# Joint estimation and compensation of transceiver IQ imbalance for PDM CO-OFDM system*

**MA Xiu-rong (马秀荣)**, YANG Meng (杨梦), SHAN Yun-long (单云龙), and LI Jing-jing (利晶晶)**

*Engineering Research Center of Optoelectronic Devices and Communication Technology, Ministry of Education, Tianjin University of Technology, Tianjin 300384, China*

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In this paper, a frequency-domain mathematical model is deduced for polarization-division multiplexing (PDM) coherent optical orthogonal frequency division multiplexing (CO-OFDM) system with transceiver in-phase and quadrature (IQ) imbalances. A novel training symbol structure is designed in which the mirror-subcarriers of pilot subcarriers were modulated with zero signal so that the channel distortion with transceiver IQ imbalances can be estimated. It proves that the channel distortion and transmitter IQ imbalances cannot be separated using the training symbols; therefore, the channel equalization method was used to recover the signals. Simulation results show that at 2 dB transmitter amplitude imbalances, 15° transmitter phase imbalances, 100 Gbit/s transmission rate, 1 040 km standard signal-mode fiber link, and $1\times10^{-3}$ bit error rate, 23.5 dB optical signal-to-noise ratio (*OSNR*) is necessary for the proposed method. However, the compared schemes cannot achieve effective transmission.

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Coherent optical orthogonal frequency division multiplexing (CO-OFDM) has shown its extreme robustness against chromatic dispersion (CD) and polarization mode dispersion (PMD) in optical communication[1-4]. However, polarization-division multiplexing (PDM) CO-OFDM system is sensitive to non-idealities in-phase and quadrature (IQ) imbalance in the transmitter (Tx) and receiver (Rx) sides which is caused by non-ideal 90° phase shift and amplitude imbalance between I and Q branches[5-8]. A generalized training symbol is proposed based hybrid (time and frequency)-domain Tx and Rx IQ imbalance calibration algorithm in which two same training symbols are used for Rx IQ compensation in time domain, and the same ones for channel estimation and Tx IQ calibration in frequency domain in Ref.[9]. However, the imperfect first process estimation value result will degrade the second process estimation performance. Ref.[10] proposed a blind estimation scheme to estimate the transmitter IQ imbalances based on the combination of analytical modeling and clustering algorithm. Ref.[11] proposed a simplified Gram-Schmidt orthogonalization process (SG) method based on the Gram-Schmidt orthogonalization procedure (GSOP) to compensate IQ imbalance, but SG method is not able to suppress the IQ imbalance completely, and the channel distortion estimation would be affected by the residual IQ imbalance. In Ref.[12], a training symbol structure is proposed for PDM CO-OFDM system with receiver IQ imbalance. Combining with the GSOP method, the channel distortion can be estimated using the training symbols.

In this paper, we built the frequency-domain mathematical model of PDM CO-OFDM system in presence of transceiver IQ imbalances. A novel training symbol structure is designed, in which the mirror-subcarriers of pilot subcarriers were modulated with zero signal so that the channel distortion with transceiver IQ imbalances can be estimated. The plots subcarriers spacing *M* is 2-subcarriers so that channel distortion with transceiver IQ imbalances of $(N-2)/M$, $M$=2 pilot subcarriers can be estimated. The channel distortion and transceiver IQ imbalance cannot be estimated respectively; therefore, channel equalization was used to recover the signals. The performance of SG method in Ref.[11], training symbols in Ref.[12] and the proposed training symbols are numerically simulated for 100 Gbit/s PDM CO-OFDM with different transmission scenarios, respectively.

Fig.1 shows a typical schematic diagram of PDM CO-OFDM system in direct up/down conversion architecture. Ignoring the noise, the received signal $R_{x/y}^{t,r}(k,i)$ in the presence of transceiver IQ imbalances can be written in frequency domain as Eq.(1).

$$\begin{pmatrix} R_x^{t,r}(k,i) \\ R_y^{t,r}(k,i) \end{pmatrix} = \begin{pmatrix} H_{1xx}^{IQ}(k) \\ H_{1xy}^{IQ}(k) \end{pmatrix} \left( S_x(k,i) \right) +$$

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** E-mail: lunwentoudi@163.com

$$\begin{pmatrix} H_{2xx}^{IQ}(k) \\ H_{2xy}^{IQ}(k) \end{pmatrix}\left(\bar{S}_x^*(k,i)\right)+\begin{pmatrix} H_{1yx}^{IQ}(k) \\ H_{1yy}^{IQ}(k) \end{pmatrix}\left(S_y(k,i)\right)+$$
$$\begin{pmatrix} H_{2yx}^{IQ}(k) \\ H_{2yy}^{IQ}(k) \end{pmatrix}\left(\bar{S}_y^*(k,i)\right). \quad (1)$$

Here,

$$\begin{aligned}
&H_{1xx}^{IQ}(k)=G_{1tx}G_{1rx}H_{xx}(k)+G_{2tx}^*G_{2rx}\bar{H}_{xx}^*(k),\\
&H_{1xy}^{IQ}(k)=G_{1tx}G_{1ry}H_{xy}(k)+G_{2tx}^*G_{2ry}\bar{H}_{xy}^*(k),\\
&H_{1yx}^{IQ}(k)=G_{1ty}G_{1rx}H_{yx}(k)+G_{2ty}^*G_{2rx}\bar{H}_{yx}^*(k),\\
&H_{1yy}^{IQ}(k)=G_{1ty}G_{1ry}H_{yy}(k)+G_{2ty}^*G_{2ry}\bar{H}_{yy}^*(k),\\
&H_{2xx}^{IQ}(k)=G_{2tx}G_{1rx}H_{xx}(k)+G_{1tx}^*G_{2rx}\bar{H}_{xx}^*(k),\\
&H_{2xy}^{IQ}(k)=G_{2tx}G_{1ry}H_{xy}(k)+G_{1tx}^*G_{2ry}\bar{H}_{xy}^*(k),\\
&H_{2yx}^{IQ}(k)=G_{2ty}G_{1rx}H_{yx}(k)+G_{1ty}^*G_{2rx}\bar{H}_{yx}^*(k),\\
&H_{2yy}^{IQ}(k)=G_{2ty}G_{1ry}H_{yy}(k)+G_{1ty}^*G_{2ry}\bar{H}_{yy}^*(k),\\
&G_{1tx/1ty}=\frac{1+\varepsilon_{tx/ty}\mathrm{e}^{\mathrm{j}\phi_{tx/ty}}}{2},\ G_{2tx/2ty}=\frac{1-\varepsilon_{tx/ty}\mathrm{e}^{\mathrm{j}\phi_{tx/ty}}}{2},\\
&G_{1rx/1ry}=\frac{1+\varepsilon_{rx/ry}\mathrm{e}^{-\mathrm{j}\phi_{rx/ry}}}{2},\ G_{2rx/2ry}=\frac{1-\varepsilon_{rx/ry}\mathrm{e}^{\mathrm{j}\phi_{rx/ry}}}{2},\\
&\bar{S}_{x/y}^*(k,i)=S_{x/y}^*\left(\mathrm{mod}\left((N-k),N\right),i\right),0\le k\le N-1,\\
&\bar{H}_{xx}^*(k)=H_{xx}^*\left(\mathrm{mod}\left((N-k),N\right)\right),0\le k\le N-1,\\
&\bar{H}_{yx}^*(k)=H_{yx}^*\left(\mathrm{mod}\left((N-k),N\right)\right),0\le k\le N-1,\\
&H_{xy}^*(k)=H_{xy}^*\left(\mathrm{mod}\left((N-k),N\right)\right),0\le k\le N-1,\\
&\bar{H}_{yy}^*(k)=H_{yy}^*\left(\mathrm{mod}\left((N-k),N\right)\right),0\le k\le N-1, \quad (2)
\end{aligned}$$

where $(\cdot)^*$ represents conjugation, mod(($N$−$k$), $N$) represents the remainder of ($N$−$k$)/$N$. $x$ and $y$ denote polarization states. $N$ is the total number of OFDM subcarriers, $i$ and $k$ represent the $i$th symbol and $k$th subcarrier. $S_{x/y}(k,i)$ are modulation signals without IQ imbalance. $\varepsilon_{tx/ty}$ ($\varepsilon_{rx/ry}$) is the Tx (Rx) amplitude imbalance between I and Q channels, while $\phi_{tx/ty}$ ($\phi_{rx/ry}$) is the Tx (Rx) phase imbalance between I and Q channels. The $H_{ab}(k)$ represents channel frequency response from polarization state $a$ to polarization state $b$ in the $k$th subcarrier.

According to Eq.(1), in order to recover the modulation signals $S_{x/y}(k,i)$, channel distortions with IQ imbalance expressed as Eq.(2) need to be estimated which include eight independent parameters: channel distortion parameters $H_{xx}(k)$, $H_{yx}(k)$, $H_{xy}(k)$, $H_{yy}(k)$ $\bar{H}_{xx}^*(k)$, $\bar{H}_{xy}^*(k),\bar{H}_{yx}^*(k),\bar{H}_{yy}^*(k)$ and IQ imbalance factors $G_{1tx}$, $G_{1rx}$, $G_{1ty}$, $G_{1ry}$. In ideal case of transceiver IQ imbalance, $\varepsilon_{tx/ty}=\varepsilon_{rx/ry}=1$ and $\phi_{tx/ty}=\phi_{rx/ry}=0°$[13], therefore, $H_{1xx}^{IQ}(k)=H_{xx}(k)$ , $H_{1xy}^{IQ}(k)=H_{xy}(k)$ , $H_{1yx}^{IQ}(k)=H_{yx}(k)$ , $H_{1yy}^{IQ}(k)=H_{yy}(k)$ , $H_{2xx}^{IQ}(k)=0$ , $H_{2xy}^{IQ}(k)=0$ , $H_{2yx}^{IQ}(k)=0$ , and $H_{2yy}^{IQ}(k)=0$ .

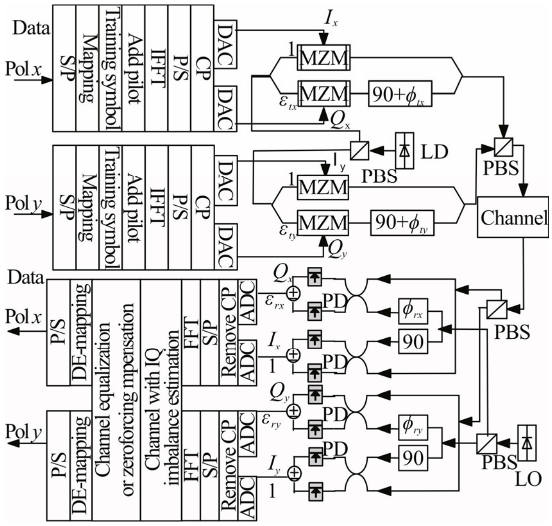


**Fig.1 Schematic diagram of PDM CO-OFDM system**

In order to estimate the channel distortions with IQ imbalance in Eq.(2), a novel zero-mirror-subcarri-ers training symbol structure is proposed shown in Fig.2. Two training symbols are inserted at the beginning of the OFDM frame in each polarization which satisfy Eq.(3). The subcarrier $k$=$N$/2 is mirror subcarrier. $p_x(k,1)$ and $p_y(k,2)$ are known modulation training symbols for polarization states $x$ and $y$.

$$\begin{cases} S_x(k,1)=\begin{cases} p_x(k,1), & k\in A \\ 0, & k\in\bar{A}\cup B \end{cases} \\ S_y(k,1)=0 \end{cases}, \quad (3a)$$

$$\begin{cases} S_x(k,2)=0 \\ S_y(k,2)=\begin{cases} p_y(k,2), & k\in A \\ 0, & k\in\bar{A}\cup B \end{cases} \end{cases}, \quad (3b)$$

where

$$\begin{aligned}
&A=\left\{2,4,6,\ldots,N/2-2\right\}\cup\\
&\left\{N/2+1,N/2+3,\ldots,N-1\right\},\\
&\bar{A}=\left\{1,3,5,\ldots,N/2-1\right\}\cup\\
&\left\{N/2+2,N/2+4,\ldots,N-2\right\}, \quad (3c)\\
&B=\left\{0,N/2\right\}.
\end{aligned}$$

As Eq.(3a) and Eq.(3b) show, the plots subcarriers spacing $M$ in subset $A$ is 2-subcarriers and the number pilots subcarriers in subset $A$ is $N_A$=($N$−2)/$M$, $M$=2.

Substituting Eq.(3) into Eq.(1), $S_x(k,1)=0$ can avoid the interference cross-interferences from $y$ polarization state to $x$ polarization state. $S_y(k,2)=0$ can avoid the interference cross-interferences from $x$ polarization state to $y$ polarization state. $S_x(k,1)=0$, $k\in\bar{A}$ can avoid the interference from the mirror-image-$k$th $S_x^*(k,1)$, $k\in\bar{A}$ to $R_x^{t,r}(k,1)$, $k\in A$. $S_y(k,2)=0$, $k\in\bar{A}$ can avoid the interference from the mirror-image-$k$th $S_y^*(k,2)$, $k\in\bar{A}$ to $R_y^{t,r}(k,2)$, $k\in A$. So that the received training symbols in both polarization states can be written as:

$$\begin{cases} R_x^{t,r}(k,1)=H_{1xx}^{IQ}(k)\,p_x(k,1) \\ R_y^{t,r}(k,1)=H_{1xy}^{IQ}(k)\,p_x(k,1) \end{cases} \quad k\in A\,, \tag{4a}$$

$$\begin{cases} R_x^{t,r}(k,1)=H_{2xx}^{IQ}(k)\,\bar{p}_x^*(k,1) \\ R_y^{t,r}(k,1)=H_{2xy}^{IQ}(k)\,\bar{p}_x^*(k,1) \end{cases} \quad k\in \bar{A}\,, \tag{4b}$$

$$\begin{cases} R_x^{t,r}(k,2)=H_{1yx}^{IQ}(k)\,p_y(k,2) \\ R_y^{t,r}(k,2)=H_{1yy}^{IQ}(k)\,p_y(k,2) \end{cases} \quad k\in A\,, \tag{4c}$$

$$\begin{cases} R_x^{t,r}(k,2)=H_{2yx}^{IQ}(k)\,\bar{p}_y^*(k,2) \\ R_y^{t,r}(k,2)=H_{2yy}^{IQ}(k)\,\bar{p}_y^*(k,2) \end{cases} \quad k\in \bar{A}\,. \tag{4d}$$

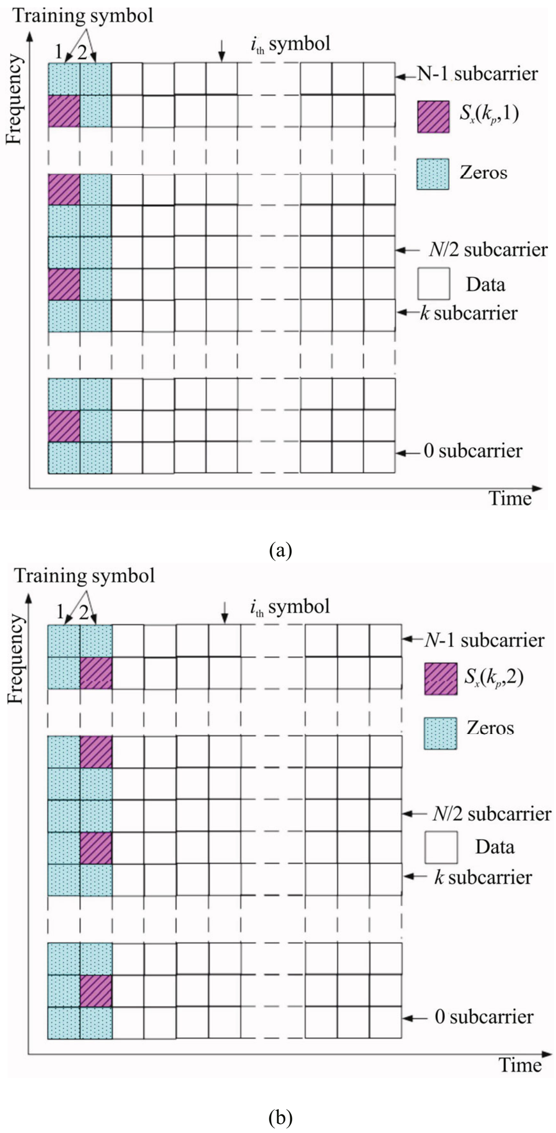


**Fig.2 The designed training symbols for (a) *x* polarization state and (b) *y* polarization state**

From Eq.(4), the channel distortion with transceiver IQ imbalances can be estimated as

$$\begin{cases} \hat{H}_{1xx}^{IQ}(k)=\dfrac{R_x^{t,r}(k,1)}{p_x(k,1)},\hat{H}_{1xy}^{IQ}(k)=\dfrac{R_y^{t,r}(k,1)}{p_x(k,1)} \\ \hat{H}_{1yx}^{IQ}(k)=\dfrac{R_x^{t,r}(k,2)}{p_y(k,2)},\hat{H}_{1yy}^{IQ}(k)=\dfrac{R_y^{t,r}(k,2)}{p_y(k,2)} \end{cases} \quad k\in A \quad ,\tag{5a}$$

$$\begin{cases} \hat{H}_{2xx}^{IQ}(k)=\dfrac{R_x^{t,r}(k,1)}{\bar{p}_x^*(k,1)},\hat{H}_{2xy}^{IQ}(k)=\dfrac{R_y^{t,r}(k,1)}{\bar{p}_x^*(k,1)} \\ \hat{H}_{2yx}^{IQ}(k)=\dfrac{R_x^{t,r}(k,2)}{\bar{p}_y^*(k,2)},\hat{H}_{2yy}^{IQ}(k)=\dfrac{R_y^{t,r}(k,2)}{\bar{p}_y^*(k,2)} \end{cases} \quad k\in \bar{A} \quad .\tag{5b}$$

After getting the channel distortion of $N_A=(N-2)/M$, $M=2$ subcarriers, we can use the method of cubicspline interpolation to get all the channel information with transceiver IQ imbalance[14].

The Eq.(4) shows that only 8 independent equations can be got using the proposed training symbols. Therefore, 12 independent unknown parameters cannot be estimated respectively. Any other training symbol structure can only get the linear combinations of these 8 independent equations in Eq.(4). In order to recover the modulation signals $S_x(k,i)$ and $S_y(k,i)$, channel equalization method was used to compensate channel distortion with transceiver IQ imbalances shown in Eq.(6).

$$\hat{S}(k,i)=\begin{pmatrix} \hat{H}_{1xx}^{IQ}(k) & \hat{H}_{2xx}^{IQ}(k) & \hat{H}_{1yx}^{IQ}(k) & \hat{H}_{2yx}^{IQ}(k) \\ \hat{H}_{1xy}^{IQ}(k) & \hat{H}_{2xy}^{IQ}(k) & \hat{H}_{1yy}^{IQ}(k) & \hat{H}_{2yy}^{IQ}(k) \end{pmatrix}^{-1} \times \begin{pmatrix} R_x^{t,r}(k,i) \\ R_y^{t,r}(k,i) \end{pmatrix}, \tag{6}$$

where

$$\hat{S}(k,i)= \left(\hat{S}_x(k,i) \quad \hat{\bar{S}}_x^*(k,i) \quad \hat{S}_y(k,i) \quad \hat{\bar{S}}_y^*(k,i)\right)^{\mathrm{T}}. \tag{7}$$

To investigate the performance of the SG method in Ref.[11], training symbols (TS I) in Ref.[12] and the proposed training symbols (TS II), Monte-Carlo simulation of a PDM CO-OFDM system is built using MATLAB R2016b and OptiSystem 7.0 for 100 Gbit/s 4-QAM signal. For each polarization, I and Q channels of the OFDM signals were generated by MATLAB R2016b with sampling rate of 25 Gbit/s. The IQ modulation and demodulation as well as the transmission fiber link are achieved by OptiSystem 7.0. Tx and Rx amplitude imbalances are dominated in the signals of Q channel by optical attenuators. Tx and Rx phase imbalances are dominated in the signals of Q channel by phase shifters. Each frame consists of 400 OFDM data symbols. The OFDM data frame is designed as follows. The signals have IFFT and FFT size of 256 points. 128 subcarriers are modulation with among 12.5 GHz bandwidth. 32 cyclic prefix (CP) is inserted into every OFDM symbol. The transmission fiber link consisted of re-circulating loops which has 80 km long SSMF with dispersion parameter of 16.75 ps/ns/km, PMD coefficient of 0.5 ps/$\sqrt{\text{km}}$, attenuation parameter of 0.2 dB/km, nonlinearity with self-phase modulation, and erbium-doped fiber amplifier (EDFA) with 16 dB gain and 6 dB noise figure. The optical dispersion is compensated[15].

Fig.3 shows the bit error rate (*BER*) performance of the SG method, TS I and TS II versus optical signal-to-noise ratio (*OSNR*) between 16 dB and 25 dB. It can be observed that the proposed TS II outperforms the SG method and TS I. It owns to that the proposed scheme can estimate the channel distortion with IQ imbalances of $(N-2)/M=127$, $N=256$, $M=2$ subcarriers, larger than that in Ref.[12] which is $|N/M|=85$, $N=256$, $M=3$ so that the accuracy of channel distortion estimation can be improved by TS II. Moreover, the SG method cannot completely eliminate the influence of IQ imbalances. At *BER* of $1\times10^{-3}$, when transceiver amplitude imbalances are 2 dB, transceiver phase imbalances are 15°, and the *OSNR* of TS II is 23.5 dB, while SG method and TS I cannot achieve the effective transmission; when Tx amplitude imbalances are 2 dB, Tx phase imbalances are 15°, with no Rx IQ imbalance, and the *OSNR* of TS I and TS II are 24.1 dB and 22.8 dB, while SG method cannot achieve the effective transmission; when Rx amplitude imbalances are 2 dB, Rx phase imbalances are 15°, with no Tx IQ imbalance, and the *OSNR* of TS I and TS II are 24.5 dB and 23.3 dB, while SG method cannot achieve the effective transmission.

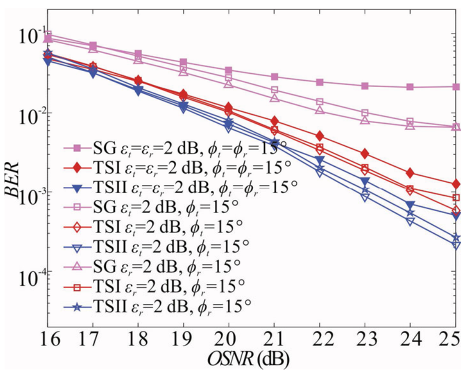

**Fig.3 *BER* versus *OSNR* over 1 040 km transmission at $\varepsilon_{tx}=\varepsilon_{ty}=\varepsilon_{rx}=\varepsilon_{ry}=2$ dB and $\phi_{tx}=\phi_{ty}=\phi_{rx}=\phi_{ry}=15°$**

To verify accuracy of channel distortion estimation of SG method, TS I and TS II, we tested the mean squared error (*MSE*) performance versus *OSNR*. The *MSE* was calculated by Eq.(8). As Fig.4 shows, at the same *OSNR*, the *MSE* of the TS II is smaller than that of SG method and TS I, which means that the proposed TS II can reduce the estimation error of channel distortion with IQ imbalances.

$$MSE=\frac{1}{N_f N}\sum_{i=0}^{N_f-1}\sum_{k=0}^{N-1}\left|\hat{S}_{x/y}(k,i)-S_{x/y}(k,i)\right|^2, \tag{8}$$

where $\hat{S}_{x/y}(k,i)$ are the recovered modulation signals, $S_{x/y}(k,i)$ are the transmitted modulation signals, $N_f$ is the number of OFDM symbols, and $N$ is the total number of OFDM subcarriers.

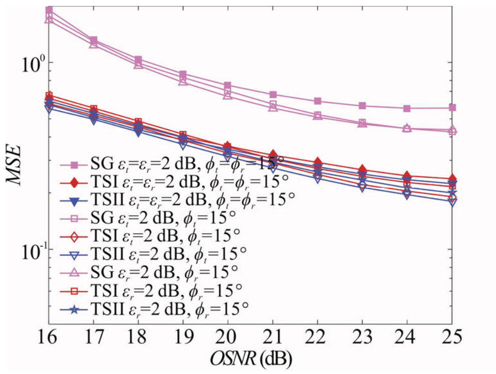

**Fig.4 *MSE* versus *OSNR* over 1 040 km transmission**

Fig.5 shows the *BER* performance of the SG method, TS I and TS II versus transceiver amplitude imbalances. At *BER* of $1\times10^{-3}$, with no transceiver phase imbalances, SG method, TS I and TS II can tolerate 0.4 dB, 1.7 dB and 2.3 dB amplitude imbalances, respectively; when transceiver phase imbalances are 20°, TS I and TS II can tolerate 0.8 dB and 1.7 dB amplitude imbalances, respectively, while SG method cannot tolerate any amplitude imbalance; when transceiver phase imbalances are 25°, TS II can tolerate 0.2 dB amplitude imbalances, while SG method and TS I cannot tolerate any amplitude imbalance.

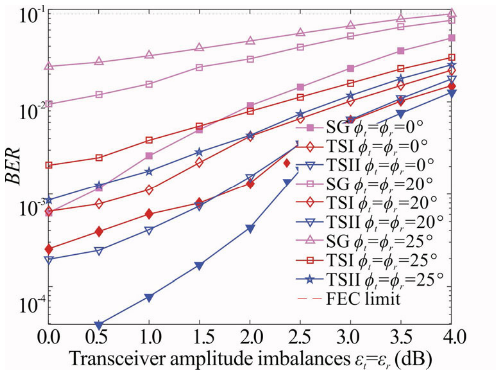

**Fig.5 *BER* versus transceiver amplitude imbalances over 1 040 km transmission at 2 dBm laser power**

Fig.6 shows the *BER* performance of the SG method, TS I and TS II versus transceiver phase imbalances. At *BER* of $1\times10^{-3}$, with no transceiver amplitude imbalances, SG method, TS I and TS II have 4.5°, 22° and 25.5° phase imbalances tolerance, respectively; when transceiver phase imbalances are 1 dB, TS I and TS II have 18.5° and 23° amplitude imbalances tolerance, respectively, while SG method cannot tolerate any phase imbalance; when transceiver amplitude imbalances are 2 dB, TS II has 15.5° phase imbalances tolerance, while SG method and TS I cannot tolerate any amplitude imbalance.

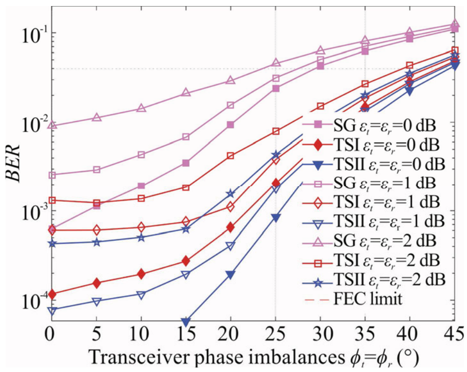


**Fig.6 *BER* versus transceiver phase imbalances over 1 040 km transmission at 2 dBm laser power**

Fig.7 shows the *BER* performance of the SG method, TS I and TS II versus fiber length. At *BER* of $1\times10^{-3}$, when transceiver amplitude imbalances are 2 dB, transceiver phase imbalances are 15°, TS I and TS II can achieve 850 km and 1 100 km transmission, respectively, while SG method cannot achieve the effective transmission; when Tx amplitude imbalances are 2 dB, Tx phase imbalances are 15°, with no Rx IQ imbalance, SG method, TS I and TS II can achieve 420 km, 1 050 km and 1 250 km transmission, respectively; when Rx amplitude imbalances are 2 dB, Rx phase imbalances are 25°, with no Tx IQ imbalance, SG method, TS I and TS II can achieve 570 km, 1 000 km and 1 200 km transmission, respectively.

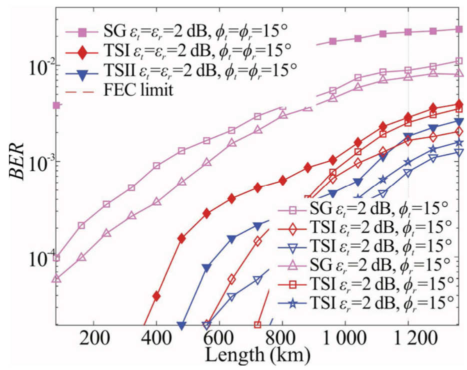


**Fig.7 *BER* versus fiber length at 2 dBm laser power**

In this paper, we built the mathematical model of PDM CO-OFDM system with transceiver IQ imbalance. A novel training symbol structure is designed, in which the plots subcarriers spacing $M$ is 2-subcarriers so that channel distortion with transceiver IQ imbalances of $(N-2)/M$, $M$=2 pilot subcarriers can be estimated. Simulation results show that compared with the reported SG method and the zero-mirror-subcarriers training symbols with 3-subcarriers subcarriers spacing, the proposed training symbols can reduce the *OSNR* punishment, improve the transmission length, enhance the transmitter IQ imbalances tolerance of the PDM CO-OFDM system.

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