Zero-guard-interval coherent optical OFDM with overlapped frequency-domain CD and PMD equalization

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Abstract: This paper presents a new channel estimation/equalization algorithm for coherent OFDM (CO-OFDM) digital receivers, which enables the elimination of the cyclic prefix (CP) for OFDM transmission. We term this new system as the zero-guard-interval (ZGI)-CO-OFDM. ZGI-CO-OFDM employs an overlapped frequency-domain equalizer (OFDE) to compensate both chromatic dispersion (CD) and polarization mode dispersion (PMD) before the OFDM demodulation. Despite the zero CP overhead, ZGI-CO-OFDM demonstrates a superior PMD tolerance than the previous reduced-GI (RGI)-CO-OFDM, which is verified under several different PMD conditions. Additionally, ZGI-CO-OFDM can improve the channel estimation accuracy under high PMD conditions by using a larger intra-symbol frequency-averaging (ISFA) length as compared to RGI-CO-OFDM. ZGI-CO-OFDM also enables the use of ever smaller fast Fourier transform (FFT) sizes (i.e. <128), while maintaining the zero CP overhead. Finally, we provide an analytical comparison of the computation complexity between the conventional, RGI- and ZGI-CO-OFDM. We show that ZGI-CO-OFDM requires reasonably small additional computation effort (~13.6%) compared to RGI-CO-OFDM for 112-Gb/s transmission over a 1600-km dispersion-uncompensated optical link.

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OCIS codes: (060.1660) Coherent communications; (060.2330) Fiber optics communications; (060.4080) Modulation.

References and links


1. Introduction

Coherent orthogonal frequency-division multiplexing (CO-OFDM) has emerged as a promising transmission technique for next-generation long-haul optical communication systems beyond 100 Gb/s [1–5]. The principal reasons that drive the interests in CO-OFDM are the simple and efficient channel estimation and equalization, the ability to allocate signal power and modulation format on a subcarrier basis, and the flexible oversampling rate [6,7]. However, there are also issues that may reduce the effectiveness of CO-OFDM when compared to the coherent single-carrier (SC) system. These issues include the vulnerability of CO-OFDM to the fiber nonlinearity, the phase noise (PN) from the transmitter and local oscillator (LO) lasers, and the dynamic range of the digital-analog converter (DAC) and analog-digital converter (ADC); CO-OFDM is also known to have large energy and spectral overhead from the insertion of cyclic prefix (CP), training symbol (TS) and pilot subcarrier (PS) [6,7]. Although these issues are OFDM specific, their impact on system performance has a strong dependence on the fast Fourier transform (FFT) size of OFDM symbols, \( N_{\text{FFT}} \). When \( N_{\text{FFT}} = 1 \), OFDM becomes essentially the same as SC. In this extreme scenario, we can circumvent all these OFDM-specific issues, but can no longer harness the OFDM benefits. Therefore, to design a coherent optical system for a given fiber link, bit rate and optical hardware, there may be an optimal \( N_{\text{FFT}} \) when considering the overall system sensitivity (e.g. Q-factor) and the computation complexity for real-time implementation [8].

In the conventional CO-OFDM, however, the design freedom to select \( N_{\text{FFT}} \) is constrained by the fiber dispersion, particularly the chromatic dispersion (CD) and polarization mode dispersion (PMD). It is known that CP or guard interval (GI) insertion between OFDM time symbols offers a convenient means to remove inter-symbol interference (ISI) from CD and PMD. However, CP limits the minimum value of \( N_{\text{FFT}} \) for a given CP overhead, which is defined as the ratio of the CP length \( N_{\text{CP}} \) and \( N_{\text{FFT}} \) and denoted as CP% in the reminder of this paper. Though one can reduce CP% by simply increasing \( N_{\text{FFT}} \), this will inevitably increase the CO-OFDM vulnerability to the fiber nonlinearity, laser PN and frequency offset [6,7]. In short we encounter a design tradeoff between having a small \( N_{\text{FFT}} \) (to achieve better sensitivity) and a small CP% (to achieve better energy and spectrum efficiency). For the conventional CO-OFDM demonstrated in previous works [1–3], the CP overhead was typically kept in the range of 10–25% (shown later in Table 1).

Furthermore, in conventional CO-OFDM \( N_{\text{CP}} \) needs to increase proportionally with the transmission distance and baud rate. Hence, to keep CP% unchanged, \( N_{\text{FFT}} \) must increase accordingly to support a link with a longer reach and higher data rate, but large \( N_{\text{FFT}} \) in turn...
makes CO-OFDM more susceptible to the laser PN and fiber nonlinearity. Therefore, the tradeoff between $N_{\text{FFT}}$ and CP% has fundamentally limited the scalability of the conventional CO-OFDM. In light of this, a variety of approaches have been studied to reduce CP% [4,9–11], among which a CO-OFDM receiver with two electronic channel equalization stages has shown a superior ability to circumvent this tradeoff. The essence of this approach is to compensate CD by using a separate overlapped frequency-domain equalizer (OFDE) stage in prior to the stage where the channel estimation and OFDM demodulation occurs. This two-stage equalization approach effectively decouples $N_{\text{CP}}$ from CD, thus $N_{\text{FFT}}$ can be reduced virtually independent of the transmission distance and baud rate. It was demonstrated experimentally that CO-OFDM with a small FFT of 128 and a small CP% of 3.13% can be achieved [3]. $N_{\text{CP}}$ of only 4 was used to compensate ISI from the time-varying PMD and transmitter bandwidth limitation [4]. This kind of CO-OFDM is termed as the reduced guard interval (RGI) CO-OFDM.

In this paper we present a modified equalization scheme for the two-stage CO-OFDM receiver to completely remove $N_{\text{CP}}$, by performing a joint CD and PMD compensation at the OFDE. We have proposed the same scheme in our earlier work to achieve a 0.8% CP% [11]. The objective of this work is to provide a systematic analysis of this equalization scheme, and for a specific example we demonstrate a 112-Gb/s polarization-division-multiplexing (PDM) CO-OFDM over 1600-km SMF with zero CP%. To distinguish from RGI-CO-OFDM in [4,11] and NGI-CO-OFDM for superchannel transmission in [12,13], we term our new system as the zero-GI CO-OFDM (ZGI-CO-OFDM).

To further illustrate the system limitation and tradeoff associated with CP%, Table 1 summarizes the key design parameters of the several CO-OFDM systems demonstrated in the recent years. With a great number of OFDM work accomplished worldwide, the CO-OFDM systems included in Table 1 are based primarily on experimental demonstrations. In Table 1, Refs [1–3,5,14] represent the conventional CO-OFDM, while Ref [4] is an example of RGI-CO-OFDM. We also include the proposed ZGI-CO-OFDM for comparison. The conventional CO-OFDM has a typical CP% of 10-25%, and CP% must scale up proportionally with increased fiber transmission distance and baud rate. However, CP% does not necessarily scale with the total bit rate, due to the multi-band CO-OFDM implementation [1,3–5]. In Table 1 we use the single-band bandwidth (SB-BW) as an equivalent to the baud rate. Note that Ref [5] has an exceptionally low CP%, simply due to its short transmission distance of 80 km and relatively low SB-BW of 6 GHz. Furthermore, compared to the conventional CO-OFDM, RGI- in [4] and ZGI-CO-OFDM in this work not only significantly reduces CP%, but it also makes CP% to have little dependence on the fiber distance and SB-BW. This unique property will become extremely useful when we scale CO-OFDM for ever higher data rates and longer reach in the future. Also ZGI-CO-OFDM can be readily used for multiband CO-OFDM, though we only demonstrate its concept and implementation in a single-band setting in this paper.
The remainder of the paper is organized as follows. In Section B, we describe the operation principle and algorithm of the new equalization scheme. In the new scheme, the main functionality of the first-stage OFDE is to compensate CD and PMD simultaneously, which offers the key to the zero CP transmission. We also compare the equalization process of ZGI-CO-OFDM to that of the conventional and RGI CO-OFDM, and reveal an underlying design tradeoff between CP% and the computation complexity of the algorithm. Unique to the ZGI-CO-OFDM receiver, a frequency domain interpolation (FDI) is required, because the FFT size used in the OFDE, $N_{OFDE}$, is usually larger than that used for OFDM symbols, $N_{FFT}$. Examples of the FDI implementation are also shown. In Section C, we demonstrate the system performance of a 112-Gb/s PDM ZGI-CO-OFDM. Besides the zero CP%, the new scheme provides three additional system benefits. We show that not only can it achieve a larger PMD tolerance, but also it can improve the channel estimation accuracy by increasing the length of the intra-symbol frequency averaging (ISFA) [14], when comparing to RGI-CO-OFDM under a high PMD scenario. Moreover, our system enables the use of an even smaller $N_{FFT}$ (e.g. 16 and 32) while still keeping zero CP%. This ability would become beneficial, when we design to achieve the optimal $N_{FFT}$ for a given fiber link and baud rate. Finally in Section D, we provide an analytical comparison of the computational complexity between the conventional, RGI- and ZGI-CO-OFDM. We show that ZGI-CO-OFDM require reasonably small additional computation effort compared to RGI-CO-OFDM, while providing several system benefits.

2. Principle of Operation

Figure 1(a) illustrates the CO-OFDM receiver structure used in this work. The OFDM transmitter and fiber link are not shown, as they are the same as those in the conventional OFDM. Besides CD compensation, one key feature of the OFDE in our new scheme is to acquire the channel estimation (in form of a 2-by-2 matrix $H[k]$ for each $k^{th}$ modulated subcarrier) from the OFDM demodulator once the TS’s have been processed, and then to compensate PMD by applying the inverse of the channel matrix. Therefore, with both CD and PMD compensated at the OFDE, subsequent data symbols can be transmitted without any CP. Figure 1(b) shows the OFDM frame used with the new equalization scheme. A small CP is still required for TS’s for accurate channel estimation, while no CP is allocated for data symbols. Note that this OFDM frame is different from the typical ones used in the conventional or RGI CO-OFDM, in which the same CP is encapsulated in both TS’s and data symbols. When compared to RGI-CO-OFDM in [4], where $N_{CP} = 4$ was used for all TS’s and data symbols, in our system $N_{CP} = 4$ is only used for each TS. Note that a very short CP length (for example, $N_{CP} = 1$) can be still used for each data symbol, so our system would become more robust against any residual ISI from the imperfect channel equalization at the OFDE. However, a small CP% of 0.8% ($ = 1/128$) will be required as a tradeoff. The origin of the equalization imperfection at the OFDE will be explained later.
2.1 Two-stage equalization algorithm

To elaborate the new equalization scheme, we describe the algorithm step-by-step as follows:

Step 1: At the OFDE, perform CD compensation on the incoming TS’s ($t_1$ and $t_2$) and produce $t_1'$ and $t_2'$. Note that $t_1$ and $t_2$ are received from the x and y polarizations, respectively. The CD compensation is done by simply multiplying the incoming signals with the inverse of the CD transfer function $H_{cd}$,

$$H_{cd} = e^{jDL(k_{OFDE}N_{sc}^f \Delta f_s / c)}, k_{OFDE} = 1, \ldots, N_{scud}. \quad (1)$$

where $D$ is the group velocity dispersion parameter, $L$ is the fiber transmission distance, $\lambda$ is the optical carrier wavelength in vacuum, $\Delta f_s$ is the frequency spacing in the OFDE, and $c$ is the speed of light in vacuum. The FFT size used in the OFDE, $N_{OFDE}$, is often larger than that used for OFDM symbols, $N_{FFT}$, thus $\Delta f_s$ is smaller than the subcarrier spacing. $k_{OFDE}$ is the frequency index at the OFDE.

Step 2: At the OFDM demodulator, estimate the channel matrix $H[k]$ for each of the $N_{sc}$ modulated subcarriers using $t_1'$ and $t_2'$. The ISFA can be applied, but the averaging length $m$ is limited by the amount of the uncompensated PMD [14]. For the $k$th modulated subcarrier, the estimated channel matrix is

$$H[k] = \begin{bmatrix} a[k] & b[k] \\ c[k] & d[k] \end{bmatrix}, k = 1, \ldots, N_{sc}. \quad (2)$$

Step 3: The OFDE acquires $H$ from the OFDM demodulator, and performs the frequency-domain interpolation (FDI) to map $H$ to $H_{FDI}$. $H$ represents $N_c$ 2-by-2 matrices, while the number of 2-by-2 matrices in $H_{FDI}$ is equal to $N_c \cdot (N_{scud} \cdot N_{rev})$, and $N_{OFDE}$ is usually larger than $N_{FFT}$.
\[
H_{FDI}[k'] = \begin{bmatrix}
               a_{FDI}[k'] & b_{FDI}[k'] \\
               c_{FDI}[k'] & c_{FDI}[k']
             \end{bmatrix}, \quad k' = 1, \ldots, N_{\text{SC}} \cdot (N_{\text{OFDE}} / N_{\text{FFT}}). \tag{3}
\]

Step 4: At the OFDE, multiply \( t_1' \) and \( t_2' \) by \( H_{FDI}^{-1} \) to produce \( t_1'' \) and \( t_2'' \). In essence, this is a multiple-input-multiple-output (MIMO) demodulation process, which compensates PMD and restores any polarization rotation. In the conventional and RGI CO-OFDM, however, this MIMO demodulation occurs at the OFDM demodulator instead of at the OFDE stage. This step is the key to enable our system to operate with zero CP%.

Step 5: At the OFDM demodulator, estimate the new channel matrix \( H[k'] \) for each of the \( N_{\text{sc}} \) subcarriers now using \( t_1'' \) and \( t_2'' \). Same as \( H \), \( H' \) represents \( N_{\text{sc}} \) 2-by-2 channel matrices. Now a large ISFA length \( m \) can be used to improve the estimation accuracy of \( H' \), because both CD and PMD have been compensated. \( H' \) will be saved and used to demodulate the subsequent data symbols. Subsequently, we can perform PN estimation and compensation after applying \( H' \) to the data symbols.

Unlike the channel estimation/compensation used for the conventional and RGI CO-OFDM, our new scheme relies on a collaborative effort between the OFDE and OFDM demodulator, and it requires processing the same TS and performing channel estimation twice. To understand our algorithm from a mathematical standpoint, we express each of the \( N_{\text{sc}} \) demodulated OFDM subcarriers \( r_{\text{out}} \) obtained after Step 5 as,

\[
Y_{\text{out}}[k] = H_{\text{cd}}^{-1} \cdot H_{\text{FDI}}^{-1} \cdot H'[k]^{-1} \cdot Y_{\text{in}}[k]. \tag{4}
\]

In Eq. (4) \( r_{\text{in}}[k] \) is the \( k^\text{th} \) received OFDM subcarrier at the OFDE input. Both \( r_{\text{out}} \) and \( r_{\text{in}} \) are the 2-by-1 column vectors, representing I-Q data on both x- and y-polarization. \( (H_{\text{cd}}^{-1} \cdot H_{\text{FDI}}^{-1}) \) is applied to the received signal at the OFDE, while \( H'[k]^{-1} \) is multiplied at the OFDM demodulator. Also \( (H_{\text{cd}}^{-1} \cdot H_{\text{FDI}}^{-1}) \) is a function of \( N_{\text{OFDE}} \) frequency points, and we don’t express it as function of \( k \), because there is a lack of the explicit one-to-one correspondence between \( n \) and \( k \). Note that the term for PN compensation is not included in Eq. (4), since it is the same process as that for the conventional and RGI-CO-OFDM.

In comparison with ZGI-CO-OFDM, RGI-CO-OFDM uses a relatively straightforward two-stage equalization process but it has to incorporate a larger CP. In RGI-CO-OFDM the OFDE only compensates CD, while the channel compensation for PMD, polarization demultiplexing and PN are performed at the OFDM demodulator. Mathematically, we can express the equalization process of RGI-CO-OFM as,

\[
Y_{\text{out}}[k] = H_{\text{cd}}^{-1} \cdot H_{\text{RGI}}[k]^{-1} \cdot Y_{\text{in}}[k]. \tag{5}
\]

All the channel equalization operations at the OFDM demodulator are lumped in \( H_{\text{RGI}}[k]^{-1} \), which is obtained using the TS based channel estimation. Finally, the channel equalization for the conventional CO-OFDM is even more straightforward, which can be simply expressed as

\[
Y_{\text{out}}[k] = H_{\text{CONV}}[k]^{-1} \cdot Y_{\text{in}}[k]. \tag{6}
\]

But this system must have a very large CP to ensure an accurate estimation of \( H_{\text{CONV}}[k] \) from TS’s. Comparing Eqs. (4), (5), and (6), a tradeoff can be observed between the equalization algorithm complexity and CP%. We will compare the computation complexity of the three CO-OFDM algorithms analytically in Section D of this paper.
2.2 Frequency domain interpolation (FDI)

As mentioned earlier ZGI-CO-OFDM requires FDI to map $H$ to $H_{FDI}$ at Step 3, due to the mismatch between $N_{OFDE}$ and $N_{FFT}$. To illustrate FDI implementation, we use a ZGI-CO-OFDM system with $N_{OFDE} = 2048$ and $N_{FFT} = 128$, so the interpolation factor is $N_{OFDE}/N_{FFT} = 16$. In Fig. 2(a) the open circles are used to indicate the channel matrix $H$, and the curve connecting the open circles is the result from FDI. In this particular example, we assume each OFDM symbol contains 64 modulated subcarriers occupying the center half of the $N_{FFT} = 128$ window, so that the oversampling factor at the ADC/ADC is 2. Figure 2(b) illustrates the spectrum of the real part of the x-pol OFDM signal before and after applying $H_{FDI}^{-1}$ at Step 4, where we assume that the fiber link has a deterministic DGD of 320 ps. We show only the real part of the signal in order to highlight the difference in spectrum before and after applying $H_{FDI}^{-1}$, because $H_{FDI}^{-1}$ only imparts a phase rotation given our assumption of a deterministic DGD. Note that a large DGD of 320 ps is not typically encountered in practice, but it is used in Fig. 2 to illustrate the FDI process. Smaller DGD values are be used to study the transmission system performance in the later sections.

Moreover, the OFDM signal is confined in the center half of the OFDE FFT window as expected from only modulating 64 middle subcarriers, and $H_{FDI}^{-1}$ is only multiplied by the OFDM signal within this center half of the spectrum. On the top of the signal spectrum, the curve shows the real part of $a$ in the channel matrix $H$ (open dot) and the interpolated channel matrix $H_{FDI}$ (thin line). Phase variations across modulated subcarriers before and after PMD compensation in Step (4) for a (c) deterministic DGD = 320ps and (d) stochastic PMD with $\langle DGD \rangle = 100$ps.

![Fig. 2. (a) Schematic of FDI. (b) OFDM spectrum before and after applying $H_{FDI}^{-1}$. The top curve shows the real part of $a$ in the channel matrix $H$ (open dot) and the interpolated channel matrix $H_{FDI}$ (thin line). Phase variations across modulated subcarriers before and after PMD compensation in Step (4) for a (c) deterministic DGD = 320ps and (d) stochastic PMD with $\langle DGD \rangle = 100$ps.](image-url)
imperfection and thus the residue ISI. However, \( N_{\text{FFT}} \) for data symbols can remain unchanged. Finally, Fig. 2(c) illustrates the phase variations across the 64 modulated subcarriers on both polarizations before and after Step 4. Before Step 4, the phase varies linearly across the subcarriers due to a deterministic DGD of 320 ps. After Step 4, the phase variations approach to zero for all subcarriers. This indicates a successful PMD compensation by removing the PMD-induced phase rotation from each subcarrier. Figure 2(d) shows the same kind of plot but considering a stochastic PMD with a mean DGD (i.e. \(<\text{DGD}>)\) of 100 ps. Before Step 4, the phase varies across the subcarriers but not in a linear fashion due to a random PMD; after Step 4, the phase variations converge to zero as Fig. 2(c), indicating a successful PMD compensation.

3. System performance

We investigate the transmission performance of ZGI-CO-OFDM numerically using OptiSystem 9.0 and Matlab. The system parameters are chosen similar to [4] and [13,14] so that we can compare directly to the conventional and RGI CO-OFDM. Specifically, we simulate a 112-Gb/s PDM CO-OFDM system over a link that consists of 1600-km (20 spans of 80-km) SMF without optical dispersion compensation. For each polarization 56-Gb/s data are encoded on the 80 QPSK-modulated subcarriers (including one pilot subcarrier for laser phase compensation). We use \( N_{\text{FFT}} = 128 \) for both TS and data symbols, resulting in an oversampling factor of 1.6 (\( = 128/80 \)). The SMF has an attenuation of 0.2 dB/km, dispersion of 16.75 ps/nm/km, and nonlinearity coefficient of \( \gamma = 1.32 \text{ W}^{-1}\text{km}^{-1} \). For the OFDM frame design, we use 4 TS’s for every 300 data symbols for synchronization and channel estimation (see Fig. 1(b)). At the receiver side, the OFDE uses \( N_{\text{OFDE}} = 2048 \) and the overlapped sample size of 512. This overlapped sample size is sufficiently large to support a DGD of \(-2000\) ps in addition to compensating CD from the 1600-km SMF link. The launch power is fixed at \(-6\) dBm to avoid fiber nonlinearities, and the linewidths of the transmitter and local oscillator lasers are assumed to be zero unless indicated otherwise.

3.1. Improved PMD tolerance

Figure 3 compares the received signal \( Q \) factor as a function of deterministic DGD’s for the conventional, RGI- and ZGI- CO-OFDM systems. The same ISFA length of 5 (i.e. \( m = 2 \)) and received OSNR of 14.5 dB are used for a fair comparison, though we will show later that a larger \( Q \) can be achieved using a larger \( m \) for a given DGD. In Fig. 3 the conventional CO-OFDM shows the best PMD tolerance, as the \( Q \) remains constant across the DGD values up to 320 ps. However, this superior PMD tolerance is achieved at the expense of a large \( N_{\text{FFT}} \) of 2048 and a large CP% of 25% [14].

In contrast, RGI-CO-OFDM in Fig. 3 shows the worse PMD tolerance of \(-70\) ps, which is consistent with the discussion in [4]. The \( Q \) degrades drastically as DGD further increases beyond 70 ps. This limited PMD tolerance is directly resulted from the small \( N_{\text{CP}} = 4 \) used for TS’s and data symbols. A larger PMD tolerance can be achieved by increasing \( N_{\text{CP}} \), but this will increase CP% further above 3.13%. For ZGI-CO-OFDM (CP% = 0%), however, the \( Q \) does not roll off until a DGD of \(-125\) ps. The \( Q \) starts to decrease when DGD becomes larger than 125 ps, because the channel estimation accuracy from the TS’s eventually degrades when the PMD-induced ISI becomes significant. Interestingly, ZGI-CO-OFDM shows a superior PMD tolerance than RGI-CO-OFDM, while both systems use exactly the same TS with \( N_{\text{CP}} = 4 \). We believe this improved PMD tolerance is related to the fact that ZGI-CO-OFDM processes the same TS and performs channel estimation twice. This can be seen from Fig. 2(c) and 2(d) that multiplying \((H_{\text{cd}}^{-1} \cdot H_{\text{di}}^{-1})\) effectively removes the CD and PMD induced phase variations, even though the deterministic and mean DGD of 320 ps and 100 ps used are both greater than the prescribed DGD tolerance of \(-70\) ps from \( N_{\text{CP}} = 4 \) in the TS. However, it is also worth noting that ZGI-CO-OFDM produces a lower \( Q \) than RGI-CO-OFDM for the
DGD values between 0 and 50 ps. This difference can be attributed to the Step 4 of the ZGI-CO-OFDM algorithm, where the equalization is applied only to the center part of the OFDM spectrum (see Fig. 2(b)) and thus induces a $Q$ penalty. The slight increase of $Q$ around DGD = 100 ps for ZGI-CO-OFDM may be a result of PMD-supported transmission [15].

In reality PMD is a time-varying stochastic process, and it needs to be estimated and compensated periodically using TS’s. Therefore, to verify the PMD tolerance of the ZGI-CO-OFDM, a statistical investigation of the system performance is needed for a large set of PMD realizations. Before discussing the $Q$ factor results, we compare the phase variations across the 80 modulated subcarriers before and after Step 4 for 25 different PMD realizations. Figure 4 illustrates these phase variations in a contour plot assuming a $<\text{DGD}>$ of 5 ps. Note that the same TS with $N_{CP} = 4$ is used. Comparing Fig. 4(a) and 4(b), the phase variations across subcarriers approach zero after Step 4 for all 25 PMD realizations, indicating our algorithm achieves a robust PMD compensation. This enables us to eliminate the CP from the data symbols without incurring $Q$ degradation. It is worth mentioning that the phase variations in Fig. 4(a) and 4(b) are irregular as a result of a random higher-order PMD, and the phase variations across the subcarriers are normalized to the phase variation of the first subcarrier.

Figure 5 and 6 illustrate the same contour plot but for a larger $<\text{DGD}>$ of 10 ps and 25 ps, respectively. Before Step 4, we observe more drastic phase variations across subcarriers due to an increased $<\text{DGD}>$. This can be seen by noting that the color bar has a larger maximum value. The residual phase variations after Step 4 are also nearly zero. It is worth noting that the residual phase variations will become significantly larger, when $<\text{DGD}>$ further increases and causes the TS-based channel estimation to degrade. Therefore, for a practical CO-OFDM design one can use a larger $N_{CP}$ in the TS to reduce the residual phase variations, which effectively increases the system PMD tolerance. Alternatively, one may consider including a very short CP in each data symbol to guard against the residual dispersion. For example, using $N_{CP} = 1$ for each data symbol for our current system with $N_{FFT} = 128$ will increase the CP% from 0% to 0.8% (see Fig. 7–9). However, this CP% is still significantly smaller than that in the conventional and RGI-CO-OFDM.
Fig. 4. Contour plot of the estimated phase variations across modulated subcarriers on x-polarization before (a) and after (b) Step 4. \(<\text{DGD}> = 5\) ps is assumed.

Fig. 5. Contour plot of the estimated phase variations across modulated subcarriers on x-polarization before (a) and after (b) Step 4. \(<\text{DGD}> = 10\) ps is assumed.

Fig. 6. Contour plot of the estimated phase variations across modulated subcarriers on x-polarization before (a) and after (b) Step 4. \(<\text{DGD}> = 25\) ps is assumed.

Furthermore, we evaluate the system performance in term of \(Q\) for 500 different PMD realizations. Figure 7, 8, and 9 compare the \(Q\) factor distribution for ZGI- and RGI-CO-OFDM for three different \(<\text{DGD}>\) values of 10, 25 and 50 ps, respectively. We use the same design parameters for RGI- and ZGI-CO-OFDM as those in Fig. 3. We also simulate the
scenario where $N_{CP} = 1$ is used for each data symbol of ZGI-CO-OFDM. When $<\text{DGD}>$ is small (i.e. 10 ps and 25 ps) so that the instantaneous DGD has large probability within the designed PMD tolerance of ~70 ps from using $N_{CP} = 4$, Fig. 7 and 8 show that RGI-CO-OFDM produces a smaller variance of $Q$ than ZGI-CO-OFDM in the presence of varying PMD. This observation agrees with the fact that ZGI-CO-OFDM produces a lower $Q$ for the DGD between 0 and 50 ps in Fig. 3, which is related to the equalization-induced signal degradation as discussed earlier. However, as $<\text{DGD}>$ increases the $Q$ distribution of RGI-CO-OFDM experiences a drastic change. This is apparent by comparing Fig. 7–9(c) that the $Q$ distributions become wider and more flattened. This is because with RGI-CO-OFDM, $Q$ penalty occurs whenever the instantaneous DGD exceeds the system’s prescribed PMD tolerance of ~70 ps. In contrast, with ZGI-CO-OFDM if the OFDE makes effort to equalize the PMD, thus its $Q$ histograms remain relatively stable against the varying PMD realizations. For large $<\text{DGD}>$ values, ZGI-CO-OFDM is expected to achieve a superior PMD tolerance by having a smaller $Q$ variance as shown in Fig. 9. This is also consistent with our observation in Fig. 3. Additionally, Fig. 7–9(b) show that ZGI-CO-OFDM with $N_{CP} = 1$ produces an even smaller $Q$ variance and better PMD tolerance than that with $N_{CP} = 0$ as expected. Ultimately, to transmit over a given fiber link, one needs to tradeoff the system PMD tolerance with the CP overhead. For a link with a small $<\text{DGD}>$ and the overhead requirement is not stringent, one may prefer RGI-CO-OFDM to achieve smaller fluctuation in system $Q$ (as shown in Fig. 7 and 8). In the opposite case, ZGI-CO-OFDM is an excellent candidate to achieve both the small CP overhead and large PMD tolerance simultaneously.

![Fig. 7. Q factor distribution after transmission over a fiber link with 500 different PMD for (a) ZGI-CO-OFDM (0% CP) (b) ZGI-CO-OFDM (0.8% CP) and (c) RGI-CO-OFDM (3.13% CP). We assume $<\text{DGD}> = 10$ ps.](image)

![Fig. 8. Q factor distribution after transmission over a fiber link with 500 different PMD for (a) ZGI-CO-OFDM (0% CP) (b) ZGI-CO-OFDM (0.8% CP) and (c) RGI-CO-OFDM (3.13% CP). We assume $<\text{DGD}> = 25$ ps.](image)
3.2. Improved channel estimation accuracy under high PMD

In conventional CO-OFDM, ISFA can effectively improve the TS based channel estimation accuracy in the presence of optical noise; with a sufficiently large averaging length $m$, the channel estimation can reproduce the ideal channel matrices, so that an optimal $Q$ can be achieved [14]. However, the averaging length $m$ is limited by the residual CD and PMD on the TS, because a large CD and PMD reduce the coherency between the subcarrier channels [14]. Derived from Eq. (6) in [14], we can write the upper bound for $m$ due to the residual CD as,

$$m < \frac{10^6 \cdot N_{sc}}{8 \pi \cdot \Delta f_{OFDM}^2 \cdot |CD_{ISFA}|}, \quad (7)$$

where $N_{sc}$ is the number of the modulated subcarriers, $\Delta f_{OFDM}$ is the optical bandwidth of the OFDM signal and $CD_{ISFA}$ is the residual CD. For example, in a 112-Gb/s conventional PDM CO-OFDM system with $N_{sc}$ and $N_{FFT}$ are 1280 and 2048, respectively. The $N_{sc}$ and $N_{FFT}$ are chosen so that the same oversampling factor of 1.6 is used for the RGI- and ZGI-CO-OFDM. $\Delta f_{OFDM}$ (GHz) is $56 \times 1280/2048 = 35$, and $CD_{ISFA}$ (ps/nm) is 26800 for the 1600-km SMF link used in this work. In this case we require $m<1.53$, thus only a very small averaging length (i.e. $m = 1$) can be used, unless CD is pre-compensated before the channel estimation. In [14] when $D_{ISFA}$ (ps/nm) reduces to 6850, $m = 6$ can be used to produce the ideal channel estimation and thus achieve the optimal $Q$. Note that the optimal $Q$ is the $Q$ factor when ideal channel estimation is assumed [14]. Furthermore, following Eq. (7) in [14] we express the upper bound for $m$ due to the PMD as,

$$m < \frac{10^3 \cdot N_{sc}}{\Delta f_{OFDM} \cdot |PMD_{ISFA}|}, \quad (8)$$

where $PMD_{ISFA}$ measures the instantaneous DGD in ps. In [14] $PMD_{ISFA} < 610$ ps is required to achieve $m = 6$.

It is apparent from Eqs. (6) and (7) that the upper bound for $m$ scales proportionally with $N_{sc}$. For RGI- and ZGI-CO-OFDM, where $N_{sc}$ and $N_{FFT}$ are significantly smaller, the residual CD and PMD must be very small in order to use the same $m$. For example, both the RGI- and ZGI-CO-OFDM in Fig. 3 have $N_{sc} = 80$, which is 16 times smaller than that used in the conventional CO-OFDM. Therefore, the residual CD and PMD need to reduce 16 times to keep $m = 6$. Though CD can be compensated close to zero using the OFDE, PMD is left uncompensated before the channel estimation in RGI-CO-OFDM. This significantly limits the ISFA effectiveness for RGI-CO-OFDM. Specifically, $PMD_{ISFA}$ must be smaller than 610/16 = 38 ps for $m = 6$. However, ZGI-CO-OFDM can circumvent this limitation, because both CD and PMD are compensated before the final channel estimation at Step 5. Though some residue...
dispersion may exist, it is much smaller compared to the uncompensated PMD. Therefore, a large ISFA averaging length \( m \) can be used to improve channel estimation accuracy at Step 5.

Figure 10 compares the \( Q \) as a function of the ISFA length \( m \) for different deterministic DGD values. For RGI-CO-OFDM in Fig. 10(a), due to the small \( N_{sc} = 80 \), \( m \) tends to be upper bounded at a small value, and increasing \( m \) beyond this bound results in \( Q \) degradation. This same effect has been observed in a conventional CO-OFDM system [16]. For example, for DGD = 75 and 100 ps the maximum \( Q \) of 9.05 dB and 8.5 dB are achieved at \( m = 4 \) and \( m = 3 \), respectively; and the \( Q \) cannot be improved further with a larger \( m \). However, for a smaller DGD of 25 ps, the \( m \) upper bound moves near to 6, where the maximum \( Q \) of 10 dB is achieved. We have also calculated the \( Q \) curves for DGD smaller than 25 ps, and found these curves coincide with that for DGD = 25 ps. The maximum \( Q \) of 10 dB achieved at DGD = 25 ps is essentially the optimal \( Q \).

With ZGI-CO-OFDM in Fig. 10(b), however, we do not observe the \( m \) upper bound for the same DGD and \( m \) values used in Fig. 10(a). This is because the residue CD and PMD are almost zero when the OFDM signal arrives at the OFDM demodulator at Step 5. As a result, a large \( m \) can be applied to approach an optimal \( Q \). Note that the same \( Q \) vs. \( m \) curves are achieved for DGD = 50 ps and smaller. The maximum \( Q \) of 10 dB achieved with DGD = 50 ps is the optimal \( Q \), the same as that in Fig. 10(a). Compared to RGI-CO-OFDM, the maximum \( Q \) achieved using ZGI-CO-OFDM is 0.45 dB larger for DGD = 50 ps. Additionally, in Fig. 10(b) the maximum \( Q \)'s achieved with DGD = 75 and 100 ps are slightly smaller than that achieved with DGD = 50 ps. This should be attributed to the degraded channel estimation by using a small \( N_{cp} = 4 \).

![Fig. 10. Q vs. ISFA length for 3 different deterministic DGD values for (a) RGI and (c) ZGI CO-OFDM.](image)

3.3. Scalability to small FFT size

Finally, ZGI-CO-OFDM completely circumvents the design tradeoff between \( N_{FFT} \) and \( N_{CP} \). This unique property enables one to choose an optimal \( N_{FFT} \) for CO-OFDM transmission while still maintaining the zero CP%, which cannot be achieved with the conventional and RGI-CO-OFDM. As an example of RGI-CO-OFDM in [4], using \( N_{FFT} = 64 \) instead of 128 will double the CP% from 3.13% to 6.26%, if \( N_{CP} = 4 \) is kept to achieve the same PMD tolerance. For ZGI-CO-OFDM, however, \( N_{FFT} = 64 \) can be used with zero CP% (\( N_{CP} = 4 \) in the TS is still required). Many system parameters can impact significantly on the optimal \( N_{FFT} \), such as transmission distance/fiber type, baud rate, launch power, laser linewidth, etc. Therefore, the optimum of \( N_{FFT} \) needs to be evaluated on a case-by-case basis. Nevertheless, ZGI-CO-OFDM allows us to search and eventually operate at the optimal \( N_{FFT} \) without a concern of varying CP%.

To exemplify this unique ability, Fig. 11 shows the \( Q \) of ZGI-CO-OFDM as a function of \( N_{FFT} \) for three different DGD values. We assume the oversampling factor is 1.6, and the DAC/ADC has 5 resolution bits and a clipping ratio of 8 dB for all \( N_{FFT} \). Each \( Q \) is optimized...
using the largest possible ISFA \( m \) that satisfies the condition (8) but the maximum \( m \) is set to 6. Moreover, we assume a launch power is \(-2\) dBm and both the transmitter and local oscillator lasers have the linewidth of 100 kHz. The PS overhead (PS\%) of 10% is used for all \( N_{FFT} \), except that the minimum number of PS’s is kept at 8 to ensure sufficient averaging to improve phase estimation accuracy. Thus for the small \( N_{FFT} = 32 \) (\( N_{SC} = 20 \)) and \( N_{FFT} = 64 \) (\( N_{SC} = 40 \)), PS\% are 40% and 20%, respectively. Note that these large PS\% can be reduced significantly, if novel PN compensation approaches such as RF pilot tone [2] or decision directed processing are employed [17].

It can be seen from Fig. 11, the \( Q \) is maximized at the intermediate \( N_{FFT} \) values. A \( Q \) penalty occurs for a large \( N_{FFT} \) of 1024 and 2048, due to the increased peak-to-power ratio (PAPR) and phase noise. We have also verified that this \( Q \) penalty would become even larger, if a larger launch power (i.e. larger fiber nonlinearity) [18] or a larger laser linewidth (i.e. larger PN) is considered. On the other hand, when \( N_{FFT} \) decreases from 128, the \( Q \) begins to roll off due to the combination of the spectral leakage and the limited channel estimation/equalization ability. Particularly, the \( Q \) rolls off drastically for a large DGD. This is because for a small \( N_{FFT} \), a large DGD limits the channel estimation accuracy so that only a small \( m \) can be used at Step 2; and this is also attributed to the coarse frequency resolution before FDI, which limits the PMD compensation capability. As in Fig. 11 the maximum \( Q \) of 11.14 dB is achieved at \( N_{FFT} = 512 \), while the \( Q \) for \( N_{FFT} = 64 \) when DGD = 0 ps only decreases slightly to 10.72 dB. Hence in this particular case, one would probably choose \( N_{FFT} = 512 \) as the optimal FFT size. Nevertheless, for all \( N_{FFT} \) values in Fig. 11 ZGI-CO-OFDM operates with a zero CP\%

![Fig. 11. Q vs. \( N_{FFT} \) with different deterministic DGD values. The launch power of \(-2\) dBm and the laser linewidth of 100 kHz are assumed.](image)

4. Computation complexity

Computation complexity directly affects the implementation cost and power consumption of digital signal processing hardware [8]. In this section we analyze and compare the computation complexity of the equalization algorithms used in the conventional, RGI- and ZGI-CO-OFDM. Understanding this will allow one to make an intelligent tradeoff between the system performance, overhead and algorithm complexity. Table 2 summarizes the number of the required complex multiplications for the three CO-OFDM systems. Several operations are common to all three systems, which include the FFT, TS-based channel estimation/equalization, and PS-based PN estimation/compensation. A factor of 2 in \( 2 \cdot N \cdot (1 - TS\%) \) and \( 2 \cdot N \cdot TS\% \) is a result of a 2-by-2 matrix multiplication for each polarization; a factor of 4 is used for ZGI-CO-OFDM channel estimation, because we process the same TS twice at the OFDM demodulator. It is worth noting that RGI- and ZGI-CO-OFDM enable the use of smaller \( N_{FFT} \) and \( N_{SC} \) and a smaller TS overhead (TS\%) compared to the conventional CO-
OFDM. TS% is defined as the ratio between the number of TS’s and data symbols within one OFDM frame (see Fig. 1(b)). Given the same time duration for one OFDM frame, RGI- and ZGI-CO-OFDM with a smaller $N_{FFT}$ and $N_{CP}$ can transmit a larger number of data symbols, resulting in a lower TS%. In Fig. 1(b) TS% is 0.67% ( = 2/300) for example, and thus computation effort for TS-based channel estimation is very small.

Table 2. Number of Complex Multiplications for Conventional, RGI-, and ZGI-CO-OFDM for Each Polarization

<table>
<thead>
<tr>
<th></th>
<th>Conventional</th>
<th>RGI-CO-OFDM</th>
<th>ZGI-CO-OFDM</th>
</tr>
</thead>
<tbody>
<tr>
<td>FFT at demodulator</td>
<td>$N_{FFT}/2^{\log N_{FFT}}$</td>
<td>$N_{FFT}/2^{\log N_{FFT}}$</td>
<td>$N_{FFT}/2^{\log N_{FFT}}$</td>
</tr>
<tr>
<td>Channel estimation (TS)</td>
<td>$2 \cdot N_w \cdot TS%$</td>
<td>$2 \cdot N_w \cdot TS%$</td>
<td>$4 \cdot N_w \cdot TS%$</td>
</tr>
<tr>
<td>Channel equalization</td>
<td>$2 \cdot N_w \cdot (1 - TS%)$</td>
<td>$2 \cdot N_w \cdot (1 - TS%)$</td>
<td>$2 \cdot N_w \cdot (1 - TS%)$</td>
</tr>
<tr>
<td>OFDE</td>
<td>n/a</td>
<td>$2 \cdot (N_{NOR} \cdot \log z N_{NOR} + N_{NOR})$</td>
<td>$2 \cdot (N_{NOR} \cdot \log z N_{NOR} + 2 \cdot N_{NOR})$</td>
</tr>
<tr>
<td>FDI (linear)</td>
<td>n/a</td>
<td>n/a</td>
<td>$2 \cdot N_{NOR}$</td>
</tr>
<tr>
<td>PN estimation</td>
<td>$N_w \cdot PS% \cdot (1 - TS%)$</td>
<td>$N_w \cdot PS% \cdot (1 - TS%)$</td>
<td>$N_w \cdot PS% \cdot (1 - TS%)$</td>
</tr>
<tr>
<td>PN compensation</td>
<td>$N_w \cdot (1 - PS%) \cdot (1 - TS%)$</td>
<td>$N_w \cdot (1 - PS%) \cdot (1 - TS%)$</td>
<td>$N_w \cdot (1 - PS%) \cdot (1 - TS%)$</td>
</tr>
</tbody>
</table>

Unlike the conventional, RGI- and ZGI-CO-OFDM requires a computationally intensive OFDE. Specifically, for each polarization in RGI-CO-OFDM, the OFDE performs two $N_{OFDE}$-point FFT’s which require the number of multiplications of $N_{NOR} \cdot \log z N_{NOR}$; and the OFDE performs frequency-domain CD equalization, requiring the number of multiplications of $N_{NOR}$. Thus the total number of multiplications is $2 \cdot (N_{NOR} \cdot \log z N_{NOR} + N_{NOR})$, if we discard $N_{NOR}$ overlapped samples for each $N_{NOR}$ block. Compared to RGI-CO-OFDM, ZGI-CO-OFDM employs a more complex OFDE algorithm, where $2 \cdot N_{NOR}$ multiplications are required for the joint frequency-domain CD and PMD equalization for each polarization. Note that only when a very large oversampling factor of $\gamma = N_{FFT} / N_{OC} > 2$ is used, one can consider performing CD and PMD equalization in two separate steps to reduce the computation effort to $N_{NOR} + 2 \cdot N_{NOR} / \gamma$. This is because the number of 2-by-2 matrices in $H_{FDI}$ is $N_{OC} \cdot (N_{NOR} / N_{FFT}) = N_{NOR} / \gamma$ (see Step 3 in Section 2.1). Moreover, $2 \cdot N_{NOR}$ number of multiplications is required for each polarization, if a simple linear interpolation is used for the FDI.

To compare the computation complexity of three CO-OFDM systems, we evaluate it using the total number of complex multiplications per useful bit. The useful bits are the transmitted bits except those used for overheads. $M$ is used to represent the number of constellation points for each modulated subcarrier, thus each data symbol transmits $N_w \cdot \log z M \cdot (1 - PS\%)$ useful bits, where PS% accounts for the pilot subcarrier overhead. We assume all $N_w$ subcarriers carry the same modulation format. Table 3 summarizes the total number of the complex multiplications per useful bit for three CO-OFDM systems. TS% and PS% are often below 5%, and their contribution to computation complexity can be ignored to obtain a simplified expression. It is apparent from the simplified expressions, the computation complexity depends largely on $N_{NOR}$, $N_{NOR}$ and the oversampling factor $\gamma$. Specifically, we use ZGI-CO-OFDM as an example to illustrate how the number of complex multiplications for each useful bit is calculated. First, the total number of complex multiplications for ZGI-CO-OFDM is calculated by summing up all the operations in Table 2.
\[ \frac{N_{\text{FFT}}}{2} \cdot \log \left( \frac{N_{F_{\text{FFT}}} + 4N_{\omega} - TS\% + 2N_{\omega} \cdot (1 - TS\%) + 2(N_{\text{OFDE}} \cdot \log \frac{N_{\text{OFDE}}}{N_{\text{FFT}}} + 2N_{\text{OFDE}}) + 2N_{\text{OFDE}}}{N_{\omega} \cdot \log M (1 - PS\%) + N_{\omega} \cdot (1 - PS\%) \cdot (1 - TS\%) + 2(N_{\text{OFDE}} \cdot \log \frac{N_{\text{OFDE}}}{N_{\text{FFT}}} + 2N_{\text{OFDE}}) + 2N_{\text{OFDE}}.} \]

Then the total number of complex multiplications for each useful bit will be given as,

\[ \frac{N_{\text{FFT}} \cdot \log \left( 2 \cdot N_{\omega} + 4N_{\omega} - TS\% + 2N_{\omega} \cdot (1 - TS\%) + 2(N_{\text{OFDE}} \cdot \log \frac{N_{\text{OFDE}}}{N_{\text{FFT}}} + 2N_{\text{OFDE}}) + 2N_{\text{OFDE}} \right)}{N_{\omega} \cdot \log M (1 - PS\%) + N_{\omega} \cdot (1 - PS\%) \cdot (1 - TS\%) + 2(N_{\text{OFDE}} \cdot \log \frac{N_{\text{OFDE}}}{N_{\text{FFT}}} + 2N_{\text{OFDE}}) + 2N_{\text{OFDE}}}. \]

Note that in Eq. (10) each FFT block at OFDE processes \( N_{\text{OFDE}}/N_{\text{FFT}} \) number of OFDM symbols, which is accounted by dividing the number of complex multiplications related to OFDE by \( N_{\text{OFDE}}/N_{\text{FFT}} \).

<table>
<thead>
<tr>
<th>Conventional</th>
<th>( \frac{\gamma \log \frac{N_{F_{\text{FFT}}} / 2 + 3 + TS% + 2 \gamma (\log N_{\text{OFDE}} + 2)}{\log M (1 - PS%)}}{\log M} )</th>
<th>( \frac{\gamma \log \frac{N_{F_{\text{FFT}}} / 2 + 3 + TS% + 2 \gamma (\log N_{\text{OFDE}} + 2) + 3}{\log M (1 - PS%)}}{\log M} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>RGI-CO-OFDM</td>
<td>( \frac{\gamma \log \frac{N_{F_{\text{FFT}}} / 2 + 3 + TS% + 2 \gamma (\log N_{\text{OFDE}} + 2) + 3}{\log M (1 - PS%)}}{\log M} )</td>
<td>( \frac{\gamma \log \frac{N_{F_{\text{FFT}}} / 2 + 3 + TS% + 2 \gamma (\log N_{\text{OFDE}} + 2) + 3}{\log M (1 - PS%)}}{\log M} )</td>
</tr>
<tr>
<td>ZGI-CO-OFDM</td>
<td>( \frac{\gamma \log \frac{N_{F_{\text{FFT}}} / 2 + 3 + TS% + 2 \gamma (\log N_{\text{OFDE}} + 6)}{\log M (1 - PS%)}}{\log M} )</td>
<td>( \frac{\gamma \log \frac{N_{F_{\text{FFT}}} / 2 + 3 + TS% + 2 \gamma (\log N_{\text{OFDE}} + 6) + 3}{\log M (1 - PS%)}}{\log M} )</td>
</tr>
</tbody>
</table>

Figure 12(a) illustrates the total number of complex multiplications per useful bit as a function of \( N_{\text{FFT}} \) for the three CO-OFDM systems. We assume \( N_{\text{OFDE}} = 2048 \) and the oversampling factor \( \gamma \) is 1.6, same as those used in the previous calculations. As expected from Section B, the conventional CO-OFDM has significantly smaller computation effort due to its simple estimation and equalization algorithm. In contrast, the computation complexity of RGI- and ZGI-CO-OFDM is governed largely by \( N_{\text{OFDE}} \) and has a relatively weak dependence on \( N_{\text{FFT}} \). For example, ZGI-CO-OFDM with \( N_{\text{FFT}} \) of 128 and 256 results in virtually the same total number of multiplications per bit of 26.7 and 27.1, respectively. Moreover, ZGI-CO-OFDM does not require significantly greater computation effort compared to RGI-CO-OFDM.

Figure 12(b) shows the additional computation effort that ZGI-CO-OFDM requires for different \( N_{\text{OFDE}} \) and oversampling factor \( \gamma \), while \( N_{\text{FFT}} \) is kept at 128. It can be seen that the complexity difference between the two algorithms decreases for smaller \( \gamma \) and larger \( N_{\text{OFDE}} \). Note that \( N_{\text{OFDE}} = 2048 \) is kept for all our previous discussions, because we assume to transmit an OFDM signal over a 1600-km SMF. However, one must increase \( N_{\text{OFDE}} \) proportionally as the fiber reach and CD increase. Specifically, in Section C we use \( N_{\text{OFDE}} = 2048 \) and \( \gamma = 1.6 \), and ZGI-CO-OFDM requires only 13.6% more computation effort than RGI-CO-OFDM does. When \( N_{\text{OFDE}} = 4096 \) (used for 3200-km SMF) and \( \gamma = 1.2 \), the complexity difference will reduce to 11.8%. With this reasonably small complexity difference, one would prefer ZGI-CO-OFDM to attain 0% CP overhead, larger PMD tolerance and improved channel estimation accuracy. Furthermore, for practical implementations of a real-time CO-OFDM receiver, many DSP techniques such as parallel processing and pipelining have been adopted [19,20]. The added complexity by ZGI-CO-
OFDM compared to RGI-CO-OFDM may deviate depending on implementation techniques, and this will be quantified in our future work.

Fig. 12. (a) Number of complex multiplications per useful bit as a function of $N_{\text{FFT}}$ for the conventional, RGI- and ZGI-CO-OFDM. (b) Percentage of extra computation complexity of ZGI- over RGI-CO-OFDM, as a function of $N_{\text{OFDE}}$ and oversampling factor.

5. Conclusion

In this work we present a new channel estimation/equalization algorithm for CO-OFDM to eliminate CP overhead from OFDM data symbols. Because of the zero CP overhead achieved, we term this new CO-OFDM system as zero-guard-interval (ZGI)-CO-OFDM. Unlike the previous RGI-CO-OFDM, ZGI-CO-OFDM employs an OFDE in the digital coherent receiver to compensate both CD and PMD before the final OFDM demodulation. We elaborate the operation principle of our new algorithm and provide comparison with the algorithms used for the conventional and RGI-CO-OFDM. Despite the zero CP overhead, ZGI-CO-OFDM demonstrates excellent PMD tolerance compared to RGI-CO-OFDM, which is verified using a stochastic PMD modeling for 500 different realizations. In addition, ZGI-CO-OFDM demonstrates the ability to achieve more accurate channel estimation through ISFA, and it can effectively improve system $Q$ factor under high PMD conditions. As another system benefit, ZGI-CO-OFDM enables the flexibility of using ever smaller $N_{\text{FFT}}$ (<128) without a concern for increasing the CP overhead, and this ability will be appreciated particularly when the laser PN is large and the frequency synchronization condition is stringent. Finally, we provide an analysis to compare the computational complexity between conventional, RGI- and ZGI-CO-OFDM systems. In particular, the additional computation effort of ZGI-CO-OFDM compared to RGI-CO-OFDM is reasonably small (~13.6%) for 112-Gb/s PDM CO-OFDM over a 1600-km dispersion-uncompensated optical link. Combined with the zero CP overhead and other system benefits demonstrated, ZGI-CO-OFDM would become a practical solution for next-generation high-capacity and long-reach CO-OFDM.