Experimental demonstration of 10 Gb/s multi-level carrier-less amplitude and phase modulation for short range optical communication systems

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Abstract: Carrier-less amplitude and phase (CAP) modulation can be a good candidate for short range optical communications for considerable computational complexity reduction and simple system structure. In this paper, a detailed investigation on the digital filters in CAP modulation system is presented. An adaptive equalizer based on cascaded multi-modulus algorithm (CMMA) is used for the demodulation at the receiver. The impact of digital filter taps on system performance is investigated through comprehensive simulations and a 10 Gb/s CAP16 modulation system is demonstrated experimentally. The BER performance for different length of fiber link is measured. Compared with back-to-back (BTB) transmissions, 2 dB and 3.5 dB receiver power penalty are observed at BER of 10^{-3} for 20 km and 40 km fiber link respectively. It clearly demonstrates the feasibility of the CAP16 modulation for the short range transmission systems.

OCIS codes: (060.0060) Fiber optics and optical communications; (060.4080) Modulation.

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1. Introduction

A wide variety of data and services drive the demand for higher capacity optical communication systems. To meet this demand, there have been significant amount of research effort on improving the performance of an optical communication system. In long haul systems, the effort has mainly been on increasing system spectral efficiency and system reach. For the system of relative short distance, such as passive optical network (PON) and short reach optical communication system for data center interconnection, the effort has been more on finding the best tradeoff among cost, complexity, data rate and sensitivity due to large amount of transceiver units needed in such applications [1–5] and low-cost equipment is a must. A candidate technique that may provide good system performance using low cost optical components such as direct modulated laser (DML) and vertical cavity surface emitting laser (VCSEL) is carrier-less amplitude and phase (CAP) modulation. It allows relatively high data rate to be achieved using optical components of limited bandwidth. Comparing with alternative schemes such as quadrature amplitude modulation (QAM) and orthogonal frequency division multiplexing (OFDM), no electrical or optical complex-to-real-value conversion is necessary which involves a complex mixer and radio frequency (RF) source or optical IQ modulator. Neither does it require the discrete Fourier transform (DFT) that utilized in OFDM signal generation and demodulation. The CAP signal can be generated by using a digital filter with several taps and a higher order modulation can be realized, thus reducing the complexity of computation and system structure considerably.

A number of optical communication systems based on CAP have been demonstrated recently [7–9]. In [7] and [8], systems based on CAP16 and CAP64 are proposed but the bit rate is only 1.25 Gb/s and 300 Mb/s respectively. References [2]. and [9] present a CAP16 modulation system with bit rate up to 40 Gb/s. However, only limited results were presented with no BER characterizations. In addition, since the channel response is not known in practical situation, adaptive equalizers are required. However, there is no detailed description about the design of adaptive equalizer and the digital filters. Their impact on system performance of optical CAP modulation systems was also not analyzed in previously published works.

In this paper, we present a detailed investigation on the design of digital filters and their impact on system performance. An adaptive equalizer based on cascaded multi-modulus algorithm (CMMA) [10] is presented and used for the demodulation of CAP16 signal at the receiver. The impact of digital filter taps including shaping filters and matched filters on system performance is investigated through extensive simulations and a 10 Gb/s CAP16 modulation system is demonstrated experimentally. The BER performance for back-to-back (BTB) transmission and over 20 km and 40 km fiber link is measured.

2. Operating principle

2.1 System structure for CAP modulation

Figure 1 shows the schematic diagram of system structure based on CAP modulation. For the CAP signal transmitter, the original bit sequence is first fed to an encoder, which maps blocks of bits into complex symbols and let T be the symbol period. The coded sequence is up-sampled by a factor M, i.e. M-1 zeros is inserted between two consecutive input symbols. Then the in-phase and quadrature components of the up-sampling sequence are separated and sent into the digital shaping filter respectively. The outputs of the filters are subtracted. The generated CAP signal is passed through a digital-to-analog (D/A) converter and subsequently performs optical up-conversion by driving an intensity modulator (IM) between the minimum

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and maximum transmission. Note that the digital shaping filters and D/A converter are operating at a rate of M/T, but the bandwidth of the optical modulator is decided by the symbol rate of the system.

At the receiver side, direct detection is used and the received signal after analog-to-digital (A/D) converter is fed into two different matched filters to separate the in-phase and quadrature components. The matched filters and A/D converter are also operating at a rate of M/T. After down-sampling, an equalizer is employed for the complex signal and a decoder is utilized to obtain the original bit sequence.

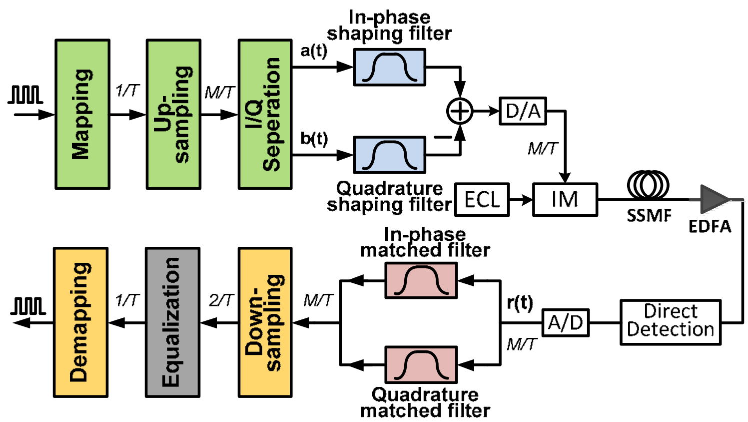

Fig. 1. Schematic diagram of system structure based on CAP modulation.

2.2 Theoretical foundations for of CAP modulation and demodulation

The generated CAP signal can be expressed as

$$s(t) = a(t) \otimes f_1(t) - b(t) \otimes f_2(t) \tag{1}$$

where $a(t)$ and $b(t)$ are the in-phase and quadrature components of transmitted bit sequence after coding and up-sampling process respectively. The functions $f_1(t)$ and $f_2(t)$ are the corresponding shaping filters and form a so-called Hilbert pair, as described in [6]. Note that the shaping filter should work at higher rate than the system symbol rate. So the up-sampling process here is used to match the rate of shaping filter and to obtain the output analog signal without the aliasing products. If the square-root raised-cosine function is employed as the baseband pulse response, the impulse responses of shaping filters are shown in Fig. 2(a).

Assuming that the channel response is ideal, the output of two matched filters at the receiver is expressed as

$$\begin{aligned} r_i(t) &= s(t) \otimes m_1(t) = \left(a(t) \otimes f_1(t) - b(t) \otimes f_2(t)\right) \otimes m_1(t) \\ r_q(t) &= s(t) \otimes m_2(t) = \left(a(t) \otimes f_1(t) - b(t) \otimes f_2(t)\right) \otimes m_2(t) \end{aligned} \tag{2}$$

Here $m_1(t) = f_1(-t)$ and $m_2(t) = f_2(-t)$ are the impulse response of the corresponding matched filters [11,12] and "$\otimes$" denotes convolution. Because the impulse response $f_1(t)$ and $f_2(t)$ is even and odd functions respectively [6], Eq. (2) can be simplified as

$$\begin{aligned} r_i(t) &= s(t) \otimes m_1(t) = a(t) \otimes h_{11}(t) - b(t) \otimes h_{12}(t) \\ r_q(t) &= s(t) \otimes m_2(t) = -a(t) \otimes h_{12}(t) + b(t) \otimes h_{22}(t) \end{aligned} \tag{3}$$

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where $h_{11}(t)=f_1(t)\otimes f_1(t)$, $h_{12}(t)=f_1(t)\otimes f_2(t)$, $h_{22}(t)=f_2(t)\otimes f_2(t)$. Figures 2(b)-2(d) indicates the joint impulse response $h_{11}(t)$, $h_{12}(t)$, $h_{22}(t)$ respectively. For the in-phase component in Eq. (3), the first term at the left side is the desired signal component. However, it is corrupted by the second term which comes from the quadrature component. It can be observed from the joint impulse response $h_{11}(t)$ and $h_{12}(t)$ that the maximum $h_{11}(t)$ coincides with the zero of $h_{12}(t)$. Therefore, the desired in-phase component can be extracted without inter-symbol interference (ISI) and the distortion that comes from the quadrature components at an appropriate sample time. The same conclusion can be obtained for the quadrature component in Eq. (3).

Because of the serious ISI, synchronization is very important in CAP demodulation. However, the appropriate sampling time is hard to decide and sampling time offsets will lead to subsequent signals seriously affected by ISI and the crosstalk between the in-phase and quadrature components. Considering that the distortions induced by either the ISI or crosstalk is linear, an equalizer is needed to recover the output signal of matched filters. The classic CMA [10] becomes popular because it is a blind algorithm and easy to realize. However, it is much less effective for multi-level signal modulation system, as the symbols in our system do not have constant amplitude and the error function in the adaptation process will not approach zero even for an ideal multi-level signal without distortion. Therefore, the cascaded multi-modulus algorithm (CMMA) with modified error function is used for the multi-level CAP signal [10] studied in this paper. It is modified slightly for the multi-level CAP system as there is only one complex input for the equalizer as supposed to polarization-multiplexed (PM) transmission in long-haul systems. Therefore, the synchronization process can be considered to be implicitly done through the blind adaptive CMMA.

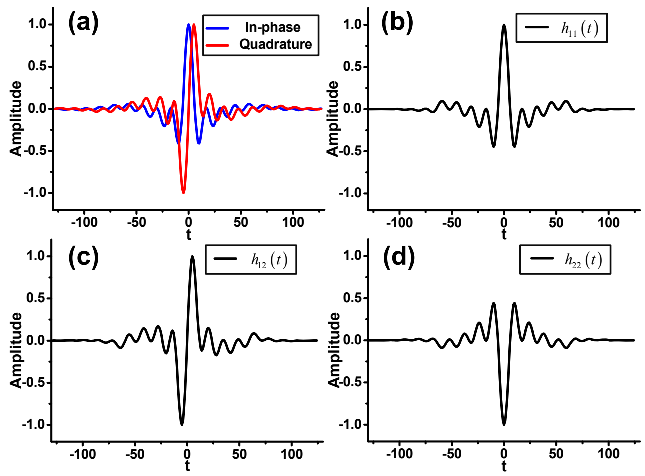

Fig. 2. The impulse responses of (a) in-phase and quadrature shaping filters, (b) $h_{11}(t)$, (c) $h_{12}(t)$, (d) $h_{22}(t)$.

3. Simulation results

The system described in the previous section is investigated through VPI Transmission Maker simulation. The original bit sequence is mapped into 4 levels and the symbol rate is set at 10 Gbaud. The computational complexity to generate and demodulate CAP signals is directly related with the number of taps of the shaping and matched filters. Therefore, it is important to investigate the requirements of the CAP filter in order to reduce the computational complexity while maintaining a good performance.

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Considering the relations between up-sampling factor and the required number of filter taps, we study a parameter ε defined as the ratio between the number of taps of shaping or matched filters and the up-sampling factor. The parameter ε can be also considered as the number of symbols that contribute to the ISI. The frequency response of the filters with different ε is shown in Fig. 3(a), and the impact of different number of taps of the shaping filters and matched filters on the system performance is shown in Fig. 3(b) at a received power of −22 dBm. It is observed that the BER decreases with the increasing ε of matched filters at the receiver for different ε of shaping filters at the transmitter, because the main lobe of the frequency response of the matched filters is narrower and the side lobe decreases faster with larger ε as shown in Fig. 3(a), thereby the inband OSNR after matched filter increases with larger ε, which leads to better BER performance. When ε exceeds 4, the frequency responses of matched filters are almost the same, and so the BER almost remains unchanged. It can be also found that BER performance degrades with increasing ε. This is because the induced ISI is more serious when ε is larger. Although the CMMA-based equalizer can compensate this linear distortion, the equalizer performance degrades in the presence of noise.

In order to make a further illustration, the BER performance of back-to-back (BTB) transmission with different ε is presented in Fig. 3(c). The BER curve at ε = 2 performs better than others. It is also found that the BER performance are closer with each other when the received optical power is increasing, suggesting that the OSNR of received signal has an influence on the tolerance of the equalizer to ISI. However, it does not mean that ε should be reduced. Figure 3(d) shows the power penalty at BER of 10^{-3} induced by chromatic dispersion (CD) for different values of ε. It shows that larger ε will improve the system performance. There is no inline dispersion compensation, so the bandwidth of generated CAP signal, which depends on ε, will affect the BER performance.

In conclusion, the parameter ε of shaping and matched filters will impact the BER performance of the transmission link. An appropriate value of ε should be chosen based on the requirement of computational complexity and transmission distance.

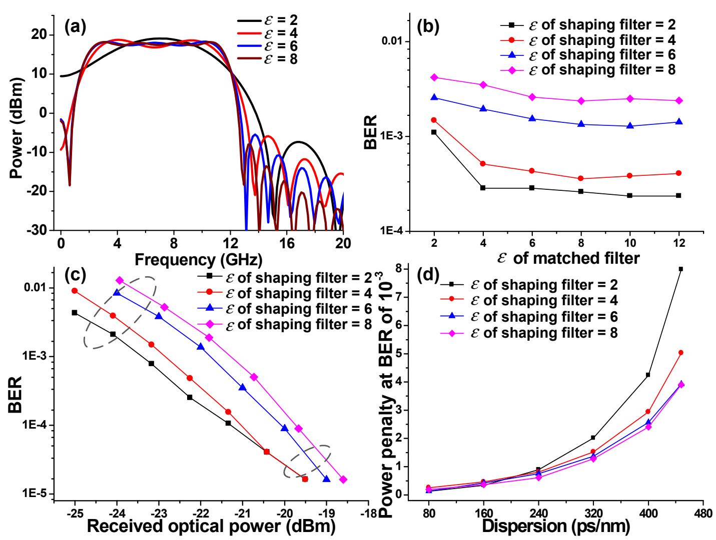

Fig. 3. (a) Frequency response of shaping or matched filters with different ε, (b) BER performance with different ε at received power of −22 dBm, (c) BER versus received optical power with different ε when ε equals 8, (d) Dispersion-induced power penalty for different ε.

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4. Experimental setup and results

The experimental setup of CAP16 transmission system is shown in Fig. 4. Firstly, the CAP16 signal is generated offline by MATLAB and the original data sequence is mapped into 4 levels and M-1 zeros are inserted between two consecutive input symbols, making the original mapping data up-sampled by a factor of 4 [7]. After I/Q separation, the two branches of data sequences is sent into two shaping filters with 32 taps. An arbitrary waveform generator (AWG) is used to produce the RF signal at 10 GSa/s so that the symbol rate 1/T of the CAP16 signal is 2.5 Gbaud. It should be noted that the symbol rate is limited by the bandwidth of the AWG used in our experiment, because there is power attenuation at higher frequency component. Some pre-emphasis techniques can be adopted to further increase the system's symbol rate. The square-root raised-cosine function is used as the baseband impulse response and its roll-off coefficient is set to 0.1. The excess bandwidth is set to 4%. A CW laser at 1551.6 nm from external cavity laser with a linewidth less than 100 kHz is used as the signal source. Then output RF signal is subsequently used to drive the intensity modulator (IM) between the minimum and maximum transmission. The fiber launch power is set at 2.8 dBm.

After fiber transmission, the CAP16 signal is detected by a photodiode (PD) with a responsivity of 0.65 A/W and sampled by an oscilloscope at a sampling rate of 10 GSa/s and processed off-line by MATLAB. Figure 5(a) shows the electrical spectrum of the received CAP16 signal. For the offline processing, the sampled signal is sent into two matched filters with 32 taps, so the in-phase and quadrature signals are separated and followed by a down-sampling process, because the CMMA algorithm operates at a rate of 2/T. Finally, the carrier phase recovery is employed based on the conventional 4th power Viterbi-Viterbi phase estimation algorithm. The captured and processed CAP16 bits for bit error counting are over 2 x 10^5.

We then measured the BER performance of the CAP16 system with BTB and over different transmission length and the results are shown in Fig. 5(b). It is found that the required received power at BER of 10^{-3} for BTB transmission is −24.9 dBm. Because there is no inline and offline dispersion compensation in this CAP16 system, there is receiver power penalty after fiber transmission. 2 dB and 3.5 dB receiver power penalty are observed at BER of 10^{-3} with 20 km and 40 km fiber link respectively. The recovered constellations at received power of −22.8 dBm with or without fiber link are also shown in Fig. 5(b), clearly demonstrating the feasibility of the proposed CAP16 transmission system.

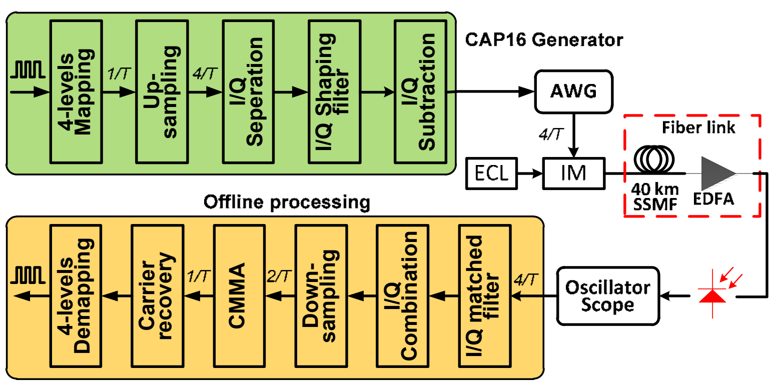

Fig. 4. Experiment setup of 10 Gb/s CAP16 modulation system. ECL: external cavity laser, IM: intensity modulator, AWG: arbitrary waveform generator.

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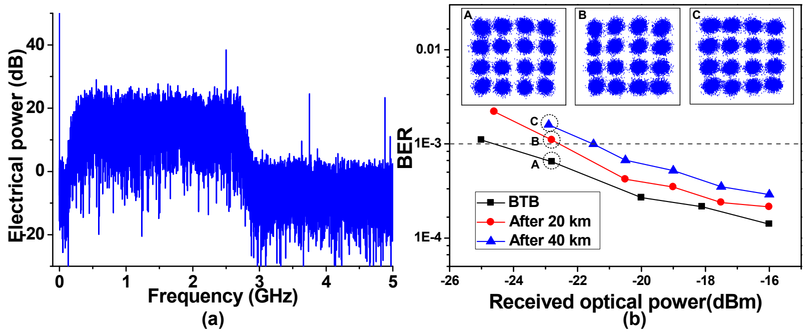

Fig. 5. (a) Electrical spectrum of received CAP16 signal, (b) BER performance versus received optical power with different transmission lengths.

4. Conclusions

In this paper, a detailed investigation on the digital filters in CAP modulation systems was presented through comprehensive simulations. An adaptive equalizer based on cascaded multi-modulus algorithm (CMMA) is used for the demodulation at the receiver and the impact of digital filter taps on system performance is investigated. It can be concluded that the shaping and matched filter taps will have considerable influence on the BER performance of the transmission system. The appropriate number of digital filter taps will be chosen based on computational complexity and transmission distance considerations. In addition, a 10 Gb/s CAP16 modulation system is demonstrated experimentally and the BER performance for different transmission distances are obtained. Compared with back-to-back transmission, 2 dB and 3.5 dB receiver power penalties are observed at BER of 10^{-3} with transmission distance of 20 km and 40 km respectively. It clearly demonstrates the feasibility of using CAP16 modulation for future short range transmission systems.

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