Advanced modulation formats for high-performance short-reach optical interconnects

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Abstract: The explosive growth of the traffic between data centers has led to an urgent demand for high-performance short-reach optical interconnects with data rate beyond 100G per wavelength and transmission distance over hundreds of kilometers. Since direct detection (DD) provides a cost-efficient solution for short-reach interconnects, various advanced modulation formats have been intensively studied to improve the performance of DD for high-performance short-reach optical interconnects. In this paper, we report the recent progress on the advanced DD modulation formats that provide superior electrical spectral efficiency (SE) and transmission reach beyond that of simple direct modulation (DM) based direct detection (DM/DD). We first provide a review of the current advanced modulation formats for high-performance short-reach optical interconnects. Among these formats, Stokes vector direct detection (SV-DD) achieves the highest electrical spectrum efficiency, presenting itself as a promising candidate for future short-reach networks. We then expound some novel algorithms to achieve high-performance SV-DD systems under severe impairments of either polarization mode dispersion (PMD) or polarization dependent loss (PDL).

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References and links


1. Introduction

High-speed fiber optic communication has been extensively explored to keep up with the relentless growth of the Internet traffic. During the last decade, long-haul optical networks have witnessed a capacity advance to multi-Terabit with the revival of coherent communications [1–3]. By employing polarization-diversity transceiver and sophisticated DSP, coherent communication achieves great robustness against the linear fiber impairments caused by chromatic dispersion (CD) and polarization mode dispersion (PMD). In the meantime, short-reach networks are also facing a challenge to upgrade their capacity per wavelength beyond 100 Gb/s to support the growing traffic demand of data center interconnects. Different from long-haul networks, short-reach networks accommodate wide geographic coverage by deploying a massive number of transceivers with reach shorter than hundreds of kilometers. As a result, the transceiver cost becomes the primary consideration for the short-reach network design. The costly polarization-diversity transceiver may prevent coherent detection from the widespread use in short-reach networks. On the contrary, direct detection (DD) requires simpler transceiver structure, making it a promising technology for short-reach networks [4,5].

Direct modulation (DM) based direct detection (DM/DD) is the simplest DD scheme [6]. Since the CD introduces severe nonlinear distortion to the signal due to the square-law photodetection, the transmission distance of DM/DD is limited to tens of kilometers. Furthermore, the phase and polarization dimensions are not utilized by DM/DD, leading to low electrical spectrum efficiency (SE). Aiming to improve the electrical SE and extend the transmission distance, advanced DD modulation formats have been proposed through a so-called self-coherent (SCOH) approach, where a carrier is transmitted together with the signal to achieve a linear mapping between the electrical signal and the optical field [7–11]. In that way, the impact of the CD can be easily removed from the received signal, greatly extending the transmission distance of the DD system.

One of the widely investigated SCOH schemes is single sideband (SSB) modulation format, where the signal is filled at one side of the carrier, enabling the recovery of the complex signal using a single-ended photodiode (PD) [7–9]. To avoid the interference of the signal-to-signal beat noise (SSBN) produced at the square-law detection, guard band or subcarrier interleaving is adopted, leading to the reduced electrical SE [7,8]. Gapless SSB modulation format is then proposed to improve the electrical SE by removing the guard band and employing iterative SSBN cancellation [9]. However, single-polarization SSB modulation format cannot achieve electrical SE higher than 1/2 since it utilizes only half spectrum of the baseband.

The electrical SE is a critical metric to evaluate the performance of short-reach networks. Compared to SSB modulation formats, double sideband (DSB) modulation formats utilize full
spectrum of the baseband, presenting greater potential to achieve higher electrical SE. The technical challenge is to obtain the phase-diverse DSB signal using DD scheme. In the past few years, significant efforts have been made to improve the electrical SE of DD system using DSB modulation formats [12–14]. Block-wise phase shift (BPS) DD and signal carrier interleaved (SCI) DD have been proposed, achieving 1/2 and 2/3 electrical SE compared to single-polarization coherent detection [12,13] (throughout the text, the relative electrical SE is normalized to that of single-polarization coherent detection). Recently, DD with single-polarization modulation achieves 100% electrical SE for the first time by using Stokes vector direct detection (SV-DD) [14]. In this paper, we extend our previous work by presenting novel algorithms to improve the performance of SV-DD systems under severe impairments of PMD and PDL, extending SV-DD to either longer transmission distance or higher Baud rate.

This paper is organized as follows: Section 2 gives a brief background of conventional DD modulation format and the motivation of applying advanced DD modulation formats. Section 3 provides a review of various advanced DD modulation formats in the form of both SSB and DSB. Sections 4 and 5 present study on the mitigation of the PMD and PDL induced impairments in SV-DD systems, respectively. Section 6 gives the comparison of the advanced DD modulation formats. The conclusion is drawn in Section 7.

2. Motivation of advanced modulation formats for direct detection

Conventional DD, referring to DM/DD, has the merits of simple implementation and low cost. As shown in Fig. 1, DM/DD modulates the non-negative signal directly onto a laser, and converts it to the photocurrent by a single-ended PD at the receiver [6]. There are two limiting factors for conventional DD. The first one is the nonlinear mapping between the transmitted baseband signal and the optical field, as well as the nonlinear mapping between the optical field and the received signal after the square-law detection. Due to the nonlinear mapping, the CD causes severe nonlinear distortion to the signal, significantly degrading the performance of the system. Therefore, the conventional DD is only suitable for very short application, where the dispersion is not significant. The second limiting factor is the reduced spectral efficiency. As DM/DD uses single-polarization intensity modulation without utilizing the phase and polarization dimensions, the spectral efficiency is significantly reduced.

Fig. 1. Conceptual diagram of a conventional DD system. PD: photo-detector.

Since the standardization of 100G Ethernet LAN interfaces in 2010, data center interconnects are also facing the increasing challenge to upgrade their capacity. High-performance DD system, with capacity per wavelength at 100 Gb/s and beyond and transmission distance over hundreds of kilometers, is urgently in demand. Various advanced modulation formats have been proposed to achieve this goal. We classify the advanced modulation formats in to two categories: SSB modulation formats and DSB modulation formats, where SSB modulation formats fill the signal into half side of the baseband, while the DSB modulation formats utilize the full band. For both formats, the baseband signal is linearly mapped to the optical field, leading to a linear communication channel under the impact of CD. A main carrier is transmitted together with the signal to down-convert the optical signal to the baseband at the receiver. Since the CD introduces a frequency dependent phase shift to the optical field, the CD induced impairment can be easily compensated by removing the phase shift from the down-converted signal. The linear mapping between the baseband signal and the optical field contributes to high-performance DD system using advanced modulation formats.
3. High-performance advanced modulation formats for direct detection

In this section we will divide the high-performance advanced DD modulation formats into 2 categories: SSB formats and DSB formats. Their principles and transceiver architectures will be discussed in detail.

3.1 Single sideband modulation formats

Fig. 2. SSB generation schemes: (a) offset SSB, and (b) RF tone assisted SSB which can be (i) with guard band, or (ii) subcarrier-interleaving, or (iii) without guard band. S: signal, C: (main) carrier, W: signal bandwidth.

To detect the complex optical signal using a single-ended PD, SSB modulation format fills the signal at one side of the carrier. SSB signal can be generated by different approaches. As shown in Fig. 2(a), offset SSB signal is generated by combing a passband signal and the main carrier at DC [7]. The electrical signal is fed into an intensity modulator, generating output optical signal with Hermitian symmetric spectrum around the main carrier. An optical filter is then used to filter out one side of the spectrum, yielding the SSB signal. Denoting the signal as $S$ and carrier as $C$, the square-law detection produces three beating terms from the SSB signal: (1) the linear signal term $\text{Re}\{SC^*\}$, (2) the nonlinear noise term $SS^*$, and (3) the DC term $CC^*$. To avoid the interference from the nonlinear noise, a guard band is reserved between the carrier and signal that has the same bandwidth as the signal. In that way, the offset SSB signal takes 1/4 spectrum of the baseband, leading to the electrical SE of 1/4 at both transmitter and receiver. The electrical SE at the transmitter can be improved by using RF tone assisted SSB signal, where the RF tone introduces a virtual carrier at the frequency $\omega \neq 0$ [8,9]. As shown in Fig. 2(b), RF tone assisted SSB signal reaches electrical SE of 1/2 at the transmitter when using (i) guard band, or (ii) subcarrier-interleaving [8,9] (see insets (i) and (ii) in Fig. 2(b)), and reaches electrical SE of 100% if no guard band is used (see inset (iii) in Fig. 2(b)). The obtained electrical signal is then fed into an IQ modulator. The electrical spectra at the receiver for different types of SSB signal are shown in Fig. 3. For the methods of using (i) guard band, and (ii) subcarrier-interleaving, the nonlinear noise fills half of the Hermitian symmetric spectrum, leading to the electrical SE of 1/4 at the receiver. For the method (iii) without guard band, the electrical SE is improved to 1/2 while the signal is contaminated by the nonlinear noise. Iterative SSBN cancellation has to be used to remove the nonlinear noise, increasing the DSP complexity. For single-polarization SSB modulation...
format, the electrical SE at the receiver cannot be higher than 1/2 due to the Hermitian symmetry of the baseband spectrum.

Though higher electrical SE could be obtained by employing dual-polarization SSB signal [10,11], the implementation complexity of the transceiver approaches that of polarization multiplexed coherent systems. Electrical SE is an important metric to evaluate the performance of DD system. Considering the limited bandwidth of electronic components, higher electrical SE means higher data rate when the same QAM size is used, or means better receiver sensitivity when smaller QAM size is used to achieve the same data rate. Several DD schemes using DSB modulation formats have been proposed to improve the electrical SE. The technical challenge is to obtain the phase-diverse DSB signal using DD schemes. In the next subsection, we will introduce three novel DD schemes which can successfully solve the problem of phase diversity.

3.2 Double sideband modulation formats

3.2.1 Block-wise phase shift direct detection

BPS-DD uses a transmitter architecture shown in Fig. 4. The DSB signal is modulated on one path using an IQ modulator, and combined with the carrier from the other path. In order to achieve the phase diversity, the DSB signal is repeated in two consecutive time blocks, where a phase shift is introduced between them using three different approaches depicted in Fig. 5 [12].

![Fig. 4. Transmitter structure for block-wise phase switching (BPS) DD.](image)

For the first approach shown in Fig. 5(a), the phase of the carrier $C$ is switched by 90° in two consecutive blocks, while the signal $S$ remains the same. The outputs of the two blocks at the square-law detection are

$$I_1 = |S + C|^2 = |C|^2 + 2 \text{Re}(SC^*) + |S|^2$$

$$I_2 = |S + iC|^2 = |C|^2 + 2 \text{Im}(SC^*) + |S|^2,$$

where $I_1$ and $I_2$ contain the real and imaginary parts of the DSB signal respectively. By combining Eqs. (1) and (2), we have

$$I = I_1 + iI_2 = (1 + i)|C|^2 + 2SC^* + (1 + i)|S|^2,$$
where the fading free signal is provided by the second term $SC^*$ on the right side of the equation. The first term is a DC term, which can be simply ignored. The third term is the nonlinear noise term, by which the linear signal is contaminated. To obtain the linear signal $SC^*$, iterative SSBN cancellation is applied to remove the nonlinear noise from the received signal $I$. High carrier-to-signal power ratio (CSPR) is needed for an effective SSBN cancellation. In the second approach shown in Fig. 5(b), the 90° phase shift is implemented in the signal $S$. The principle of the second approach is the same as the first one.

In the third approach shown in Fig. 5(c), the phase shift is carried out by switching the lower sideband $S_L$ of the signal by 180° and maintaining the upper sideband $S_U$ the same. The outputs of the two consecutive blocks at the square-law detection are

$$I_1 = |S_R + S_L + C|^2 = |C|^2 + 2\text{Re}\{S_R S_L C^*\} + |S_R + S_L|^2$$

$$I_2 = |S_R - S_L + C|^2 = |C|^2 + 2\text{Re}\{(S_R - S_L) C^*\} + |S_R - S_L|^2.$$  

Combining Eq. (4) and (5), we have

$$I_3 = I_1 + I_2 = 2|C|^2 + 4\text{Re}\{S_R C^*\} + 2|S_R|^2 + 2|S_L|^2$$

$$I_4 = I_1 - I_2 = 4\text{Re}\{S_R C^*\} + 4\text{Re}\{S_R S_L^*\}.$$  

The upper and lower sideband of the DSB signal are obtained from $I_3$ and $I_4$ respectively. Similar to the previous approaches, large CSPR and iterative SSBN cancellation are needed to mitigate the nonlinear noise.

For BPS scheme, full baseband is utilized to transmit the phase-diverse signal. Since the information is repeated in in two consecutive blocks, 1/2 electrical SE is achieved by BPS scheme, which is the same as the gapless SSB-DD [8]. By using the first approach in Fig. 5, 49.4 Gb/s direct detection optical OFDM over 80-km standard single-mode fiber (SSMF) has been experimentally demonstrated [12].

3.2.2 Signal-carrier interleaved direct detection

In SCI-DD, the carrier and DSB signal are transmitted in separate time blocks. As shown in Fig. 6, the received signal is first divided into two paths. By introducing one block delay into one path, the signal and carrier blocks are aligned with each other. The two paths are then fed into the inputs of a standard balanced receiver, where the phase-diverse DSB signal gets recovered. The electrical SE of SCI-DD is determined by how the signal and carrier blocks are arranged [13]. In the first approach shown in Fig. 7(a), a carrier block is inserted after each signal block. After the delay at the receiver, Block A contains signal $S_1$ in the upper path and carrier $C$ in the delayed path, while Block B contains carrier $C$ in the upper path and signal $S_1$ in the delayed path. Conjugate outputs, $S_1 C^*$ and $S_1^* C$ are obtained from Block A and B via balanced receiver, leading to the electrical SE of 1/2. The electrical SE can be improved by the second approach, where a carrier block is inserted after every two signal blocks, as shown in Fig. 7(b). After the delay at the receiver, the outputs, $S_1 C^*$, $S_1 S_2^*$, and $S_2^* C$, are obtained from Block A, B and C via the balanced receiver. The DSB signal $S_1$, $S_2$ can be recovered from Block A and C, leading to the electrical SE of 2/3.

Fig. 6. Receiver structure for signal carrier interleaved (SCI) DD.
Fig. 7. Conceptual diagram of signal carrier interleaved (SCI) schemes: (a) SCI-DD with 1/2 SE, and (b) SCI-DD with 2/3 SE.

Beside the improved electrical SE, SCI-DD has another advantage due to the usage of the balanced receiver. The high common mode rejection ratio of the balanced PD makes SCI-DD naturally resistant to the impairment of SSB N. Without the need for SSBN cancellation, DSP becomes simpler, and low CSPR can be used to improve the receiver sensitivity. 102.4-Gb/s direct detection optical OFDM over 80 km SSMF has been successfully demonstrated using the SCI scheme with 2/3 electrical SE in [15].

3.2.3 Stokes vector direct detection

For SV-DD scheme, the DSB signal $S$ and carrier $C$ are transmitted in two orthogonal polarizations as shown in Fig. 8(a) [14]. To recover the signal after the random polarization rotation due to fiber transmission, SV-DD adopts the Stokes vector to represent the polarization state. After the Stokes vector is obtained by a 3-dimensional detection at the receiver, the polarization recovery is then implemented in Stokes space.

Given a polarization state $[E_x | E_y]^T$ in Jones space, the corresponding representation in Stokes space is $[I | Q | S | C]^T$, where $E_x$ and $E_y$ are the electrical fields in the two orthogonal polarizations and the superscript ‘T’ represents transpose [16]. The Stokes vector after fiber transmission is detected using the direct detection scheme shown in Fig. 8(b) [14]. After splitting the two orthogonal polarizations, $X_{out}$ and $Y_{out}$, using a polarization beam splitter (PBS), each polarization is divided into two paths by a 3 dB coupler. One path from $X_{out}$ and $Y_{out}$ is fed into a 90° optical hybrid followed by two balanced detectors (BDs), producing the outputs of $\text{Re}\{X_{out}Y_{out}^*\}$ and $\text{Im}\{X_{out}Y_{out}^*\}$. The other path is fed into a BD directly, producing the output of $|X_{out}|^2 - |Y_{out}|^2$. The three outputs provide us the three elements of the received Stokes vector $[X_{out}^2 | Y_{out}^2 | 2\text{Re}\{X_{out}Y_{out}^*\} | 2\text{Im}\{X_{out}Y_{out}^*\}]^T$.

Assuming the transmitted signal of $[S C]^T$, the received Stokes vector is related to the transmitted Stokes vector $[|S|^2 | |C|^2 | 2\text{Re}\{SC^*\} | 2\text{Im}\{SC^*\}]^T$ by a 3x3 channel matrix $V$, given by

$$
\begin{bmatrix}
|X_{out}|^2 - |Y_{out}|^2 \\
2\text{Re}\{X_{out}Y_{out}^*\} \\
2\text{Im}\{X_{out}Y_{out}^*\}
\end{bmatrix}
= V
\begin{bmatrix}
|S|^2 - |C|^2 \\
2\text{Re}\{SC^*\} \\
2\text{Im}\{SC^*\}
\end{bmatrix},
\quad V = \begin{bmatrix}
h_{11} & h_{12} & h_{13} \\
h_{21} & h_{22} & h_{23} \\
h_{31} & h_{32} & h_{33}
\end{bmatrix},
$$

where $h_{mn}$ is the element of the channel matrix $V$ of $m$-th row and $n$-th column. The channel matrix $V$ indicates the linear effect of fiber channel. Once $V$ is known, the transmitted Stokes vector can be recovered by multiplying $V^{-1}$ to the received Stokes vector.
The channel matrix $V$ can be acquired using polarization training symbols. As shown in Fig. 9, a training period is divided into three slots. For each slot, $S$ is set to be 0, 1 or $i$, while $C$ maintains to be 1. The polarization states of the three slots, $[0\ 1]^T$, $[1\ 1]^T$ and $[i\ 1]^T$, correspond to the Stokes vectors of $[1\ 0\ 0]^T$, $[0\ 1\ 0]^T$ and $[0\ 0\ 1]^T$ (ignoring the constant factors). Plugging them into Eq. (8), the three columns of the channel matrix $[h_{11}\ h_{12}\ h_{13}]^T$ and $[h_{21}\ h_{22}\ h_{32}]^T$ are obtained at the outputs. After recovering the transmitted Stokes vector, $[|S|^2-|C|^2\ 2\text{Re}\{SC^*\}\ 2\text{Im}\{SC^*\}]^T$, we arrive at $SC^*$ by combining the last two elements in the vector, from which the phase-diverse DSB signal is fully developed. The SSBN is moved to the first element of the Stokes vector during the polarization recovery, leaving the signal free from the impact of SSBN. The high-performance of SV-DD is manifested by its two major advantages: (1) SV-DD can be used to detect signal with any polarization state, while achieving high electrical SE of 100%, and (2) the SSBN is automatically cancelled during the polarization recovery, without needing for high CSPR. A 160 Gb/s SV-DD systems over 160 km SSMF has been experimentally demonstrated in [14].

SV-DD provides a novel way to implement the DSP enabled polarization demultiplexing in DD system. The application of this scheme has been extended to the intensity modulation based direct detection (IM/DD) to improve its SE. By transmitting the intensity modulated signal in two polarizations and detecting the signal using SV-DD, polarization division multiplexed (PDM) IM/DD is achieved in [17] with the data rate of 224 Gb/s. By modulating the signal on 4 polarization states and tracking the Stokes vector at the receiver, [18] is able to obtain IM/DD link with the data rate of 128 Gb/s.

4. PMD impact and mitigation in SV-DD systems

PMD effect induces a time delay $\Delta \tau$ between the two principal polarizations during fiber transmission. Under the first-order PMD approximation, the input polarization state $[X_{in}\ Y_{in}]^T$ in Jones space are related by [19]

$$
\begin{bmatrix}
X_{out}(\omega) \\
Y_{out}(\omega)
\end{bmatrix} =
\begin{bmatrix}
a_1 & b_1 \\
a_2 & b_2
\end{bmatrix}
\begin{bmatrix}
e^{j\omega\Delta\tau/2} & 0 \\
0 & e^{-j\omega\Delta\tau/2}
\end{bmatrix}
\begin{bmatrix}
a_1^* & b_1^* \\
-a_2^* & a_2^*
\end{bmatrix}
\begin{bmatrix}
X_{in}(\omega) \\
Y_{in}(\omega)
\end{bmatrix},
$$

(9)

where the first and third matrices at the right side of the equation denote the polarization rotations at the fiber input and output, and the second matrix denotes the differential phase shift $\omega\Delta\tau$ between two principal polarizations caused by the time delay $\Delta \tau$. Equation (9) indicates the frequency dependence of the polarization evolution.

Since the polarization training symbols are constructed by signals at the carrier frequency $\omega_c$, the channel matrix $V(\omega_c)$ obtained from the training symbols represents the polarization evolution at the frequency $\omega_c$. After the polarization recovery, only the polarization state at the frequency $\omega_c$ can get fully recovered. The misalignment of the polarization states between the transmitted signal and the recovered one at the other frequencies causes impairment to SV-DD systems. Denoting the transmitted signal as $S(t) = \sum S_i e^{j\omega_i t}$, where $\omega_i$ and $S_i$ are the baseband frequency and modulation symbol for the $i$-th subcarrier, we calculate the recovered Stokes vector from Eq. (9). Combining the second and third components of the Stokes vector, we arrive at [20]

$$
XY^* = \sum S_i e^{j\omega_i t} \cdot \left( \sum \left[ \frac{\cos(\omega \Delta \tau/2) + j \left( a_i^* - a_i \right) \sin(\omega \Delta \tau/2) S_i e^{j\omega_i t} + 2 j a_i^* b_i}{2} \right] \right) + \left( \sum \left[ \frac{\cos(\omega \Delta \tau/2) + j \left( a_i^* - a_i \right) \sin(\omega \Delta \tau/2) S_i e^{j\omega_i t}}{2} \right] \right) \cdot \sum S_i e^{j\omega_i t}.
$$

(10)
where the first term on the right side of the equation denotes the beating between the recovered signal in $X$ polarization and the carrier in $Y$ polarization, and the second term denotes the beating between the recovered signal in $X$ polarization and the residual signal in $Y$ polarization due to PMD. Under the condition of $|\omega \Delta \tau| < 1$, which is valid in short- and medium-reach optical transmissions, the first and second terms can be simplified as $S(t) = \sum S_i e^{i\omega t}$ and $N_{NL} = F \cdot S(t) \cdot (\sum \omega S_i e^{i\omega t})^*$, where $N_{NL}$ is the PMD-induced nonlinear noise and $F = ja_i b_i \Delta \tau$ is a constant factor determined by the rotation matrix in Eq. (9). When $F$ and $S(t)$ are known, the PMD induced noise $N_{NL}$ can be estimated and subtracted from the recovered signal. We propose two different algorithms to estimate the PMD induced noise $N_{NL}$. The flow charts of algorithms A and B are shown in Figs. 10 and 11.

![Flow chart of algorithm A](image1)

**Fig. 10.** Flow chart of algorithm A. (a) Calculate $F$ using training symbols. (b) Mitigate the PMD induced noise in data symbols.

![Flow chart of algorithm B](image2)

**Fig. 11.** Flow chart of algorithm B. (a) Calculate $F$ using training symbols. (b) Mitigate the PMD induced noise in data symbols.

For both algorithms, specially-designed training symbols are transmitted to calculate the factor $F$, where only the odd subcarriers are filled with data, the even subcarriers are left empty. For the received training symbols, the first term of Eq. (10) falls into the odd subcarriers, while the second term falls into the even subcarriers. As shown in the flow charts of Figs. 10(a) and 11(a), after performing FFT for the received training symbols, the first and second terms of Eq. (10) are obtained by separating the odd and even subcarriers. Different approaches are adopted by algorithms A and B when calculating $F$. In algorithm A shown in Fig. 10(a), $\omega S_i$ is calculated in the frequency domain. $S_i$ and $\omega S_i$ are then converted to $S(t)$ and $\sum \omega S_i e^{i\omega t}$ by IFFT. By comparing $S(t) \cdot (\sum \omega S_i e^{i\omega t})^*$ with the second term, $F$ is obtained. In algorithm B shown in Fig. 11(a), signal in the frequency domain $S_i$ is directly converted to $S(t)$ by IFFT. Then, $S(t)$ is advanced and delayed with one sampling interval $\Delta T$ to obtain $S(t + \Delta T)$ and $S(t - \Delta T)$. Subtracting $S(t + \Delta T)$ from $S(t - \Delta T)$, we obtain
\begin{align*}
S(t - \Delta T) - S(t + \Delta T) &= \sum_j S_j e^{j\omega_j t} e^{-j\omega_j \Delta T} - \sum_j S_j e^{j\omega_j t} e^{j\omega_j \Delta T} = -2j \sum_j \sin(\omega_j \Delta T) S_j e^{j\omega_j t} . (11)
\end{align*}

Assuming that \(\sin(\omega_j \Delta T) = \alpha \omega_j \Delta T\) under the condition of \(|\omega_j \Delta T| \leq \pi/2\), where \(\alpha\) is a constant, and \(|\omega_j \Delta T| \leq \pi/2\) is satisfied when the oversampling rate is larger than or equals to 2, the second term in Eq. (10) can be rewritten as \(F/(2j\alpha \omega) \cdot S(t) \cdot (S(t - \Delta T) - S(t + \Delta T))'\). By comparing \(S(t) \cdot (S(t - \Delta T) - S(t + \Delta T))'\) with the second term, we obtain \(F/(2j\alpha \omega)\). The PMD induced noise in the data symbols is mitigated following the flow charts of Figs. 10(b) and 11(b). In algorithm A, FFT is performed to calculate \(\omega_i S_i\), and then IFFT is performed to obtain \(F / (2 \alpha \omega) \cdot S(t) \cdot (S(t - \Delta T) - S(t + \Delta T))'\). In algorithm B, \(F / (2 \alpha \omega) \cdot S(t) \cdot (S(t - \Delta T) - S(t + \Delta T))'\) is obtained directly from the time domain signal \(S(t)\). Though the assumption of \(\sin(\omega_j \Delta T) = \omega_j \Delta T\) may induce slight inaccuracy during the noise estimation, algorithm B does not need an additional pair of FFT/IFFT for PMD mitigation in contrast with algorithms A. This leads to a much efficient DSP for algorithm B.

![Fig. 12. Experimental setup to verify the PMD mitigation algorithms.](image)

We use the two proposed algorithms to mitigate the PMD induced impairment in a 93-Gb/s SVDD system. The experiment setup is illustrated in Fig. 12. A three-tone generator is applied in the signal path to obtain a broader optical bandwidth. 16 QAM OFDM signal is generated by an arbitrary waveform generator (AWG) and modulated onto the three tones. The FFT size is 4096 with the central 3420 subcarriers filled with data. A total of 100 data symbols are transmitted, where the first 4 symbols are used as the training symbols by setting the even subcarriers null. 128 point of cyclic prefix (CP) is added into each symbol. The polarization training symbols are added before the OFDM frame. The AWG operates at a sampling rate of 10 GSa/s, leading to an optical bandwidth of 8.35 GHz for one tone, and 25 GHz for three tones. The raw data rate is 4x25 = 100 Gb/s. Counting the training symbols and CP, the data rate is reduced to 93 Gb/s. The length of the carrier path is matched with the signal path to eliminate the phase noise. The CSPR is maintained at 0 dB. The signal is transmitted over an 80-km fiber. To adjust the PMD in the transmission link, a PMD emulator (FiberPro PE4200) is placed at the input of the fiber. A polarization controller (PC) is inserted before the PMD emulator to control the input polarization angle. In our experiment, the angle is set to be 45°, corresponding to the worst impairment. The Stokes vector is detected using a 90° optical hybrid and three BDs at the receiver. The electrical signal is sampled by an oscilloscope at a sampling rate of 50 GSa/s. The collected data is processed offline. The PMD induced impairment is to be mitigated using the two proposed algorithms after the polarization recovery.
Fig. 13. Q penalty as a function of DGD before and after PMD mitigation.

Fig. 14. Signal and noise spectra (a) before PMD mitigation, (b) after PMD mitigation using algorithm A, and (c) after PMD mitigation using algorithm B.

The system performances before and after applying the algorithms are compared. Considering the small differential group delay (DGD) (smaller than 0.36 ps) in the 80-km fiber, the DGD value in the transmission link is dominated by the PMD emulator with adjustable DGD ranging from 0 to 10 ps. The Q penalty is measured as the PMD-induced Q degradation when the initial Q is set at 16.5 dB. The Q penalty as a function of DGD is plotted in Fig. 13. According to Eq. (10), the PMD induced noise increases rapidly with the DGD value. We can observe a significant increase of Q penalty with the increase of DGD before PMD mitigation in Fig. 13. The Q penalty exceeds 1 dB (the dashed line in Fig. 13) when DGD is larger than 3.5 ps and reaches 3.5 dB at the DGD of 10 ps. After mitigating the PMD induced impairment using algorithms A and B, the Q penalty is greatly reduced. The Q penalty reaches 1 dB at DGD of 9.5 ps using algorithm A, and at DGD of 9 ps using algorithm B. Algorithms A and B extend the system’s PMD tolerance from 3.5 to 9.5 and 9 ps, respectively. Figure 14(a) shows the spectra of signal and PMD-induced noise at DGD of 10 ps before applying the algorithms; Figs. 14(b) and (c) show the spectra after applying algorithms A and B. The greatly reduced noise level in Figs. 14(b) and (c) demonstrates the effectiveness of the two proposed algorithms in mitigating the PMD induced impairment. Furthermore, algorithm B is almost as effective as algorithm A. Considering their DSP complexity, algorithm B would be the preferred choice for the practical implementation.

5. PDL impact and mitigation in SV-DD systems

The asymmetric structure of optical components and the polarization dependent gain of erbium-doped fiber amplifiers (EDFAs) lead to unequal insert losses for two polarization modes, or so-called polarization-dependent loss (PDL). Assuming the attenuation difference between the two principal polarizations is $2\alpha$, the input polarization state $[S\ C]^T$ and output polarization state $[X_{out}\ Y_{out}]^T$ in Jones space are related by [21]

$$
\begin{bmatrix}
X_{out} \\
Y_{out}
\end{bmatrix} = \begin{bmatrix}
a_2 \\
-b_2
\end{bmatrix} \begin{bmatrix}
\sqrt{1+\alpha} & 0 \\
0 & \sqrt{1-\alpha}
\end{bmatrix} \begin{bmatrix}
a_1 \\
b_1
\end{bmatrix} \begin{bmatrix}
S \\
C
\end{bmatrix}. \tag{12}
$$
where the first and third matrices on the right side of the equation are the rotation matrices, similar to those in Eq. (9), and the second matrix indicates the differential loss between two principle polarizations due to PDL. The PDL value in decibel is $\Gamma = 10 \log_{10} \left( (1 + \alpha) / (1 - \alpha) \right)$.

Converting the output vector $[X_{out} \ Y_{out}]^T$ into Stokes space, the three elements of the output Stokes vector can be calculated from Eq. (12) as

$$[X_{out}^2 - Y_{out}^2] = A_1 |S|^2 + B_1 |C|^2 + C_1 \text{Re}\{SC^*\} + D_1 \text{Im}\{SC^*\}$$

$$\text{Re}\{X_{out}^* Y_{out}^*\} = A_2 |S|^2 + B_2 |C|^2 + C_2 \text{Re}\{SC^*\} + D_2 \text{Im}\{SC^*\}$$

$$\text{Im}\{X_{out}^* Y_{out}^*\} = A_3 |S|^2 + B_3 |C|^2 + C_3 \text{Re}\{SC^*\} + D_3 \text{Im}\{SC^*\}.$$  \hspace{1cm} (13)

(14)

(15)

where symbols, $A_n$, $B_n$, $C_n$ and $D_n$, are determined by the PDL coefficient $\alpha$ and the elements of rotation matrices in Eq. (12). The expressions of $A_n$, $B_n$, $C_n$ and $D_n$ are given in the appendix.

When no PDL exists in the system, $\alpha = 0$, we have $A_n = -B_n$ ($n = 1, 2, 3$). The first and second terms in Eqs. (13), (14) and (15) can be combined as $A_n (|S|^2 - |C|^2)$.

Therefore, the 3x3 channel estimation, described in section 3.2.3, fails to acquire the channel information accurately. Impairment is invalidity of the linear relationship of Eq. (8). Therefore, the 3x3 channel estimation, by the 3x3 channel matrix in Eq. (8). On the other hand, when PDL exists in the system, we arrive at $[A_n] \neq [B_n]$. In that case, the $|S|^2$ and $|C|^2$ terms cannot be combined, leading to the invalidity of the linear relationship of Eq. (8). Therefore, the 3x3 channel estimation, described in section 3.2.3, fails to acquire the channel information accurately. Impairment is induced to SV-DD systems during the polarization recovery.

We propose two different approaches to mitigate the impact of PDL. In the first approach, we adopt 4x4 channel estimation instead of 3x3 channel estimation. We separate the first element of the Stokes vector, $|E_r|^2 - |E_i|^2$, into 2 individual elements, $|E_r|^2$ and $|E_i|^2$, and extend the 3-dimensional Stokes vector to a 4-dimensional vector $[|E_r|^2 \ |E_i|^2 \ 2\text{Re}\{|E_r|E_i^*\} \ 2\text{Im}\{|E_r|E_i^*\}]^T$. To detect the 4-dimensional vector, the ‘BD1’ in Fig. 8(b) is replaced with two single-ended PDs. The input 4-dimensional vector $[|S|^2 \ |C|^2 \ 2\text{Re}\{|SC^*\} \ 2\text{Im}\{|SC^*\}]^T$ and output 4-dimensional vector $[|X_{out}|^2 \ |Y_{out}|^2 \ 2\text{Re}\{|X_{out}|Y_{out}^*\} \ 2\text{Im}\{|X_{out}|Y_{out}^*\}]^T$ are related by a 4x4 channel matrix $H$

$$\begin{bmatrix} |X_{out}|^2 \\ |Y_{out}|^2 \\ \text{Re}\{X_{out}^* Y_{out}^*\} \\ \text{Im}\{X_{out}^* Y_{out}^*\} \end{bmatrix} = H \begin{bmatrix} |S|^2 \\ |C|^2 \\ \text{Re}\{SC^*\} \\ \text{Im}\{SC^*\} \end{bmatrix}, \hspace{1cm} H = \begin{bmatrix} h_{11} & h_{12} & h_{13} & h_{14} \\ h_{21} & h_{22} & h_{23} & h_{24} \\ h_{31} & h_{32} & h_{33} & h_{34} \\ h_{41} & h_{42} & h_{43} & h_{44} \end{bmatrix},$$  \hspace{1cm} (16)

where $h_{nm}$ denotes the element of the channel matrix $H$ at the $n$-th row and $m$-th column. By extending the Stokes vector to 4 dimensions, linear communication channel can be obtained even at the presence of PDL. To acquire the 4x4 channel matrix $H$, specially-designed polarization training symbols are used. A training symbol is divided into 4 slots, where the signal $S$ is set to be 0, 1, -1 and $i$ respectively, while the carrier $C$ remains to be 1. The 4-dimensional vectors for the 4 slots are $T_1 = [0 1 0 0]^T$, $T_2 = [1 1 1 0]^T$, $T_3 = [1 1 -1 0]^T$ and $T_4 = [1 1 0 1]^T$. Taking them to Eq. (16), we obtain the vectors $R_1$, $R_2$, $R_3$ and $R_4$ at the output. The columns of the 4x4 channel matrix can be calculated by

$$[h_{11} \ h_{21} \ h_{31} \ h_{41}]^T = (R_2 + R_3) / 2 - R_1$$

$$[h_{12} \ h_{22} \ h_{32} \ h_{42}]^T = R_1$$

$$[h_{13} \ h_{23} \ h_{33} \ h_{43}]^T = (R_2 - R_3) / 2$$

\hspace{1cm} (17)

\hspace{1cm} (18)

\hspace{1cm} (19)
\[ [h_{14} \ h_{24} \ h_{34} \ h_{44}]^T = R_4 - (R_2 + R_1)/2. \] (20)

After the 4x4 channel matrix is obtained, the rest DSP procedures are the same as for 3-dimensional Stokes vector.

In the second approach, the 3x3 channel training described in Section 3.2.3 is still applied, while additional algorithm is used to mitigate the impairment caused by PDL. For a better understanding of the impact of PDL, we rewrite Eqs. (13), (14) and (15) as

\[ \begin{align*}
|X_{\text{out}}| - |Y_{\text{out}}| &= A_1(|S|^2 - |C|^2) + (A_1 + B_1)|C|^2 + C_1 \text{Re}\{SC^*\} + D_1 \text{Im}\{SC^*\} \quad (21) \\
\text{Re}\{X_{\text{out}}Y_{\text{out}}^*\} &= A_2(|S|^2 - |C|^2) + (A_2 + B_2)|C|^2 + C_2 \text{Re}\{SC^*\} + D_2 \text{Im}\{SC^*\} \quad (22) \\
\text{Im}\{X_{\text{out}}Y_{\text{out}}^*\} &= A_3(|S|^2 - |C|^2) + (A_3 + B_3)|C|^2 + C_3 \text{Re}\{SC^*\} + D_3 \text{Im}\{SC^*\}. \quad (23)
\end{align*} \]

From Eqs. (21), (22) and (23), we can see that PDL introduces terms \((A_n + B_n)|C|^2\) (n = 1, 2, 3) into the three elements of the received Stokes vector, disturbing the linear relationship between the input and output Stokes vector, where \(A_n, B_n\) are constant numbers determined by Eq. (12), and \(|C|^2 = 1\) for SV-DD scheme. By estimating the DC terms \((A_n + B_n)|C|^2\) and removing them from the received Stokes vector, the linear relationship between the input and output Stokes vectors can be recovered. Training symbols with the input polarization states \([S]\) and \([-1 \ 1]^T\) are used to estimate the DC terms. Taking the training symbols into Eqs. (21), (22) and (23), the output Stokes vectors for the training symbols are \([A_1 + B_1 \ C_1] \ [A_2 + B_2 \ C_2] \ [A_3 + B_3 \ C_3]^T\) and \([A_1 + B_1 \ C_1] \ [A_2 + B_2 \ C_2] \ [A_3 + B_3 \ C_3]^T\). Subtracting their sum \([A_1 + B_1 \ C_1] \ [A_2 + B_2 \ C_2] \ [A_3 + B_3 \ C_3]^T\) from the received Stokes vector, the linear relationship of Eq. (8) is recovered. After that, the channel training and the polarization recovery would be the same as described in Section 3.2.3.

The proposed PDL mitigation algorithms are evaluated by simulation. The simulation uses a similar setup as illustrated in Fig. 12, where the multi-band generator is removed, and PMD emulator is replaced by PDL emulator. 16 QAM OFDM signal is generated and modulated onto the CW laser source. The FFT size of OFDM signal is 4096, in which 2048 subcarriers are filled with signal. Different polarization training symbols for algorithms A and B are added before the OFDM frame. The sampling rate of DAC is set to be 50 GSa/s, resulting in a signal bandwidth of 25 GHz. Counting the CP and training symbols, the overall data rate for the described system is 95 Gb/s. The input polarization angle of the PDL emulator is set to be \(\pi/4\), corresponding to the worst case. At the receiver, the signal is detected by 4-dimensional detection for algorithm A and by 3-dimensional detection for algorithm B. After the signal is received, algorithms A and B are carried out for the linear channel estimation and polarization recovery.

![Fig. 15. Q penalty as a function of DGD before and after PMD mitigation.](image)
measured by the Q penalty. The Q penalty as a function of the PDL, ranging from 0 to 2 dB, is plotted in Fig. 15. As explained in Eqs. (21), (22) and (23), PDL introduces DC terms into the three elements of the received Stokes vector, leading to inaccurate polarization recovery. Without applying PDL mitigation algorithm, the Q penalty increases dramatically with the PDL, and reaches 4.5 dB at PDL of 2 dB. On the other hand, the Q penalty maintains around 0 dB after either algorithm A or B is applied. The simulation result shows that both algorithms A and B can efficiently eliminate the PDL induced errors during the polarization recovery. Algorithm B is the preferred choice due to its simplicity without an additional PD.

6. Comparison of DD schemes using advanced modulation formats

<table>
<thead>
<tr>
<th>Scheme</th>
<th>E-SE</th>
<th>Mod.</th>
<th>PD</th>
<th>Required OSNR @ 10^−3 BER</th>
<th>Additional DSP</th>
</tr>
</thead>
<tbody>
<tr>
<td>Offset SSB</td>
<td>25%</td>
<td>1 IM</td>
<td>1</td>
<td>18 dB for 10Gb/s 8QAM [9]</td>
<td></td>
</tr>
<tr>
<td>Virtual SSB</td>
<td>50%</td>
<td>1 IQ-M</td>
<td>1</td>
<td>13 dB for 10Gb/s 4QAM [9]</td>
<td>Iterative SSBN canc.</td>
</tr>
<tr>
<td>BPS-DD</td>
<td>50%</td>
<td>1 IQ-M</td>
<td>1</td>
<td>32 dB for 49.4 Gb/s 8QAM [12]</td>
<td>Iterative SSBN canc.</td>
</tr>
<tr>
<td>SCI-DD</td>
<td>66.7%</td>
<td>1 IQ-M</td>
<td>2 B-PD</td>
<td>22 dB for 43.2 Gb/s 16 QAM [13]</td>
<td></td>
</tr>
<tr>
<td>SV-DD</td>
<td>100%</td>
<td>1 IQ-M</td>
<td>3 B-PD</td>
<td>27 dB for 160 Gb/s 16 QAM [14]</td>
<td>3x3 real MIMO</td>
</tr>
</tbody>
</table>

The performance and complexity of DD schemes using advanced modulation formats are compared with the polarization multiplexed coherent detection in Table 1. Obviously, coherent detection has the best performance. The high performance of coherent detection is based on its expensive polarization-diversity transceiver and sophisticated DSP. On the contrary, cost-efficient transmission system can be obtained by using DD schemes. However, their performances are sacrificed by their simple transceiver and DSP. DD schemes using single-ended PD have the lowest electrical SE. Since a guard band has to be reserved to accommodate the nonlinear noise produced at single-ended PD, offset SSB only achieves the electrical SE of 25%. The electrical SE is improved to 50% by virtual SSB and BPS-DD, where iterative SSBN cancellation is employed to eliminate the nonlinear noise. After turning to balanced receiver, SCI-DD achieves the electrical SE of 66.7%. The highest electrical SE is achieved by SV-DD with its 3-dimensional detection.

7. Conclusion

Aiming to improve the electrical spectral efficiency and transmission reach of DD systems, various advanced modulation formats have been proposed for high-performance short-reach optical interconnects. Among these schemes, SV-DD achieves the highest electrical SE (100% SE compared to the single-polarization coherent detection), making it a promising candidate for future short-reach optical networks. Additionally, the performance of SV-DD systems can be impaired by the PMD and PDL effects. Novel algorithms have been proposed in this paper to mitigate the impairments caused by PMD and PDL. By applying these algorithms, high-performance SV-DD can be achieved even under severe impairments of PMD and PDL.

Appendix

The coefficients $A_n$, $B_n$, $C_n$ and $D_n$ ($n=1, 2, 3$) in Eqs. (13), (14) and (15) are given below:


\[ A_i = \left( |a_i|^2 - |b_i|^2 \right) (|a_i|^2 (1 + \alpha) - |b_i|^2 (1 - \alpha)) - 2(a_i b_i^* a_i b_i + a_i^* b_i a_i^* b_i^*) \sqrt{(1 + \alpha)(1 - \alpha)} \]

\[ B_i = \left( |a_i|^2 - |b_i|^2 \right) (|b_i|^2 (1 + \alpha) - |a_i|^2 (1 - \alpha)) + 2(a_i b_i^* a_i b_i + a_i^* b_i a_i^* b_i^*) \sqrt{(1 + \alpha)(1 - \alpha)} \]

\[ C_i = [2(|a_i|^2 - |b_i|^2)(a_i b_i^* + a_i^* b_i) + 2(a_i b_i^*(a_i^2 - b_i^2) + a_i^* b_i(a_i^2 - b_i^2)) \sqrt{(1 + \alpha)(1 - \alpha)}] \]

\[ D_i = [2(|a_i|^2 - |b_i|^2)(a_i b_i^* - a_i^* b_i) + 2(a_i b_i^*(a_i^2 + b_i^2) + a_i^* b_i(a_i^2 + b_i^2)) \sqrt{(1 + \alpha)(1 - \alpha)}] \]
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