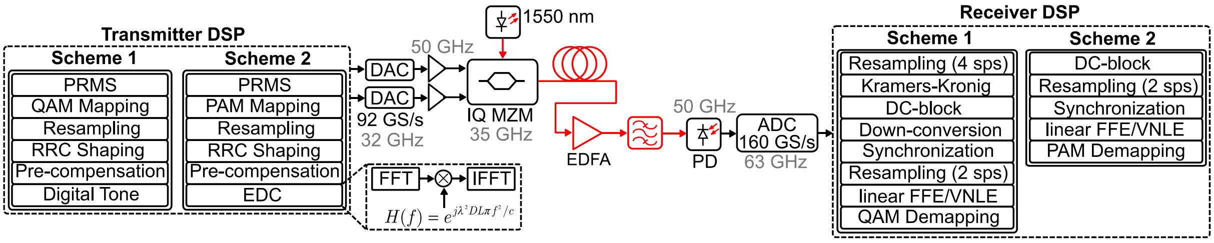

Fig. 2: Experimental transmission system setup and DSP. Scheme 1 shows the DSP steps for virtual carrier based QAM transmission and scheme 2 shows the DSP for PAM transmission with CD pre-compensation. PRMS: pseudo-random multilevel sequence, QAM: quadrature amplitude modulation, PAM: pulse amplitude modulation, RRC: root-raised cosine, EDC: electronic dispersion compensation, DAC: digital-to-analog converter, MZM: Mach-Zehnder modulator, EDFA: Erbium-doped fiber amplifier, PD: photodiode, ADC: analog-to-digital converter, FFE: feed-forward equalizer, VNLE: Volterra nonlinear equalization.

(MZM) or an IQ modulator instead of a simple intensity modulator. The amount of chromatic dispersion in the link needs to be known at the transmitter. This might make a feedback channel necessary for networks with unknown or varying dispersion.

Experimental Investigations

The experimental setup in Fig. 2 was used for the comparison of the two schemes described in the previous part. For the virtual carrier assisted transmission (scheme 1) QAM symbols were generated and resampled to the DAC sampling rate. Afterwards, root-raised cosine (RRC) pulse-shaping with a roll-off factor of 0.05 was applied and the signal was pre-compensated for bandwidth limitations. A digital tone was added on one side of the spectrum with a guard spacing of 1 GHz between signal and tone. The CSPR was optimized to 12 dB for all cases. For the PAM transmission with transmitter side EDC (scheme 2), the RRC roll-off factor was set to the value

$$\beta = f_{\mathrm{DAC}} / f_{\mathrm{sym}} - 1, \qquad (1)$$

where f_{DAC} is the DAC sampling rate and f_{sym} the symbol rate. After pre-compensation, the real valued signal was rotated by 45° in the complex plane by copying the same signal on I- and Q-parts to align the signal with the DC tone resulting from the MZM biased at quadrature point. Afterwards, EDC was applied on the complex signal.

The transceiver components configuration is identical for both schemes. The DACs were operated at a sampling rate of 92 GS/s and have a 3-dB bandwidth (BW) of 32 GHz. The analog signal was amplified by 50 GHz driver amplifiers and an IQ-MZM with a BW of 35 GHz modulates the signal on the optical C-band carrier. The IQ MZM was biased at null point for scheme 1 and at quadrature point for scheme 2, respectively. After transmission through the fiber, the signal is amplified by an EDFA followed by an optical filter. The input power into the photodiode (PD) was optimized and fixed to be 7 dBm for all measurements. The signal is detected by a 50 GHz PD and captured and digitized by a 63 GHz real time analog-to-digital converter at a sampling rate of 160 GS/s.

The receiver offline DSP for scheme 1 begins with a resampling of the signal to 4 samples per symbol (sps). This oversampling is necessary due to the nonlinear operations of the KK receiver. After applying the KK relations, the DC-part of the signal was removed and the signal is down-converted by the value of the symbol rate. The signal was synchronized and resampled to 2 sps before feed-forward equalization (FFE) was applied. Finally, the QAM symbols were de-mapped and the BER was calculated. For scheme 2 the DC-part of the signal was removed before it was resampled to 2 sps and synchronized. After FFE, PAM de-mapping and BER calculation were performed.

Results and Discussion

The results of the experimental investigations are shown in Fig. 3. In Fig. 3 a) the BER after 10 km transmission is shown as a function of the ONSR at a constant rate of 224/225 Gb/s for both schemes and different modulation formats. For scheme 1, the formats 16 QAM, 32 QAM and 64 QAM are considered, whereas PAM6 and PAM8 are considered for scheme 2. For all formats, transmission performance was measured under two DSP configurations: one is linear FFE with 120 kernels and the other is VNLE with memory lengths of 120, 5, 5 for the first, second and third order, respectively. The kernel weights were optimized using first a training-based and then a decision-directed least mean squares algorithm. While 16 QAM and 32 QAM reach a similar performance, 64 QAM shows the worst results for the considered rate. Linear FFE and VNLE show no difference in BER performance for all formats subject to scheme 1, thus in the following results only linear FFE is used. For scheme 2 higher OSNR values compared to scheme 1 can be reached, since the output power of the IQ MZM is higher due to the different bias point. PAM6 shows a better performance than PAM8 and

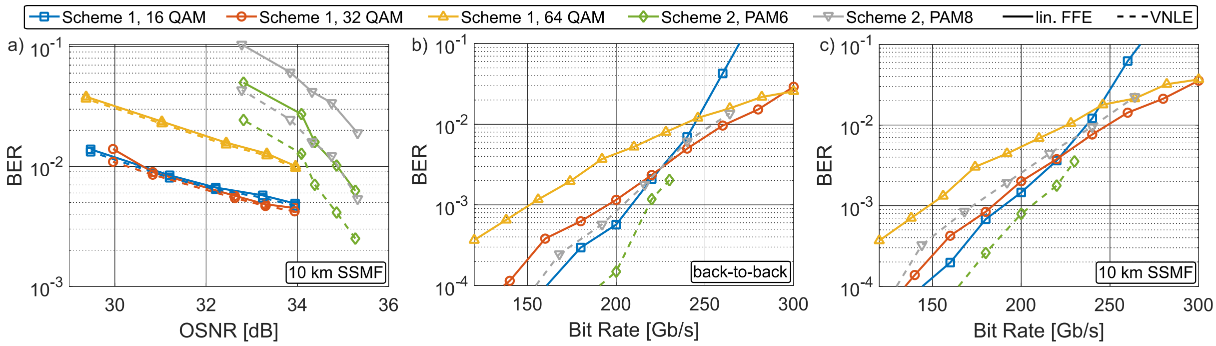

Fig. 3: Experimental results. a) shows the BER as a function of the OSNR for a bit rate of 224/225 Gb/s and 10 km transmission in C-band. b) and c) show the comparison of both schemes for rates between 120 Gb/s and 300 Gb/s for back-to-back and 10 km SSMF transmission. The solid curves were obtained based on linear FFE and the dashed curves on VNLE.

VNLE significantly improves the performance compared to linear FFE. Scheme 1 shows a better performance for lower OSNR values, while scheme 2 reaches the lower BER at the highest achievable OSNR.

In Fig. 3 b) the results for schemes 1 and 2 are compared in a back-to-back configuration for gross bit rates between 120 Gb/s and 300 Gb/s. For the back-to-back case, the EDC step of scheme 2 is skipped. Following the insights from the previous results, linear FFE is used for scheme 1 and VNLE for scheme 2. Scheme 2 with PAM6 shows the best results for rates up to 230 Gb/s because of the higher achievable OSNR as indicated in Fig. 3 a). Higher rates are not considered due to the limited DAC sampling rate. For scheme 1, 16 QAM shows the best performance for rates up to 225 Gb/s and 32 QAM for higher rates. 64 QAM performs similar to 32 QAM for a rate of 300 Gb/s. Scheme 2 with PAM8 performs slightly worse than 16 QAM for rates below 225 Gb/s and slightly worse than 32 QAM for higher rates. Fig. 3 c) shows the same comparison for transmission over 10 km standard single-mode fiber (SSMF). Scheme 2 with PAM6 still shows the best performance and the relative performance of PAM8 compared to scheme 1 is worse than for the back-to-back configuration.

All in all, scheme 1 cannot outperform scheme 2 for rates up to 230 Gb/s. However, higher bit rates can be transmitted at a limited DAC sampling rate, since the symbol rate can be halved due to the utilization of complex constellations. In contrast to scheme 1, scheme 2 needs an additional EDC step in the transmitter DSP, that requires knowledge about the chromatic dispersion in the link. The KK relations in scheme 1 require a digital oversampling to perform optimally. However, the symbol rate is significantly reduced compared to scheme 2 (halved comparing 32 QAM and PAM6 or 64 QAM and PAM8), so that the actual sampling rate is not increased. The reduced symbol rate also means that the equalization is performed at a lower sampling rate for scheme 1, but this advantage is balanced by the fact that an equalizer with kernels for the I- and Q-components is necessary. Scheme 2 needs to apply VNLE to outperform scheme 1 for 224/225 Gb/s transmission. Therefore, the better performance of scheme 2 comes at the cost of an increased DSP complexity.

Conclusions

We compared virtual carrier based self-coherent detection with Kramers-Kronig reception (scheme 1) and PAM transmission with chromatic dispersion pre-compensation (scheme 2) for datacenter campus networks that include distances of up to 10 km. The experiments were done in C-band using the same system for both schemes. Scheme 2 with PAM6 shows the best performance for rates up to 230 Gb/s, if Volterra nonlinear equalization is applied in the receiver DSP. The reachable rates are limited by the sampling rate of the digital-to-analog converter. Scheme 1 can increase the number of bits per symbol by using complex constellations. This way, higher bit rates can be realized at lower sampling rates. Additionally, linear equalization is sufficient, so that the DSP is less complex for scheme 1 compared to scheme 2.

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