Fast and robust chromatic dispersion estimation based on temporal auto-correlation after digital spectrum superposition

Fast and robust chromatic dispersion estimation based on temporal auto-correlation after digital spectrum superposition

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Abstract: We investigate and experimentally demonstrate a fast and robust chromatic dispersion (CD) estimation method based on temporal auto-correlation after digital spectrum superposition. The estimation process is fast, because neither tentative CD scanning based on CD compensation nor specific cost function calculations are used. Meanwhile, the proposed CD estimation method is robust against polarization mode dispersion (PMD), amplified spontaneous emission (ASE) noise and fiber nonlinearity. Furthermore, the proposed CD estimation method can be used for various modulation formats and digital pulse shaping technique. Only 4096 samples are necessary for CD estimation of single carrier either 112 Gbps DP-QPSK or 224 Gbps DP-16QAM signal with various pulse shapes. 8192 samples are sufficient for the root-raised-cosine pulse with roll-off factor of 0.1. As low as 50 ps/nm standard deviation together with a worst estimation error of about 160 ps/nm is experimentally obtained for 7×112 Gbps DP-QPSK WDM signal after the transmission through 480 km to 9120 km single mode fiber (SMF) loop using different launch powers.

OCIS codes: (060.2330) Fiber optics communications; (060.1660) Coherent communications; (060.4510) Optical communications; (260.2030) Dispersion.

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1. Introduction

The digital coherent receiver is central in current optical transmission system due to its capability of compensating various transmission impairments. It has been reported that accumulated CD at the range of 10^4~10^5 ps/nm can be efficiently compensated in the digital coherent receiver [1–3]. While the CD compensation is typically realized by a static equalizer, an accurate information of the accumulated CD is necessary to perform the precise channel equalization. Generally speaking, the value of the accumulated CD can be treated as a fixed parameter once the fiber link is designated. While the optical network is sometimes under dynamic reconfiguration, the routing path may vary from time to time and consequently the accumulated CD varies. Therefore, in order to realize digital compensation of transmission impairment, a fast and precise CD monitoring is highly desired

Plenty of CD estimation methods have been demonstrated until now. By analyzing the impulse response of the adaptive equalizer, it is possible to quickly capture the CD information at a range of 0-1500 ps/nm depending on the number of taps [4]. Other CD estimation methods usually rely on either tentative CD compensation or tentative cost function calculations by scanning the CD over a specific range [5–9]. In the presence of a large amount of residual CD exceeding 10^4 ps/nm, hundred thousands of samples may be required to cover the CD scanning range. Thus the processing time for CD estimation is fairly long for those schemes. In order to speed up the CD estimation, a modified Godard clock tone algorithm was proposed [10]. After the clock tone is constructed from the complex auto-correlation function of the received signal spectrum, fast Fourier transform (FFT) is used to scan the CD. The scanning range is proportional to the number of samples used for the FFT and the resolution is inversely proportional to the sampling rate. Recently, the auto-correlation of signal power waveform (ACSPW) CD estimator was proposed for fast CD estimation [11–

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13]. While no tentative CD scanning is required for the ACSPW CD estimator, 16384 samples can be used for the estimation of accumulated CD at a range of 0~10^5 ps/nm [11].

In this paper, we extend our previous work on the spectrum superposition based CD estimation method [14]. Similar to the ACSPW method, the proposed CD estimation requires no training sequences and can operate without tentative CD scanning. Furthermore, compared to the ACSPW method, it has been verified that by introducing a matched digital equalizer during the CD estimation, a higher peak-to-average-ratio (PAR) of the searching plot for our proposed CD estimation method can be achieved and a 4 times reduction of required number of samples is verified for reliable CD estimation. Therefore, the proposed CD estimation method is applicable for fast and blind CD estimation. In particular, various high-order modulation formats and pulse shaping techniques are employed to increase the spectral efficiency (SE). Therefore, the robustness of the CD estimation against fiber transmission impairments, including polarization mode dispersion (PMD), ASE noise and nonlinearity, is explored under the scenarios of various modulation formats. In this paper, the performance of our proposed CD estimation for different modulation formats with various pulse shapes is discussed under ASE noise and PMD with a mean DGD of 25 ps. Through optimizing the digital filters used for the CD estimation, the proposed CD estimation method can operate well for both return to zero (RZ) pulses with different duty cycles and Nyquist pulses with root raised-cosine pulse shaping. For 112 Gbps DP-QPSK and 224 Gbps DP-16QAM signal with RZ pulses, the standard deviation is 74 ps/nm and the worst CD estimation error is 210 ps/nm using only 4096 samples. Meanwhile, at the expense of aggressive pulse shaping, 8192 samples are required for the CD estimation of root-raised-cosine pulse with roll-off factor as low as 0.1. The standard deviation of the estimation error is 138 ps/nm and the worst CD estimation error is 375 ps/nm. We finally verify the robustness of our proposed CD estimation method against fiber nonlinearity and ASE noise through a 7×112 Gbps DP-QPSK WDM transmission experiment. After 480 km to 9120 km SMF loop transmission, the standard deviation of the CD estimation error is around 50 ps/nm, even when the 7×112 Gbps DP-QPSK signals operate in the nonlinear region.

2. Operation principle

To simplify the discussion, we first consider a single polarization coherent optical fiber transmission system. Assume $S_T(t)$ is the transmitted baseband signal, then after a certain fiber link with a channel impulse response function of $h(t)$, the coherently detected signal can be written as

$$\begin{aligned} S_R(t) &= S_T(t) * h(t)e^{\mathrm{j}(2\pi\Delta f t+\varphi(t))} + \mathrm{u}(t) \\ &= \left(\sum_{k=1}^{K} a_{\mathrm{k}} g(t-kT)\right) * h(t)e^{\mathrm{j}(2\pi\Delta f t+\varphi(t))} + u(t)\,, \end{aligned} \tag{1}$$

where $*$ stands for convolution, a_{k} is the complex information symbol, $g(t)$ is the pulse shape, T is the symbol period, Δf represents the frequency offset between transmitter laser and local oscillator, $\varphi(t)$ denotes the combined phase noise of both lasers, and $u(t)$ is noise. To analytically focus on the chromatic dispersion (CD) estimation, we ignore the effect of PMD, fiber nonlinearity, frequency offset and phase noise here. Suppose the coherently detected signal is sampled at a sampling rate of R_{s}, then the spectrum of received samples can be written as

$$\begin{aligned} S_R(n) &= S_T(n)H(n) + U(n) \\ &= S_T(n)e^{(2j\pi^2\beta_2 L(nR_s/N)^2)} + U(n)\,, \end{aligned} \tag{2}$$

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where L is the link length of fiber transmission, β_2 is the group velocity dispersion coefficient, and N is the block size of the FFT processing (i.e., the number of samples used for the CD estimation), $-N/2+1 \le n \le N/2$. As in most coherent receivers, R_s is assumed to be twice of the symbol rate. Note that for $H(n)$, the quadratic phase term can be transferred to first-order amplitude term by introducing a digital superposition of the upper and lower sideband component as

$$\begin{aligned} |P(n)|^2 &= |H(n+N/2)+H(n)|^2 \\ &= 2+2\cos(2\pi^2\beta_2 L R_s^2 (n/N+1/4)), \\ &= 2+2\cos(2\pi^2\beta_2 L R_s^2 (m/N)), \end{aligned} \tag{3}$$

where $P(n)$ is the result of spectrum superposition, $-N/2+1 \le n \le 0$, $m{=}n+N/4$ and $-N/4+1 \le m \le N/4$. Meanwhile, the first-order amplitude term can be easily detected by performing the inverse fast Fourier transform (IFFT) as

$$\left|\mathrm{IFFT}(|P(n)|^2)\right| = 2\delta(\tau)+\delta(\tau+\tau_0)+\delta(\tau-\tau_0), \tag{4}$$

where $\tau = 2\mathrm{m}/R_s$, and the peak located at $\tau_0 = \pi\beta_2 L R_s$ indicates an accumulated CD estimate according to

$$CD_{est} = -\frac{2\pi c\beta_2 L}{\lambda^2} = -\frac{2c\tau_0}{\lambda^2 R_s}. \tag{5}$$

The result in Eq. (3)-(5) suggests an accumulated CD estimation by doing the digital superposition of received signal spectrum $S_R(n)$. However, before doing the digital superposition of $S_R(n)$, the spectral properties of $S_T(n)$ need to be taken into account. Since

$$\begin{aligned} S_T(n) &= \sum_{l=0}^{N-1} e^{-j\frac{2\pi nl}{N}} \left(\sum_{k=1}^{K} a_{\mathrm{k}} g(lT_s - kT) \right) \\ &= \sum_{k=1}^{K} a_{\mathrm{k}} \sum_{l=0}^{N-1} e^{-j\frac{2\pi nl}{N}} g(lT_s - 2kT_s) \\ &= \sum_{k=1}^{K} a_{\mathrm{k}} e^{-j\frac{4\pi nk}{N}} G(n) \\ &= G(n) \sum_{k=1}^{K} a_{\mathrm{k}} e^{-j\frac{4\pi nk}{N}}, \end{aligned} \tag{6}$$

where K is the number of symbols and $G(.)$ is the FFT of $g(.)$, i.e., the frequency response of the pulse shape. Thus

$$\begin{aligned} S_T(n+N/2) &= G(n+N/2) \sum_{k=1}^{K} a_k e^{-j\frac{4\pi(n+N/2)k}{N}} \\ &= G(n+N/2) \sum_{k=1}^{K} a_k e^{-j\frac{4\pi nk}{N}}, \end{aligned} \tag{7}$$

Ignoring the noise term, the superimposed spectrum becomes

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$$S_R(n)+S_R(n+N/2)\approx\left(G(n)H(n)+G(n+N/2)H(n+N/2)\right)\sum_{k=1}^{K}a_k e^{-j\frac{4\pi nk}{N}}. \quad (8)$$

In order to obtain first-order amplitude term of CD as shown in Eq. (3), we introduce a digital filter $\mathrm{F}(n)$ with tap weights satisfying

$$G(n+N/2)F(n+N/2)=G(n)F(n)=A(n). \quad (9)$$

Then

$$\begin{aligned}&\left|S_R(n)F(n)+S_R(n+N/2)F(n+N/2)\right|^2\\&=\left|A(n)\sum_{k=1}^{K}a_k e^{-j\frac{4\pi nk}{N}}\right|^2(2+2\cos(2\pi^2\beta_2 LR_s^2(n/N+1/4))).\end{aligned} \quad (10)$$

Finally, the CD can be estimated from

$$\begin{aligned}R(\tau)&=\left|\mathrm{IFFT}(\left|S_R(n)F(n)+S_R(n+N/2)F(n+N/2)\right|^2)\right|\\&=\left|\left(2\delta(\tau)+\delta(\tau+\tau_0)+\delta(\tau-\tau_0)\right)*\mathrm{IFFT}\left(\left|A(n)\sum_{k=1}^{K}a_k e^{-j\frac{4\pi nk}{N}}\right|^2\right)\right|.\end{aligned} \quad (11)$$

In Eq. (11), the operation of $\mathrm{IFFT}(|\ |^2)$ represents the temporal auto-correlation after digital spectrum superposition. The auto-correlation result is derived with a peak whose temporal position indicates the accumulated CD value. Therefore, the accumulated CD can be estimated by searching the temporal position of peak. The calculation process is similar to the previously reported ACSPW CD estimation method, because both do auto-correlation of either signal electrical field or signal power in the time domain. Unlike the CD estimation in time domain, the CD estimation based on Godard clock tone can be realized with cross-correlation of the upper and lower-sideband signal in frequency domain. Since the magnitude of clock tone is decreased with the growing CD [15], after a tentative CD compensation by scanning the CD over a specific range, the accumulated CD value can be determined with maximum CT magnitude. The simplification of such CD estimation method is put forward, by doing a FFT calculation in frequency domain in order to realize the tentative CD compensation and the CT magnitude sweeping simultaneously [10]. Similar computation complexity can be obtained for both CD estimation methods, except that an additional digital filter $F(n)$ and a superposition operation of the upper- and lower-sideband signal spectrum are necessary for our proposed CD estimation method.

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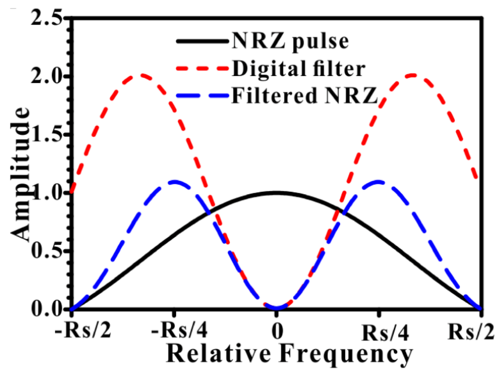

Fig. 1. Spectrum of NRZ pulse, frequency response of digital filter, and spectrum of the filtered pulse.

Consider a rectangular non-return to zero (NRZ) pulse. The digital spectrum of the rectangular NRZ pulse is

$$G(n) = \sin c(n/N). \tag{12}$$

We then choose the digital filter in Eq. (9) as

$$F(n) = 1 - \cos(3\pi n/N). \tag{13}$$

As shown in Fig. 1, the filtered spectrum of the pulse acts with an approximately symmetric property which is suitable for the spectrum superposition.

An exemplary plot of the CD estimation peaks for our CD estimation for single carrier 112Gbps DP-QPSK signal is presented in Fig. 2(a). In the simulation, complex valued sequences $S_T(t) = \{S_{TX}(t), S_{TY}(t)\}^T$ are transmitted through the fiber link and detected as $S_R(t) = \{ S_{RX}(t), S_{RY}(t) \}^{\mathrm{T}}$ at the receiver side. The subscripts X and Y represent two orthogonal polarizations. Both $S_{RX}(t)$ and $S_{RY}(t)$ can be resampled and transformed into frequency domain as $S_{RX}(n)$ and $S_{RY}(n)$ for the preparation of the CD estimation. Considering polarization effects, for each accumulated CD, four independent CD estimations can be conducted for $S_{RX}(n)$, $\mathrm{S_{RY}}(n)$ and $S_{RX}(n) \pm S_{RY}(n)$. The four combinations of the digital samples from different polarizations relate to four polarization filters with polarization angle of 0, π, π/4 and 3π/4. By applying the four polarization filters, all polarization effects can be effectively covered even in the presence of high order polarization [5]. The final CD estimation value is determined by a temporal peak with a higher PAR, which is defined as the ratio between the peak amplitude and the average noise floor, as shown in Fig. 2(b). Note that in Eq. (4), both positive and negative accumulated CDs cause two equal-amplitude peaks for $R(\tau)$ which make it difficult to determine the sign of the CD for a dispersion-managed fiber link. However, the problem can be solved by adding a positive CD value and then monitor the direction where the peaks shift.

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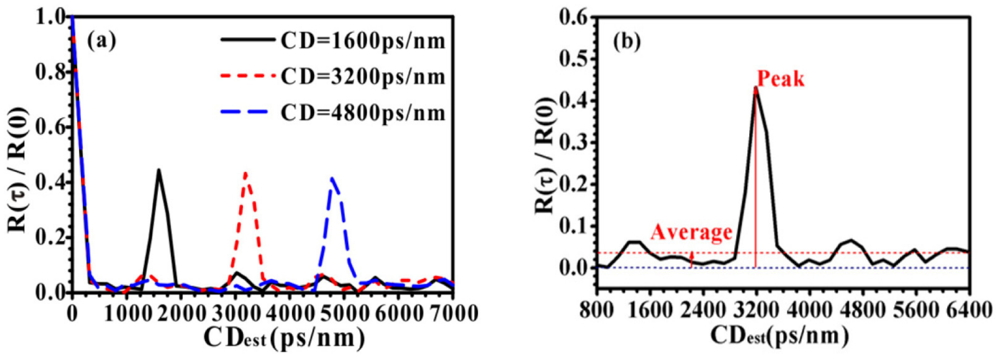

Fig. 2. CD estimation peaks for the 112Gbps NRZ DP-QPSK signal with (a) 1600, 3200 and 4800ps/nm accumulated CD, (b) 3200ps/nm accumulated CD using proposed method. N = 4096, frequency offset = 0GHz, optical signal-to-noise ratio (OSNR) = 20dB.

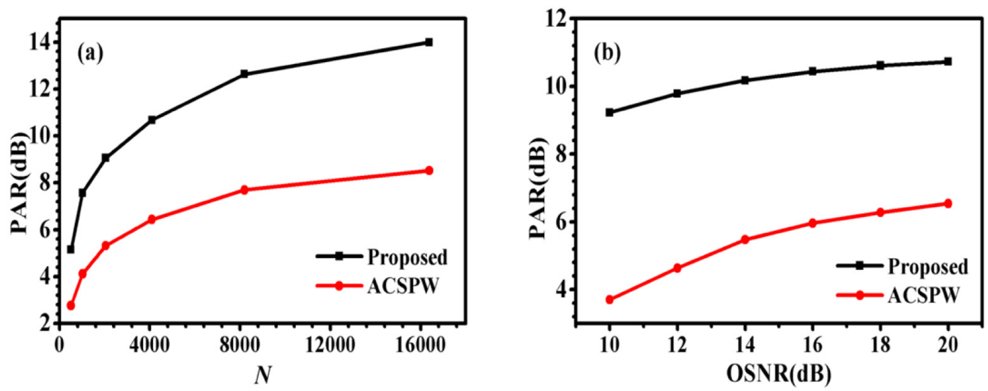

Fig. 3. PAR of CD estimation peak for 112 Gbps NRZ DP-QPSK signal with 1600ps/nm accumulated CD using the proposed CD estimation method and the ACSPW method with (a) different FFT block sizes when OSNR = 20 dB, (b) different OSNRs when N = 4096.

High PAR is of great significance for the robust CD estimation. With a low PAR, the estimation peak might be difficult to detect and leads to a CD estimation failure. Since both the proposed CD estimation and the ACSWP method are based on the temporal auto-correlation of received signal and exhibit with similar CD estimation peak in the time domain, here we carry out performance comparison. In Fig. 3(a), we compare the PARs of our proposed digital spectrum superposition method with the ACSPW method under various FFT block sizes during the calculation of searching peaks. It shows that for both CD estimation methods, the PARs rapidly increase with the FFT block size, because a large FFT block size is helpful to provide a precise transition of the quadratic phase shift component and reduce the noise perturbations. For our proposed method, the PAR is always higher than that of the ACSPW method. Thus the reduction of FFT block size is acceptable for our proposed method without sacrificing the estimation accuracy. Furthermore, in Fig. 3(b), decreasing OSNR, the PARs of our proposed estimation method yields a slightly slower decreasing speed than that of the ACSPW. Therefore, our proposed method is more robust to the ASE noise.

3. Simulation results and discussions

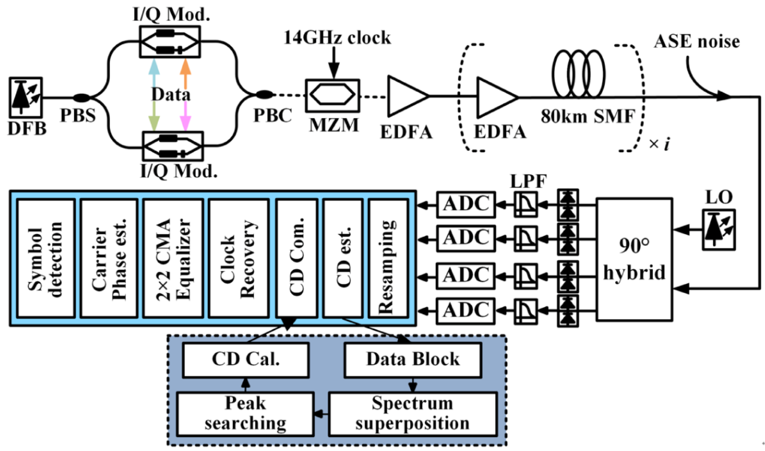

Fig. 4. Simulation setup for single carrier signal SMF transmission.

Based on VPI Transmission Maker 9.0, we investigated the proposed CD estimation method for a single carrier 112 Gbps DP-QPSK and 224 Gbps DP-16QAM signal. Figure 4 shows the simulation setup for the single carrier signal transmission system. The transmitter comprises a distributed feedback laser (DFB) with 1 MHz of linewidth and two I/Q modulators, each driven by two de-correlated binary or 4-PAM signals at 28 Gbaud to generate the QPSK or 16QAM optical signals, respectively, in each polarization. After the polarization combination, an optional Mach-Zehnder modulator (MZM) driven by a 14 GHz clock source is used for pulse carver. By carefully adjusting the bias and driving voltage, 66% duty cycle carved carrier suppressed return to zero (CSRZ) and 33% duty cycle carved RZ pulses can be obtained. The optical signal is then pre-amplified and launched to the fiber link which is composed by several spans of 80 km SMF with erbium-doped fiber amplifier (EDFA) only. The CD and nonlinear coefficient of the fiber is set to 16 ps/km/nm and 2.6 $W^{-1}\cdot km^{-1}$, respectively, while the mean DGD is set to 25 ps in all scenarios. The receiver OSNR is controlled by adding ASE noise. The receiver uses another laser with a linewidth of 1 MHz as the local oscillator (LO). The frequency offset between the LO and the signal source randomly varies from –2 GHz to 2 GHz. Using a polarization-diverse 90° hybrid and four balanced photodiodes, signals are coherently detected and then filtered by a 5th order Bessel electrical filter with 19 GHz bandwidth to remove the out-of-band noise. Before the CD estimation starts, the filtered signal is sampled at twice symbol rate. A data block with length of 4096 is used for the blind CD estimation. For the CD estimation, the spectrum of the received signal is first superimposed with the help of the introduced digital filter to obtain the CD estimation peak. Then the CD estimation peak is detected and the accumulated CD is calculated according to Eq. (5). The estimated CD value is used for static dispersion compensation subsequently. Finally, clock recovery, polarization demultiplexing and carrier phase recovery is carried out, and the symbol is detected.

To analyze the robustness of the proposed CD estimation method, a certain number of CD estimates can be carried out for the estimation statistics. Suppose V is the number of CD estimation trails in total, the statistical property of CD estimation results is then characterized by calculating the standard deviation of the estimation errors, which is defined as

$$\sigma=\sqrt{\frac{1}{V}\sum_{v=1}^{V}\left(CD_{real,v}-CD_{est,v}-\frac{1}{V}\sum_{v=1}^{V}(CD_{real,v}-CD_{est,v})\right)^{2}} \quad . \tag{16}$$

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3.1 CD estimation for NRZ and RZ signals

NRZ and RZ especially CSRZ are three most common modulation formats used in current fiber-optical communication systems. The exact expression of the spectrum for the NRZ and RZ pulses is presented in [16]. Compared to NRZ pulses, 33% RZ pulses and 67% CSRZ pulses exhibit broader spectrums. For the preparation of the CD estimation, we apply the digital filter $F(n)$ described in Eq. (13) for all three pulses. As stated before, the digital filter $F(n)$ is introduced to keep the filtered spectrum symmetric. Here we define

$$D(n){=}|(Gn+N/2)F(n+N/2)\text{-}G(n)F(n)|, \tag{17}$$

$$\overline{D}{=}\frac{1}{N}\sum_{n=1}^{N}D(n). \tag{18}$$

A good symmetric property of the filtered spectrum is obtained when $\overline{\mathrm{D}}$ is small. As shown in Fig. 5(a), for CSRZ and RZ pulses, after the digital filtering, $\overline{\mathrm{D}}$ is slightly higher than for NRZ pulses. Nevertheless, it is low enough for the proposed CD estimation. By applying the digital filter, we plot the corresponding CD estimation peaks for 112 Gbps DP-QPSK signal using NRZ, CSRZ and RZ pulse shapes in Fig. 5(b). A similar peak value and noise floor level is observed for all three plots. It indicates that the proposed CD estimation method has similar performance on NRZ, CSRZ and RZ signals.

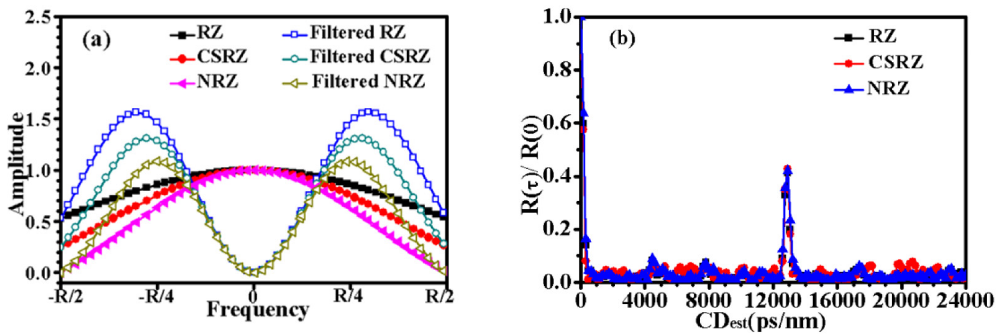

Fig. 5. (a) Spectra of NRZ/CSRZ /RZ pulses with digital filtering. (b) CD estimation peaks for 112Gbps NRZ/CSRZ /RZ DP-QPSK signals using proposed method on the condition of 12800 ps/nm CD and 20 dB OSNR

We further measure the CD estimation errors for QPSK and 16QAM signal with NRZ, CSRZ and RZ pulse shapes after 80 km to 1840 km transmission. For each combination of modulation formats and pulse shapes, 9200 independent trials are conducted to calculate the CD estimation error. The distributions of the CD estimation errors are shown in Fig. 6. For both QPSK and 16QAM signal, regardless of the pulse shapes, the standard deviation is all around 74 ps/nm. The detected worst estimation error is 210 ps/nm.

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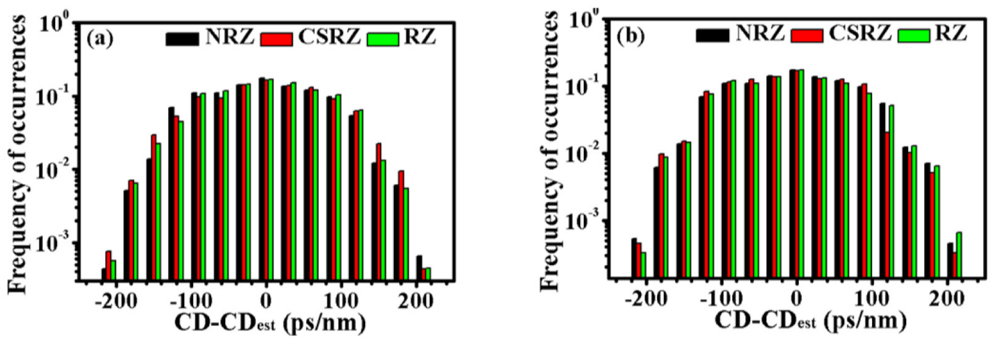

Fig. 6. Histogram of the CD estimation error for the (a) 112 Gbps DP-QPSK, (b) 224 Gbps DP-16QAM signal with NRZ/CSRZ/RZ pulse shapes after 80 km to 1840 km transmission. OSNR = 12dB, mean DGD = 25ps, $N = 4096$.

3.2 CD estimation for root-raised-cosine pulse shaped signals

During Nyquist WDM transmission, in order to achieve inter-symbol-interference (ISI) free transmission, the transmitted signal is usually shaped with a root-raised-cosine spectrum where low value of roll-off factor α is helpful to avoid extra interference from adjacent channels. In most Nyquist WDM systems, a roll-off factor of 0.1 is realistic [17]. For such root raised-cosine pulses, in order to keep the symmetric property of the signal spectrum for digital spectrum superposition, we choose the digital filter the same as the one described in [11]. As shown in Fig. 7(a), after digital filtering, some of the spectral components are lost. However, the filtered response of the upper and lower sideband is similar. Thus, they can be used for the proposed CD estimation. With the roll-off factor varying from 0.05 to 0.5, individual CD estimation peaks are shown in Fig. 7(b). When the roll-off factor is decreased, due to the limited spectral components for digital spectrum superposition, the CD estimation peak broadens and the estimation accuracy is reduced. In Fig. 8, we show the CD estimation errors for the root raised-cosine pulse signal in a 7 × 112 Gbps NRZ DP-QPSK WDM transmission with a channel spacing of 50GHz. In the simulation, for each in-phase/quadrature tributaries in two polarizations, an electrical root raised-cosine filter is applied at the transmitter side. After 800 to 1600 km of SMF transmission, a data block with 8192 samples is used for the CD estimation. To investigate the performance of the proposed CD estimation method, 100 independent trials are conducted at each transmission distance. From the histogram of Fig. 8, when the roll-off factor of the root raised-cosine filter is decreased from 0.5 to 0.1, the distribution range of the CD estimation error spreads. When the roll-off factor is equal to 0.5, the standard deviation of estimation error is 46 ps/nm and the detected worst estimation error is 175 ps/nm. Meanwhile, when the roll-off factor is further decreased to 0.1, the standard deviation of the estimation error is increased to 138 ps/nm and the detected worst estimation error is 375 ps/nm

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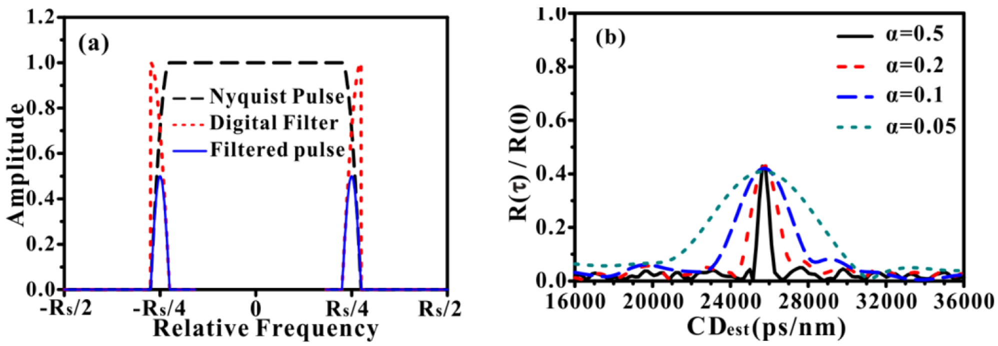

Fig. 7. (a) Frequency response of the root raised-cosine pulse shape (roll-off factor is equal to 0.1). (b) CD estimation peak for the 112Gbps DP-QPSK signal with a root raised-cosine pulse shape whose roll-off factor varies from 0.05 to 0.5. Accumulated CD = 25600 ps/nm, OSNR = 20dB, $N = 8192$.

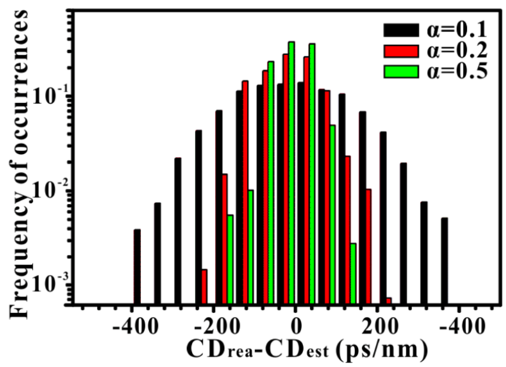

Fig. 8. Histogram of the CD estimation error for RRC pulse signal in a 7× 112 Gbps NRZ DP-QPSK WDM system after 800km-1600km SFM loop link, OSNR = 12dB, $N = 8192$.

3. Experimental results and discussions

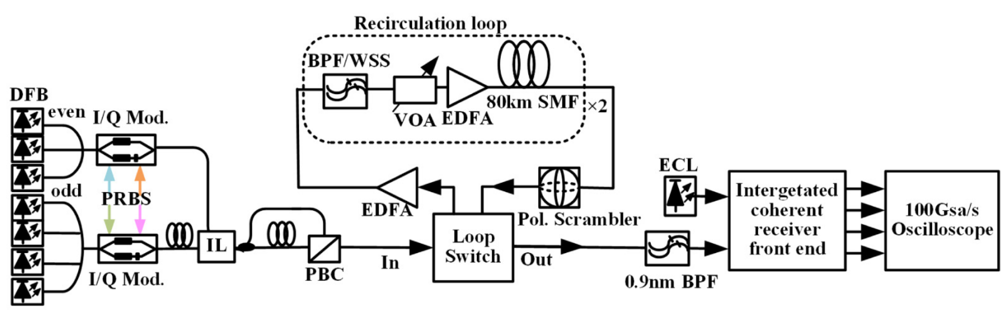

Fig. 9. Experimental setup for 7× 112 Gbps NRZ DP-QPSK WDM transmission.

The fiber optical transmission experiment of 7× 112 Gbps NRZ DP-QPSK WDM signal over 50 GHz grid is also conducted to investigate the performance of the proposed CD estimation method. As shown in Fig. 9, seven DFB lasers with linewidth around 100 kHz were used as optical sources. For QPSK modulation, lasers from odd and even channels were combined and injected to two I/Q modulators, respectively, which were driven by four de-correlated 28 Gbaud pseudorandom binary sequences (PRBS) with length of $2^{15}– 1$. The odd and even channels were further de-correlated by adding an optical time delay to the odd channels. After being combined by an optical interleaver, the entire QPSK WDM signal was polarization

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multiplexed with an additional optical delay on one polarization. The 7× 112 Gbps DP-QPSK WDM signal was then sent to the recirculation loop, which contained two spans of 80 km SMF. Before the first span of SMF, a 4 nm band-pass filter was used to suppress the out of band noise and then a waveshaper was used for both the power equalization of each channel and the suppression of noise before the second span. The launch power at each span of SMF was controlled by two variable attenuators. Besides, a polarization scrambler was also used to makes random polarization rotation between each loop round-trip. Before coherent detection at the receiver side, the WDM signal was first pre-amplified and then filtered by a 0.9 nm BPF to suppress out-of-band noise and interference from adjacent WDM channels. Using an external cavity laser with narrow linewidth as the local oscillator, the DP-QPSK signal from the central channel was detected by an integrated coherent receiver, which was composed of a polarization-diverse hybrid and four balanced photodiodes. A real-time oscilloscope with a sampling rate of 100 GSa/s was then used to collect the received signals. Finally, the received signal was resampled at 2 samples per symbol for the blind CD estimation.

To evaluate the performance of the proposed CD estimation method, the received signal after 3 to 57 fiber loop round-trips are saved for CD estimation. Figure 10(a) shows the estimation peaks using the proposed method after 1440 km, 1920 km and 2400 km fiber link

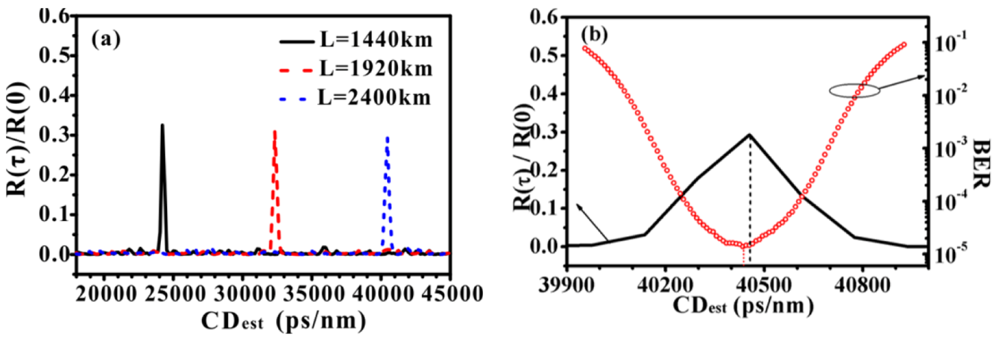

Fig. 10. (a) Experimental CD estimation peaks for 7× 112 Gbps NRZ DP-QPSK WDM signal after 1400 km, 1920 km and 2400 km SMF link. (b) Accumulated CD characterization through BER calculation after 2400 km SMF link and the tap number of 2 × 2 CMA equalizer in the DSP module is 3. The launch power is −2 dBm,N = 4096.

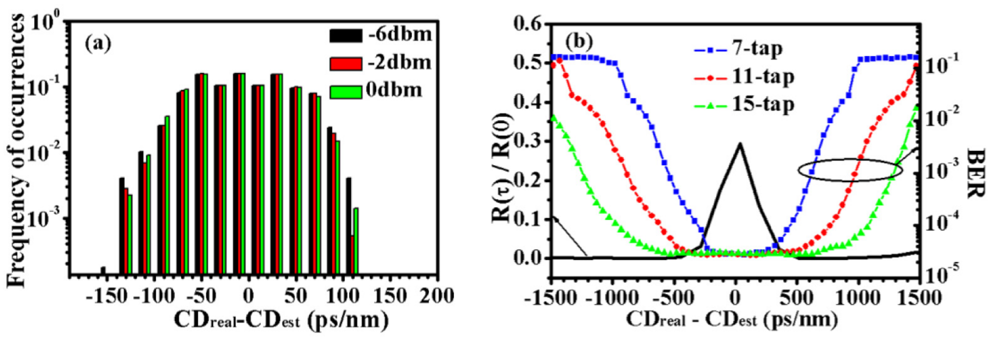

Fig. 11. (a) Histogram of the CD estimation error for the experimental 7× 112 Gbps NRZ DP-QPSK WDM signal after 480 km to 9120 km transmission at an interval of 480 km. N = 4096. (b) BER calculation in variance of CD estimation errors for the 7× 112 Gbps NRZ DP-QPSK WDM signal after 2400 km SMF transmission.

respectively. The launch power at the fiber link is −2 dBm and 4096 samples are used for the CD estimation. As expected, the CD estimation peaks are quite high compared to the noise floor which indicates a good estimation performance of our proposed CD estimation method.

To evaluate the CD estimation performance, the exact CD after the fiber loop is characterized through BER calculation of the received signal while the CD is scanned at a step size of 10 ps/nm over the range of [-500 ps/nm + CD_{est}, 500 ps/nm + CD_{est}]. In the DSP module, the tap number of 2 × 2 CMA equalizer is 3, allowing little residual CD compensation. The exact value of the accumulated CD is characterized when a lowest BER is observed. As shown in Fig. 10(b), the CD estimation error for the 7× 112 Gbps DP-QPSK WDM signal is around 20 ps/nm after 2400 km SMF transmission.

We then measure the CD estimation performance under the condition of different launch powers. By varying the input power and calculating the bit error rate of the received DP-QPSK signal, we find that the optimal launch power per channel is −2 dBm [18]. Increasing or decreasing the launch power leads to transmission performance degradation because of the OSNR degradation or enhanced nonlinearity effect. Here, −6 dBm, −2 dBm and 0 dBm launch power are taken into consideration for the CD estimation of 7× 112 Gbps DP-QPSK WDM signal. In Fig. 11(a), we carry out the CD estimation at the transmission distance from 480 km to 9120 km at an interval of 480km. For each launch power, 11400 independent estimation is processed by using different received samples. As shown in Fig. 11(a), there is no big difference among the CD estimation error distribution with different launch powers and the standard deviation is around 50 ps/nm. The worst CD estimation error detected is around 160 ps/nm. Thus, our proposed CD estimation method can operate well with different launch powers. In Fig. 11(b), we further calculate the BER of the 7× 112 Gbps DP-QPSK WDM signal in variance of CD estimation errors. As an inaccurate CD estimation may cause an incomplete CD compensation after static CD compensation, the residual CD can be further compensated by following 2 × 2 CMA equalizer. To ensure a low BER, an increased number of taps is required for 2 × 2 CMA equalizer in order to deal with severe CD estimation error. As shown in Fig. 11(b), in the presence of CD estimation error within 200 ps/nm, the residual CD can be effectively compensated by 2 × 2 CMA equalizer with a tap number of 7.

4. Conclusion

A fast and blind CD estimation method is proposed and experimentally demonstrated. By doing auto-correlation calculation in time domain after digital spectrum superposition, the temporal position of peak can be used to estimate the accumulated CD value. Without tentative CD scanning, the proposed method, has a greatly reduced processing time. Through simulation and experimental demonstration, the proposed CD estimation method is proved to be robust against ASE noise, PMD and optical nonlinearity even when using different modulation formats and pulse shaping techniques. Reliable CD estimation can be verified using only 4096 samples for 112 Gbps DP-QPSK and 224 Gbps DP-16QAM signals with either NRZ or RZ pulse shape with a standard deviation of 74 ps/nm and maximum detected estimation error of 210 ps/nm. For a root raised-cosine pulse, 8192 samples are shown to be sufficient for the CD estimation. When the roll off factor of the root-raised-cosine pulse is equal to 0.1, the standard deviation is 138 ps/nm and the worst estimation error detected is 375 ps/nm. The results reveal a comparable CD estimation precision for our proposed CD estimation method with other reported CD estimation methods. Therefore, with its fast operation speed and strong robustness, this CD estimation scheme is promising for digital coherent receivers.

Acknowledgments

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