Digital coherent superposition of optical OFDM subcarrier pairs with Hermitian symmetry for phase noise mitigation

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Abstract: Digital coherent superposition (DCS) provides an approach to combat fiber nonlinearities by trading off the spectrum efficiency. In analogy, we extend the concept of DCS to the optical OFDM subcarrier pairs with Hermitian symmetry to combat the linear and nonlinear phase noise. At the transmitter, we simply use a real-valued OFDM signal to drive a Mach-Zehnder (MZ) intensity modulator biased at the null point and the so-generated OFDM signal is Hermitian in the frequency domain. At receiver, after the conventional OFDM signal processing, we conduct DCS of the optical OFDM subcarrier pairs, which requires only conjugation and summation. We show that the inter-carrier-interference (ICI) due to phase noise can be reduced because of the Hermitain symmetry. In a simulation, this method improves the tolerance to the laser phase noise. In a nonlinear WDM transmission experiment, this method also achieves better performance under the influence of cross phase modulation (XPM).

OCIS codes: (060.0060) Fiber optics and optical communications; (060.4370) Nonlinear optics, fibers; (060.4080) Modulation.

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1. Introduction

Coherent optical OFDM (CO-OFDM) has been demonstrated as a viable solution for high-capacity and long-haul transmissions [1–4]. However it is well-known that OFDM is sensitive to phase noise leading to inter-carrier interference (ICI) [5]. It is aggravated in CO-OFDM because of the linear phase noise of laser sources and the nonlinear phase noise due to fiber nonlinearity [6], which limits the capacity of optical fiber transmissions. Some of the existing methods of reducing ICI lower the spectrum efficiency by half to combat phase noise [7]. This trade-off could be justified in certain scenarios. In fact, it has been proposed in single-carrier systems, namely digital coherent superposition (DCS) [8], where otherwise the DSP-based methods tend to have unpractical computational complexity to combat the fiber nonlinearity. For single-core fiber transmissions, Liu *et al.* have proposed phase-conjugated twin waves on two polarization tributaries over a record distance in fiber [9]. Tian *et al.* has demonstrated a conjugated copy via four-wave mixing, which can cancel the phase noise after DCS [10].

We have recently extended the concept of DCS to the optical OFDM subcarrier pairs with Hermitian symmetry centered on the optical carrier and we name it DCS-OFDM for simplicity [11]. At the transmitter, we simply use an MZ intensity-modulator biased at the null point to up-convert a real-valued OFDM signal with polarities. The so-generated OFDM signal is Hermitian in the frequency domain. At the receiver, after the conventional OFDM signal processing, we conduct DCS for OFDM subcarrier pairs, which requires only conjugation and summation. Therefore, DCS-OFDM is much simpler to be implemented, compared with the previous DCS experiments [8–10] for single carrier systems, and the existing ICI reduction methods reported in [7,12], which may have to use more complicated optical IQ modulators [12]. We show that the inter-carrier interference (ICI) resulted from phase noise can be decreased thanks to the Hermitian symmetry. In simulation, DCS-OFDM has an improved tolerance to laser phase noise. In a nonlinear WDM transmission experiment, compared with the conventional CO-OFDM, we demonstrate that DCS-OFDM can increase both the performance and the optimum launch power, and therefore has a better tolerance to cross-phase modulation (XPM).

2. ICI reduction by DCS of OFDM subcarrier pairs with Hermitian symmetry

Let $X(k)$ denote the transmitted data in subcarrier of an OFDM symbol. The low-pass equivalent output after IDFT with a length of N can be expressed as

$$s(m) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X(k) \exp(j2\pi km / N), \qquad k, m = 0, ..., N-1. \tag{1}$$

This output is real-valued if we pair up the OFDM subcarriers with Hermitian symmetry,

$$X(N-k) = X^*(k), \tag{2}$$

where * stands for conjugation. For simplicity, we have dropped the index of OFDM symbols in this paper. Equation (2) also means that half of the OFDM subcarriers carry the redundant data, which lowers the spectrum efficiency by half. To up-convert this real signal $s(m)$ with

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polarity to the optical domain, we use an intensity MZ modulator biased at the null point [13]. Note this configuration was used in the early demonstration of CO-OFDM, where the redundant OFDM subcarriers were also treated as the effective data for performance evaluation [1]. Here we re-use the redundant OFDM subcarriers by the concept of DCS to combat phase noise.

To focus on the phase noise $\phi(\mathrm{m})$, we ignore the channel response and the AWGN noise, and the received OFDM signal is

$$y(\mathrm{m}) = s(m)\cdot\exp[j\phi(m)]. \tag{3}$$

Assuming a perfect DFT window synchronization, the received OFDM subcarriers with common phase error (CPE) and ICI are [5]

$$Y_k = X_k \cdot I(0) + ICI(k), \tag{4}$$

$$ICI(k) = \sum_{l=0, l\neq k}^{N-1} X(l) I(k-l), \tag{5}$$

$$I(k) = \frac{1}{\mathrm{N}}\sum_{\mathrm{n=0}}^{\mathrm{N\text{-}1}} \exp[\mathrm{j}2\pi\mathrm{kn/N} + j\phi(\mathrm{n})]. \tag{6}$$

The CPE phase estimation is to calculate $\psi = \arg[\mathrm{I}(0)]$, i.e., the angle of *I(0)* [14]. We assume that the CPE phase estimation is accurate, then after the CPE phase compensation, the recovered OFDM subcarriers are

$$\hat{Y}_k = X_k \cdot \left|I(0)\right| + ICI(k)\cdot\exp(-j\psi). \tag{7}$$

Assuming that the transmitted data are mutually independent with zero mean and variance of E_s, and following a similar derivation in [15], the noise variance due to ICI is

$$\left\langle \left|ICI(k)\right|^2 \right\rangle = E_s \cdot \left(1 - \left\langle \left|I(0)\right|^2 \right\rangle\right), \tag{8}$$

where $\langle\cdot\rangle$ stands for the average operation. The DCS of OFDM subcarrier pairs with Hermitian symmetry can be expressed as

$$\left(\hat{Y}_k + \hat{Y}^*_{N-k}\right)/2 = X_k \cdot \left|I(0)\right| + ICI(k)\cdot\exp(-j\psi)/2 + ICI^*(N-k)\cdot\exp(j\psi)/2. \tag{9}$$

On the other hand,combining Eqs. (2) and (5) yields

$$ICI^*(N-k) = \sum_{l=0, l\neq k}^{N-1} X(l) I^*(l-k). \tag{10}$$

Inserting Eq. (10) to (9) yields

$$\left(\hat{Y}_k + \hat{Y}^*_{N-k}\right)/2 = X_k \cdot \left|I(0)\right| + ICI_{DCS}(k), \tag{11}$$

$$ICI_{DCS}(k) = \sum_{l=0, l\neq k}^{N-1} X(l) I_{DCS}(k-l), \tag{12}$$

$$I_{DCS}(k) = \frac{1}{\mathrm{N}}\sum_{\mathrm{n=0}}^{\mathrm{N\text{-}1}} \exp\left(j2\pi kn/N\right)\cdot\cos\left[\phi(n)\text{-}\psi\right]. \tag{13}$$

Comparing Eqs. (6) and (12), in the conventional OFDM, the ICI is from the phase noise $\exp[j\phi(n)]$,whereas in DCS-OFDM, the ICI is from $\cos\left[\phi(n)\text{-}\psi\right]$.

Equation (13) also means that $I_{DCS}(k)$ is the IDFT output of $\cos[\phi(n)\text{-}\psi]$ divided by a factor of $1/\sqrt{N}$ and further, the Fourier transform does not change the signal power, denoted as A. We obtain

$$A = \sum_{k=0}^{N-1} \left\langle \left| I_{DCS}(k) \right|^2 \right\rangle = \left\langle \left| \cos[\phi(\mathrm{n})\text{-}\psi] \right|^2 \right\rangle, \tag{14}$$

$$\left\langle \left| ICI_{DCS}(k) \right|^2 \right\rangle = E_s \cdot \left(A - \left\langle \left| I_{DCS}(0) \right|^2 \right\rangle \right). \tag{15}$$

To compare the values of Eqs. (8) and (15), firstly, from the definition of ψ, we have $\langle \phi(\mathrm{n})\text{-}\psi \rangle = 0$ and therefore,

$$\begin{aligned} \left\langle \left| I_{DCS}(0) \right|^2 \right\rangle &= \left\langle \left| \langle \cos[\phi(n) - \psi] \rangle \right|^2 \right\rangle \\ &\approx \left\langle \left| \langle \exp[j\phi(n) - j\psi] \rangle \right|^2 \right\rangle = \left\langle \left| \langle \exp[j\phi(n)] \rangle \right|^2 \right\rangle = \left\langle \left| I(0) \right|^2 \right\rangle. \end{aligned} \tag{16}$$

Secondly, from Eq. (14), we are certain that A<1. Therefore we conclude that $\left\langle \left| ICI_{DCS}(k) \right|^2 \right\rangle < \left\langle \left| ICI(k) \right|^2 \right\rangle$, which means that the ICI noise power is reduced. It also becomes clear that DCS-OFDM mitigates the ICI by reducing the energy of $\exp[j\phi(n)]$ to that of $\cos[\phi(n)\text{-}\psi]$.

Note that our derivation above is general and applicable to either laser phase noise or nonlinear phase noise. Our future research will quantify the ICI reduction, which requires the statistical characteristics of phase noise.

3. Simulation of DCS-OFDM under the laser phase noise

We conduct a simple simulation of CO-OFDM with single-polarization to investigate the effect of laser phase noise on DCS-OFDM. We have three cases for comparison:

Case I: A conventional CO-OFDM transmission using an IQ modulator at transmitter. This is the baseline for our comparison.

Case II: At the transmitter, we use a real-valued OFDM signal to drive an MZ intensity modulator biased at the null point. At the receiver, we follow the conventional DSP of CO-OFDM and count the BER for all the OFDM subcarriers, despite that half of them are redundant. This is identical to the configuration in [1], except that we use a direct down-conversion coherent receiver.

Case III: It is very similar to Case II, but we conduct the DSC for the OFDM subcarrier pairs with Hermitian symmetry, i.e., DCS-OFDM.

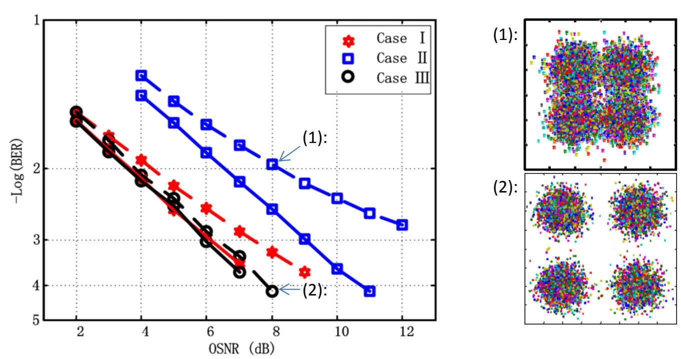

Fig. 1. BER comparison for three cases with 0-MHz laser linewdith (solid line) and 1-MHz (dash line). The insets on the left are the constellations at the indicated BER points.

In Case I, the sampling rate of DAC is 10 GS/S and IDFT length is 128. Both parameters are doubled in Case II and III. In three cases, we intentionally equal the OFDM subcarrier bandwidth, which is the main parameter to investigate the effect of the laser phase [3]. The cyclic prefix (CP) is 1/8 of IDFT length for all cases. The raw data rates are 17.8 Gb/s, 35.7 Gb/s, and 17.8 Gb/s, respectively.

Figure 1 shows the BER performance with 0-MHz or 1-MHz linewidth for both transmitter and receiver lasers. Case II has the largest OSNR penalty due to phase noise and the conventional OFDM in Case I has an OSNR penalty of 1.3 dB at BER of 10^{-3}. DCS-OFDM in Case III has the smallest OSNR penalty of 0.5 dB at BER of 10^{-3}, which means a better tolerance to the laser phase noise. With 1-MHz linewidth, from Case II to III, the OSNR is drastically reduced, much more than 3 dB, which follows the derivation in Section 2. The inset constellations in Fig. 1 also demonstrate this significant improvement, which results from the simple operation of Eq. (9).

4. Experiment of DCS-OFDM in a nonlinear WDM transmission

To verify the ICI reduction of DCS-OFDM, we conduct a nonlinear WDM transmission, where the transmitter uses a comb generator to emulate the multiple laser sources. As shown in Fig. 2, a 25-GHz clock signal drives a phase modulator followed by a wavelength selective switch (WSS) to flatten and select the comb lines. For comparison, all the comb lines pass through either an IM or IQ modulator driven by an arbitrary waveform generator (AWG) operated at 10 GS/s. The transmitted OFDM signal with 4-QAM format is generated off-line by a MATLAB program. The DFT length is 128 and we also use 1/8 of it as cyclic prefix. We use 44 OFDM subcarriers for Case I and 88 OFDM subcarriers for both Case II and III. This OFDM signal is further duplicated into three copies, or a super-channel [16], by another IM modulator driven at 6.875 GHz. Inside the transmitter, we also use two EDFAs to compensate for the losses. The inset spectra in Fig. 2 are before and after the data modulation, respectively. In short, the transmitter side includes 25 super-channels on 25-GHz WDM grid, covering 5-nm wide spectrum. The raw bit rate of one super-channel is 18.3 Gb/s for Case I and Case III, and 36.7 Gb/s for Case II. The frequency gaps among the super-channels are too narrow (a few GHz wide), and consequently, the spectrum in Fig. 2 measured by an optical spectrum analyzer seems gapless. The launch power per super-channel in the following measurement is obtained by dividing the total power of all channels by the channel number.

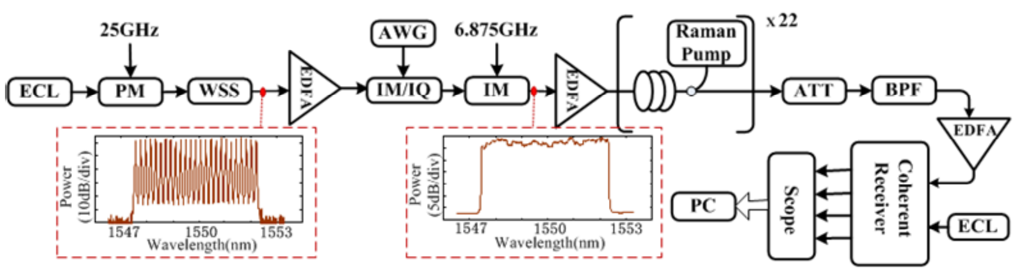

Fig. 2. Experimental setup of the emulated WDM transmission. The inset optical spectra are the optical signal before and after the data modulation. ECL: external cavity laser, PM: phase modulator, IM: intensity modulator, IQ: IQ modulator, WSS: wavelength selective switch, ATT: attenuator, BPF: bandpass filter.

The transmission link is 22 spans of 80-km standard single-mode fiber (SSMF) only using Raman amplification to compensate for the fiber loss. The launch power into each span is kept equal when we vary the launch power. The receiver side uses an optical attenuator, a bandpass filter and another EDFA to select the central channel before the coherent receiver. Both the transmitter and receiver lasers have a claimed laser linewidth below 100 kHz. Then we use a real-time scope operated at 50 GS/s as ADC and a computer to conduct off-line DSP, which covers the three cases defined in Section 3. We elect the estimated SNR to evaluate the performance [6]. Because the electrical dispersion compensation can play an important role in the nonlinear mitigation [9], we calculate the performance with or without the full dispersion compensation, which is based on the frequency domain compensation using the known dispersion value.

Figure 3 shows the performance of the three cases in the nonlinear transmission. At the linear transmission regime with the lower launch power, the conventional CO-OFDM of Case I is slightly better, which is due to that its narrower spectrum has a better tolerance to the fiber dispersion. Meanwhile, the SNR difference by DCS between Case II and Case III is 3 dB, which is as expected in the linear regime [8,9]. However, when the launch power is larger than −12 dBm per super-channel, the SNR improvement of Case III is apparently larger than 3 dB. Without electrical dispersion compensation, the maximum SNR improvement is 4.5 dB at the launch power of −7.4 dBm. With dispersion compensation, the estimated SNRs are all slightly improved for the three cases, and the maximum SNR improvement is increased to 4.9 dB. Correspondingly, Fig. 3(c) shows that the constellation spreading is dramatically reduced by DCS. Therefore, the electrical dispersion compensation can further improve the performance of DCS-OFDM in the nonlinear regime. Although it is not explained in Section 2, this similar conclusion was reported in the other experiment [9]. Finally, comparing Case III and Case I around the optimum launch power, we find that the SNRs of DCS-OFDM are apparently larger than that of the conventional CO-OFDM and its optimum launch power is increased by 3.4 dB. Based on the results in Fig. 3, we can conclude that DCS-OFDM trades off the spectrum efficiency for a better performance in the nonlinear WDM transmission, where XPM is dominant.

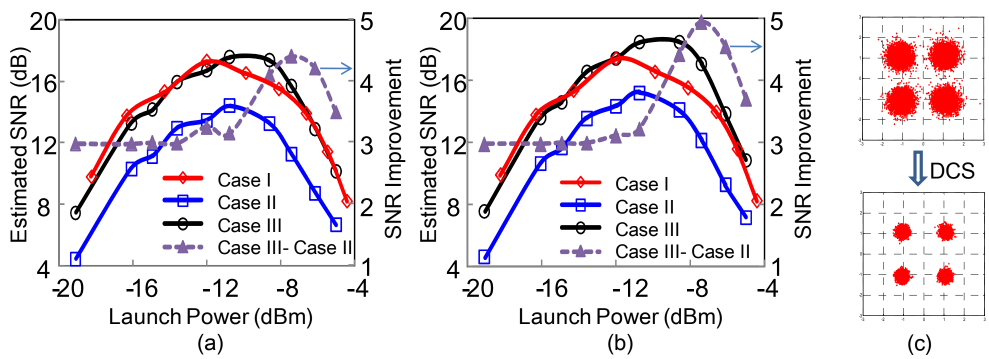

Fig. 3. Estimated SNR vs launch power in the nonlinear WDM transmission: (a) without electrical dispersion compensation, (b) with electrical dispersion compensation. The dash lines are the SNR difference between Case III and Case II. (c) Constellations before and after DCS at the maximum SNR improvement.

5. Conclusion

DCS trades off the spectrum efficiency to achieve a better transmission performance. We have extended the concept of DCS to the OFDM subcarrier pairs with Hermitian symmetry. The transmitter is simplified as an MZ intensity modulator driven with real-valued signals. The receiver's DSP only requires one additional conjugation and summation for DCS. We have shown that the ICI due to phase noise can be reduced. In simulation, DCS-OFDM has a better tolerance to laser phase noise. In a nonlinear WDM transmission experiment, DCS-OFDM increases both the optimum launch power and the maximum SNR under the influence of XPM. Further research is under plan to quantify the benefit of DCS-OFDM, both in theory and experiment.

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