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2019

Zhang, X., Zhang, C., Zhu, M., Chen, C., Jin, W., & Qiu, K. (2019). Power budget of 25-Gb/s digital orthogonal filter multiple access PONs with 10G-class optics. *Optics Communications*, 434, 196-202. doi:10.1016/j.optcom.2018.10.063

<https://hdl.handle.net/10356/105592>

<https://doi.org/10.1016/j.optcom.2018.10.063>

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Power Budget of 25-Gb/s Digital Orthogonal Filter Multiple Access PONs with 10G-Class Optics

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Abstract:

Power budget is a key factor that should be considered in intensity modulation and direct detection (IM/DD) passive optical networks (PONs). Based on the low-cost 10G-class electro-absorption modulated laser (EML) operating at C-band, we not only experimentally and theoretically investigate the interplay between the chromatic dispersion, the chirp of EML, and the self-phase modulation (SPM), but also explore the influences of the resolution and timing jitter of the analog to digital converter (ADC). Both theoretical and experimental results show that high power budget can be effectively achieved by making full use of the high launch power induced SPM effect and EML-induced frequency chirp. Based on the 10-G transmitter optical sub-assembly (TOSA) and receiver optical sub-assembly (ROSA), we successfully realize 25-Gb/s signal transmission at C-band over 25-km standard single mode fiber (SSMF) in IM/DD PON system. Even the worst performing bands can still achieve 25-dB power budget. The proposed scheme in this work has a potential application in future cost-effective optical access networks.

Key words:—Passive optical networks, power budget, self-phase modulation, electro-absorption modulated laser, chirp.

1. Introduction

The next generation optical network is expected to support services with high capacity, high reliability and high power budget. To cope with the rapid increasing of networking traffic, many researchers have paid their attention to the investigations of alleviating the bandwidth bottleneck of metro systems and last-mile access networks [1]-[4]. For the bandwidth-hungry applications of next generation optical network, new standardizations such as 400 GbE scheme based on the existing 100 GbE low-cost and power-efficient components are widely considered as promising technical solutions to realize 400 Gb/s optical broadband transmissions [5], [6]. From the optical access network perspective, the roadmap which is pursued by the IEEE is also suggesting to update the 10 Gb/s passive optical network (PON) interface with the 802.3ca standardization effort, investigating higher data rates of possible physical layer for the next generation access network, expecting to be fully released by 2019 [7]. Although coherent detection systems can achieve high data rate as well as high performance in long haul transmissions, they require complex hardware, i.e., local oscillator and hybrid, which inevitably cause high cost [8], [9]. While the intensity modulation and direct detection (IM/DD) schemes with the advantage of low cost is much more attractive for PON systems [10]-[14]. So

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far, a large amount of researches have been done for various IM/DD PONs to achieve high data rate and long distance, such as orthogonal frequency division multiplexing (OFDM)-PON [15], [16], filter bank multicarrier (FBMC)-PON [16], [17] and carrier-less amplitude phase (CAP)-PON [18]. The IM/DD based PON systems not only relax the laser line-width requirement by using relatively low-cost transceivers such as electro absorption modulated lasers (EMLs) and directly modulation lasers (DMLs) at the transmitter, but also reduce the number of photodiodes (PDs) at the receiver, as only one single-ended PD is needed.

Digital orthogonal filtering-enabled channel multiplexing/demultiplexing with excellent DSP-enabled network elasticity and flexibility has been proposed for the cost-effective PON application scenarios [19-21], since it is not limited to a fixed modulation format, but allows a board range of signal modulation formats such as pulse amplitude modulation (PAM), non-return-to-zero (NRZ) and OFDM. In [19], Bolea mainly investigated the performance of digital orthogonal filter enabled OFDM channel with different filter taps and oversampling factors. In [20], Duan explored the tolerance to inter (optical network unit) ONU sample timing offset of the digital orthogonal multiple access system between different upstream ONUs. In [21] researchers were mainly concentrated on investigating the uplink capacity of the digital orthogonal multiple access system. By utilizing the digital orthogonal filtering techniques, our previous works reported in [22], [23] have been mainly focused on exploring the feasibility and performance of the DSP-enabled reconfigurable optical add/drop multiplexers (ROADMs), which is capable of providing add/drop operations at wavelength, sub-wavelength and orthogonal sub-band levels. However, the interplay between the chromatic dispersion (CD), the chirp of EML and self-phase modulation (SPM) plays a vital role in impacting the network performances, which are not considered in all-above mentioned works. In [24] researchers have demonstrated that the SPM-induced impairments can be partial mitigated by the residual positive CD in PAM4 SSB transmission system, but the effect of chirp is not considered.

In this paper, we theoretically and experimentally demonstrate the digital orthogonal filter multiple access-based IM/DD PONs utilizing the low-cost commercially-available 10-G optics. We not only investigate the power budget of the system, but also investigate the resolution and timing jitter requirements of the analog to digital converter (ADC). Low-cost 10-G electro-absorption modulator integrated with the laser-transmitter optical subassembly (EML-TOSA) in the C-band and 10-G PIN-receiver are used in transmitter (Tx) and receiver (Rx), respectively. Meanwhile, the interplay between the chromatic dispersion, the chirp of EML and SPM are also considered. The results imply that high power budget as large as 26 dB can be effectively achieved by deliberately increasing the optical signal launch power to enhance the SPM effect and employing the negative bias EML intensity modulation-induced frequency chirp. Experimental results show that 25-Gb/s transmission can be realized by utilizing 10-G optical transmitter and receiver, and the worst performing bands can still have 25-dB power budget.

2. Principle analysis

A. Principle of Digital Orthogonal Filter Multiple Access PONs

The principle of the digital orthogonal filter multiple access PONs is clearly illustrated in Fig. 1, where N independent OFDM channels are multiplexed utilizing DSP-based digital orthogonal filtering in the transmitter and demultiplexed in the receiver. In the optical line terminal (OLT), the individual signal for each ONU is first digitally encoded with OFDM signal in digital domain. Then each OFDM signal is up-sampled to M sample-per-symbol by inserting $M-1$ zeros between every two adjacent original OFDM data samples, and subsequently fed into a corresponding shaping filter (SF). After that, all of the shaped signals are aggregated/multiplexed together in the digital domain. In order to generate the analog signal, a single digital-to-analogue converter (DAC) is employed. Then, the produced analog

signal is fed into the electro-absorption modulator (EAM) to perform the intensity modulation. To obtain high power budget, the optical signal is boost by an erbium-doped fiber amplifier (EDFA) before being launched into the fiber link. At the optical distribution network (ODN), the received optical data stream is split into N parallel parts by a power splitter and each part is delivered to an ONU. At ONU side, the received optical signal is detected via a single-end PD and digitalized by an analog-to-digital converter (ADC). After that, the received signal is digitally filtered by a corresponding matching filter to extract/obtain the desired OFDM signal and is then down-sampled by a factor of M ($M \downarrow$). Finally, the conventional OFDM demodulation process is performed to obtain the transmitted data sequence.

Fig. 1. Principle of Digital Orthogonal Filter Multiple Access PONs.

In the OLT digital signal processor (DSP), the required digital filters for digital orthogonal multiple filter access PON system is construct by the Hilbert-pair approach [19], the k -th orthogonal digital filter pair is given by,

$$h_k^w(t) = \begin{cases} \frac{\sin\left[\pi(1-\beta)\frac{t}{T}\right] + 4\beta\frac{t}{T}\cos\left[\pi(1+\beta)\frac{t}{T}\right]}{\pi\frac{t}{T}\left[1 - \left(4\beta\frac{t}{T}\right)^2\right]} \cos(2\pi f_{ck}t) & w = I \\ \frac{\sin\left[\pi(1-\beta)\frac{t}{T}\right] + 4\beta\frac{t}{T}\cos\left[\pi(1+\beta)\frac{t}{T}\right]}{\pi\frac{t}{T}\left[1 - \left(4\beta\frac{t}{T}\right)^2\right]} \sin(2\pi f_{ck}t) & w = Q \end{cases} \quad (1)$$

From the mathematical point of view, the orthogonal digital filter is constructed by multiplying a baseband pulse having a square-root raised-cosine (SRRRC) form with a cosine and/or sine function. The $\{h_k^I(t) \text{ and } h_k^Q(t)\}$ represent the digital filter type, i.e. in-phase digital filter and quadrature-phase digital filter. And f_{ck} is the central frequency of the k -th digital filtering-enabled spectrally-overlapped sub-bands, T is the symbol duration and β is the roll-off factor. The impacts of digital filtering and its optimum digital filter coefficient setting can be found in [19]. Then the shaped multi-band signals for the ONUs are combined together.

B. Theoretical Model Based On EML

After DAC, the intensity modulation is implemented, as demonstrated in Fig. 1, EML is used to perform the required electrical-to-optical (E-O) conversion. To further evaluate the EML nonlinearity-induced impact on the power budget of orthogonal digital filter multiple access PONs system, the chirp of the EML should be considered and the output of the EML is written as [25],

$$E = \sqrt{P_b(1+\chi_n)} e^{-1/2 j\alpha \ln(1+\chi_n)} \quad (2)$$

where P_b is the bias power, χ_n denotes the normalized signal component and α represents the chirp of the EML. After E-O conversion, the optical signal is then launched into the

standard single mode fiber (SSMF) with a sufficient optical power. The SPM induced the phase change of Θ_{spm} is expressed as [26],

$$\Theta_{spm} = \gamma L_{eff} P_t = \gamma L_{eff} P_b (1 + \chi_n) = \psi_{NL} (1 + \chi_n) \quad (3)$$

where $\psi_{NL} = \gamma L_{eff} P$, γ is the nonlinear coefficient of the fiber, and P_t is the optical launch power. L_{eff} is the effective fiber length. When taking both EML chirp and SPM effect (B-EMLCSPME) into consideration, we can obtain,

$$\begin{aligned} E_s &= \sqrt{P_b (1 + \chi_n)} e^{-1/2 j \alpha \ln(1 + \chi_n)} e^{j \Theta_{spm}} \\ &= \sqrt{P_b (1 + \chi_n)} e^{-1/2 j \alpha \ln(1 + \chi_n)} e^{j \psi_{NL}} e^{j \psi_{NL} \chi_n} \end{aligned} \quad (4)$$

After expanding (4) into Taylor series, the 2nd-order approximation of the B-EMLCSPME can be written as,

$$\begin{aligned} E_s &= \sqrt{P_b} e^{j \psi_{NL}} (1 + \chi_n)^{\frac{1}{2}(1 - j \alpha)} e^{j \psi_{NL} \chi_n} \\ &\approx \sqrt{P_b} e^{j \psi_{NL}} \left[1 + \frac{1}{2} (1 - j \alpha) \chi_n - \frac{1}{8} (1 + \alpha^2) \chi_n^2 \right] [1 + j \psi_{NL} \chi_n \\ &\quad - \frac{1}{2} \psi_{NL}^2 \chi_n^2] \\ &= \sqrt{P_b} e^{j \psi_{NL}} \left\{ 1 + \left[\frac{1 + j(2\psi_{NL} - \alpha)}{2} \right] \chi_n \right. \\ &\quad \left. - \left[\frac{1 + \alpha^2 + 4\psi_{NL}^2 - 4\alpha\psi_{NL} - j4\psi_{NL}}{8} \right] \chi_n^2 \right\} \end{aligned} \quad (5)$$

Note that, only B-EMLCSPME is considered in (5). When taking the chromatic dispersion into account, the optical signal at the output of the fiber can be simply estimated as $E_{out} = E_s \otimes \tilde{h}(t)$, where the impulse response of the fiber $\tilde{h}(t)$ has a Fourier transform of $H(f) = e^{j\psi_{cd}}$. At the ONU, after square-law detection, the detected electrical signal is expressed as,

$$I(t) = [E_s(t) \otimes \tilde{h}(t)][E_s(t) \otimes \tilde{h}(t)]^* \quad (6)$$

Based on the above-stated theoretical analysis, the received electrical signal can be expanded as,

$$\begin{aligned} I(t) &= \sqrt{P_b} e^{j \psi_{NL}} \left\{ 1 + \left[\frac{1 + j(2\psi_{NL} - \alpha)}{2} \right] \chi_n \right. \\ &\quad \left. - \left[\frac{1 + \alpha^2 + 4\psi_{NL}^2 - 4\alpha\psi_{NL} - j4\psi_{NL}}{8} \right] \chi_n^2 \right\} \otimes \tilde{h}(t) \\ &\quad \times \sqrt{P_b} e^{-j \psi_{NL}} \left\{ 1 + \left[\frac{1 - j(2\psi_{NL} - \alpha)}{2} \right] \chi_n \right. \\ &\quad \left. - \left[\frac{1 + \alpha^2 + 4\psi_{NL}^2 - 4\alpha\psi_{NL} + j4\psi_{NL}}{8} \right] \chi_n^2 \right\} \otimes \tilde{h}^*(t) \\ &= \overset{dc}{P_b} \left\{ 1 + \overbrace{\left[\frac{1 + j(2\psi_{NL} - \alpha)}{2} \right] \chi_n \otimes \tilde{h}(t) + \left[\frac{1 - j(2\psi_{NL} - \alpha)}{2} \right] \chi_n^* \otimes \tilde{h}^*(t)}^{\text{desired signal}} \right. \\ &\quad \left. - \overbrace{\left[\frac{1 + \alpha^2 + 4\psi_{NL}^2 - 4\alpha\psi_{NL} - j4\psi_{NL}}{8} \right] \chi_n^2 \otimes \tilde{h}(t)}^{\text{higher-order interference}} \right. \\ &\quad \left. - \overbrace{\left[\frac{1 + \alpha^2 + 4\psi_{NL}^2 - 4\alpha\psi_{NL} + j4\psi_{NL}}{8} \right] \chi_n^{*2} \otimes \tilde{h}^*(t)}^{\text{higher-order interference}} \right\} \end{aligned} \quad (7)$$

It can be seen in (7) that the received electrical signal consists of a DC component, the desired signal and the high-order interferences. The high-order interferences can be neglected due to the signal $\chi_n \ll 1$. Also considering the fact that the signal χ_n is the real-valued, $I(t)$ can be further simplified as,

$$\begin{aligned} I(t) &= P_b \{ [1 + \chi_n \otimes \left[\frac{1 + j(2\psi_{NL} - \alpha)}{2} \tilde{h}(t) + \frac{1 - j(2\psi_{NL} - \alpha)}{2} \tilde{h}^*(t) \right]] \} \\ &= P_b [1 + \chi_n \otimes \tilde{h}_{eq}(t)] \end{aligned} \quad (8)$$

And the $\tilde{h}_{eq}(t)$ equals to,

$$\begin{aligned} \tilde{h}_{eq}(t) &= \left[\frac{1 + j\xi_{EMLCSPM}}{2} \tilde{h}(t) + \frac{1 - j\xi_{EMLCSPM}}{2} \tilde{h}^*(t) \right] \\ &= [1 + j\xi_{EMLCSPM}] \frac{\tilde{h}(t)}{2} + [1 - j\xi_{EMLCSPM}] \frac{\tilde{h}^*(t)}{2} \end{aligned} \quad (9)$$

where $\xi_{EMLCSPM} = 2\psi_{NL} - \alpha$, and the frequency expression of $\tilde{h}_{eq}(t)$ can be express as,

$$\begin{aligned} H_{eq}(f) &= [1 + j\xi_{EMLCSPM}] \frac{H(f)}{2} + [1 - j\xi_{EMLCSPM}] \frac{H^*(f)}{2} \\ &= \sqrt{1 + \xi_{EMLCSPM}^2} (e^{ja \arctan \xi_{EMLCSPM}} e^{j\Psi_{CD}} + e^{-ja \arctan \xi_{EMLCSPM}} e^{-j\Psi_{CD}}) / 2 \\ &= \sqrt{1 + \xi_{EMLCSPM}^2} \frac{\cos[\Psi_{CD} + \arctan \xi_{EMLCSPM}]}{J - EMLCSPMCD} \end{aligned} \quad (10)$$

From the Eq(10), we can see that the overall system channel frequency response are jointly influenced by three main physical mechanisms, namely, EML chirp, fiber SPM and chromatic dispersion (J-EMLCSPMCD).

Figure 2 plot the overall system frequency response of 25-km and 50-km SSMF with various EML chirp value and different optical signal launch power. As show in the Fig. 2, at the same chirp factor and launch power conditions, the frequency response over 50-km fiber link is worse than 25-km fiber link due to the CD effect. Moreover, the system frequency response is relatively poor when EML has no chirp and optical signal launch power is fixed at 8dBm. When increasing the optical signal launch power to the 12 dBm, the SPM effect is non-negligible, thereby enhancing the system frequency response. It is also interesting to find that for a specific optical signal launch power, the system frequency response is better with the negative chirp of EML; while for the positive chirp, the system frequency response is narrowed. It implies that the system frequency response improvement is achievable by making full use of high optical launch power induced SPM effect and EML chirp with an optimized negative value.

Fig. 2. Frequency response with different EML chirps and different launch power for (a) 25-km SSMF and (b) 50-km SSMF.

3. Experimental setup for the digital orthogonal filter access system

Figure 3 shows the experimental setup of the digital orthogonal filter multiple access PONs with 10G-class optics. In the OLT, the input pseudorandom binary sequence (PRBS) is mapped into 16-quaternary amplitude modulation (QAM) symbols. A 64-point inverse fast Fourier transform (IFFT) and the Hermitian symmetry are imposed to generate real-valued OFDM signals. Power loading is also utilized to maximize the system capacity, where all the subcarriers are adaptively loaded with three levels of amplitudes, i.e., the 1st-5th, 6th-28th, 29th-31st subcarriers are assigned with the normalized amplitudes of 0.92, 0.96 and 1, respectively. And a cyclic prefix (CP) with its length equal to 1/8 of an OFDM symbol duration is inserted to mitigate inter-symbol interference (ISI) which is caused by fiber dispersion. And clipping ratio is fixed at 13 dB to perform the digital clipping. Six independent real-valued OFDM signals which are intended for six ONUs are produced and subsequently pass through the shaping filters, which has a tap length of 64 and a roll-off factor of 0.01. After that, the six digitally-filtered OFDM sub-bands are aggregated together and the band of I_1, Q_1, I_2, Q_2, I_3 and Q_3 are intended from ONU₁ to ONU₆, respectively, as show in the inset of Fig. 3. All of the Tx DSP procedures are carried out in Matlab and the obtained digital signal is uploaded into an arbitrary waveform generator (AWG) running at 25-GS/s sampling rate and 8-bit resolution to produce analog signal with 25-Gb/s. The bandwidth of the AWG is ~14 GHz. The output signal of the AWG first passes through an amplifier and then a DC bias is added via a bias-tee. Following that, the electrical signal is used to drive the EML which consists of a 10-GHz EAM and a distributed feedback (DFB) laser at C band. The key parameters of the system are listed in the Table 1.

Fig. 3. Experimental setup of the digital orthogonal filtering access for the IM/DD PON system.

After E-O conversion, the launch optical power is adjusted via an EDFA and a variable optical attenuator (VOA) with its dynamic range from 0 to 16 dBm. Finally, the optical signal is launched into a SSMF and the fiber attenuation coefficient is ~0.2 dB/km. At the ONU side, the received optical signal first passes through a VOA to control the received optical power and then it is fed into a commercial PIN photodiode (PIN-PD) with an inline transimpedance amplifier (TIA) receiver (the 3-dB bandwidth is 10 GHz) to perform optical-to-electrical (O-E) conversion. Then the output electrical signal is sampled by a digital processing oscilloscope (DPO) operating at 100 GS/s and the resolution is 8-bit. The electric signal waveform of the output from the DPO is depicted in Fig. 4(a), in which the header used for timing synchronization has lower peak-to-peak voltage as the desired signal, so that it has higher tolerance to noise and distortions. The obtained digital signal is then processed offline by Matlab. The major receiver DSP procedures include timing synchronization, Volterra nonlinear equalization (VNLE), matching filter, down-sampling, real-valued OFDM signal recovery, demodulation, and bit error rate calculation. The 3rd-order VNLE [27] is adopted to compensate impairments of system and the Volterra filter lengths of the first term $N_1 = 21$ is used to mitigate the linear distortion, and second and third order terms are set to, $N_2 = 7$ and $N_3 = 5$, to mitigate the nonlinear distortion.

TABLE I
PARAMETERS OF THE EXPERIMENT SETUP

Parameter	Value
DAC sampling rate	25 GSa/s
DAC bit resolution	8 bits
ADC sampling rate	100 GSa/s
ADC bit resolution	8 bits
Roll-off of SRRC	0.01
IFFT/FFT points per channel	64
CP ratio	1/8
Clipping ratio	13 dB
EML laser operation wavelength	C band
EML modulation bandwidth	10 GHz
PIN detector bandwidth	10 GHz
PIN detector sensitivity	-19 dBm

4. Experimental Results and Discussions

Figure 4(b) presents the measured transfer curves of the EML at different bias voltages of the EAM for different laser driving currents. The results show that the linear operational bias is approximate -0.6 V and the output optical power is gradually decreased when reducing the bias voltage. Figs. 4(c) and (d) present the measured frequency responses with different launch powers and EML biases over 25-km and 50-km fiber link, respectively. The frequency responses have a similar trend to the simulation frequency response given in Fig. 2. Actually, different EML biases correspond to different chirp factors, and the chirp factor can be calculated according to the measured frequency responses [28]. When the EML is biased at -0.6 V, a chirp-free system can be obtained by employing either the frequency responses of 25-km or 50-km fiber link. In addition, a negative EML chirp can be achieved by adjusting the EML's bias voltage less than this chirp-free value, and decreasing the bias voltage can increase the negative value of the chirp parameter, but the nonlinear distortion caused by the EML is also increased which is same to the characteristic of other typical EMLs. On the contrary, a

Fig. 4. (a) Waveform with the header for timing synchronization, (b) EML output power vs. bias voltage of EAM for different laser driving currents, (c) frequency response over 25-km SMF, and (d) frequency response over 50-km SMF.

Fig. 5. The optical spectra with different biases of EML for (a) without and (b) with modulated signal. positive chirp can also be obtained by increasing the EML bias larger than this chirp-free value. Figs. 5(a) and (b) show the optical spectra at different biases of the EML without and with modulated signal, respectively. The laser driving current is fixed at 120 mA to achieve a high output optical power.

We first evaluate the transmission performance versus the bias of the EML and the optical launch power for the ONU₁. Figure 6 illustrates the contour of BER versus launch optical power and the modulator bias over 25-km fiber link. As we can see, $\log_{10}(\text{BER})$ is less than -4.5 in the red-color region, while in the green-color region, $\log_{10}(\text{BER})$ is larger than -4. The results in Fig. 6 reveal that 1) by using an optimum bias voltage at ~ -0.6 V and an optimum optical signal launch power at 6 dBm the transmission performance of the ONU₁ with optimized condition under the tradeoff between CD, SPM and EML's chirp; 2) relatively high optical signal launch power enhances the transmission performance, thereby improving optical power budget; 3) when optical signal launch power > 14 dBm, the fiber nonlinearity-induced system impairment becomes significant and degrades the BER performance; 4) when the EML's bias voltage is lower than the optimum value, the negative chirp can be obtained, but the EML's nonlinear distortion will be also increased, and hence the BER performance is decreased as shown in Fig. 6. Therefore, the transmission performance should have a tradeoff between EML's chirp, SPM and CD.

Fig. 6. BER vs. the bias of EML and launch optical power over 25-km SSF.

Fig. 7 shows the receiver sensitivity of band I₁ for ONU₁ and band Q₃ for ONU₆ at a BER level of 4.5×10^{-3} (hard-decision forward error correction (HD-FEC) threshold) for various the

optical launch power. The fiber transmission distance is 25-km and the EML bias is set at -0.6 V. It should be noting that among all considered six ONUs, the ONU₁ and ONU₆ have the best and worst transmission performances, respectively. Since ONU₆ occupies the highest frequency band, it suffers from the most severe CD effect. When the optical signal launch power is 14 dBm, a link power budget of 26 dB can be achieved for ONU₁. It is also shown that, with the same optical launch power, the receiver sensitivity differences between six ONUs are less than 1 dB.

Fig. 7. Receiver sensitivity as a function of the launch optical power for ODN reach of 25 km at bias of -0.6 V.

When extending the fiber transmission to 50 km, the chromatic dispersion-induced impairment is significant. Fig. 8(a) presents the measured BER of each ONU versus different EML bias voltages. It shows that 1) all ONUs have vary similar BER developing trend; 2) all ONUs have the same optimal EML bias voltage of -0.8 V. Under this optimal EML bias voltage, the receiver sensitivities of ONU₁ and ONU₆ are explored and depicted as a function of the total launch optical power, as shown in Fig. 8(b). The receiver sensitivity differences among all ONUs are still less than 1 dB for a specific optical signal launch power. And 25-dB power budget is obtained for ONU₁ when employing 12-dBm optical signal launch power. The result also indicates that J-EMLCSPMCD effect can ensure digital orthogonal filter multiple access PON systems have excellent power budget, which is independent of sub-band spectral locations. Most importantly, many researches utilize the avalanche photodiode (APD) with higher sensitivity to get higher power budget. Due to the lack of APD, we used can still get relatively higher power budget with PIN.

Fig.8. (a) Measured BER versus the bias of EML for each ONU after 50-km SSMF transmission (the launch optical power and received optical power are fixed at 6 dBm and -10 dBm, respectively), and (b) receiver sensitivity as a function of the launch optical power for ODN reach of 50 km with bias of -0.8 V.

Finally, we investigated the combined effect of ADC jitter and resolution on the power budget of ONU₆, which has the worst transmission performances. The fiber transmission distance is 25km and the EML is biased at -0.6 V. The results are depicted in Fig. 9. The optical signal launch power is 14 dBm, the system has a power budget of 25 dB for 8-bit resolution DPO. After re-quantifying the sampled data from the 8-bit DPO into 3~7 bits, corresponding to $2^3 \sim 2^7$ discrete quantization levels. 25 dB power budget can be still obtained when the resolution of ADC is 7 bit, and the tolerance of ADC jitter is ± 1.25 ps. The certain power budget can be achieved within a region of ADC jitter and resolution, where power budget of 24 dB in the red-color region; while in the yellow-color region, the power budget is 23 dB. For a desired power budget of 24 dB, the tolerable of ADC resolution can be decreased from 7 to 5 bits and the ADC jitter can be varying within $[\pm 1.25, \pm 3.75]$ ps. When the ADC jitter varies within $[\pm 5, \pm 6.25]$ ps and the DAC resolution is gradually increased from 3 to 8 bits, the power budget is increased from 18 to 22 dB. To achieve a higher power budget, the ADC jitter and resolution should be controlled strictly.

Fig. 9. Power budget as a function of ADC jitter and resolution for ONU₆.

5. Conclusion

We have experimentally demonstrated a digital orthogonal filter multiple access-based IMDD PON system using 10G optical/electrical devices. A theoretical model has been built to reveal the impacts of the SPM, chromatic dispersion and EML chirp. The experimental results have shown that high power budget can be effectively achieved by utilizing a relatively high launch power to enhance the SPM effects and adjusting EML bias voltage to achieve negative EAM chirp. In our demonstration, 26-dB power budget is achieved in 25-Gb/s digital orthogonal filter multiple access IM/DD PONs system at C-band after 25-km transmission. The proposed scheme in this work has a potential application in future cost-effective optical access networks.

6. Acknowledgment

This work was supported partly by the Sichuan Science and Technology Program under Grant 2018JY0145, National Science Foundation of China under Grant 61571092, Joint Research Program of UESTC and Salton Tech. under Grant 2018STKY001, Zhongshan Government Funded project for National Laboratory Zhongshan Branch Lab under Grant 412S06, Research program of UESTC Zhongshan Institute.

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