# Phase Noise Self-Cancellation Scheme Based on Orthogonal Polarization for OFDM System

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## Abstract

In orthogonal frequency-division multiplexing (OFDM) systems, phase noise introduced by the local oscillators can cause bit error rate (BER) performance degradation. To solve the phase noise problem, a novel orthogonal-polarization-based phase noise self-cancellation (OP-PNSC) scheme is proposed. First, the efficiency of canceling the phase noise of the OP-PNSC scheme in the AWGN channel is investigated. Then, the OP-PNSC scheme in the polarization-dependent loss (PDL) channel is investigated due to power imbalance caused by PDL, and a PDL pre-compensated OP-PNSC (PPC -OP-PNSC) scheme is proposed to mitigate the power imbalance caused by PDL. In addition, the performance of the PPC-OP-PNSC scheme is investigated, where the signal-to-interference-plus-noise ratio (SINR) and spectral efficiency (SE) performances are analyzed. Finally, a comparison between the OP-PNSC and polarization diversity scheme is discussed. The numerical results show that the BER and SINR performances of the OP-PNSC scheme outperform the case with the phase noise compensation and phase noise self-cancellation scheme.

***Keywords:*** orthogonal polarization, phase noise, self-cancellation, PDL, OFDM

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## 1. Introduction

Currently, orthogonal frequency-division multiplexing (OFDM) is chosen for several standards, including IEEE802.11, LTE/LTE-A, Full-Duplex, and 5G, due to its high spectral efficiency and robustness to multi-path interference. However, OFDM systems are very sensitive to phase noise caused by non-ideal local oscillators (LOs) [1], [2]. The distortion of phase noise is characterized by a common phase error (CPE) term and an inter-carrier interference (ICI) term. The CPE term represents the common rotation of all constellation points, and the ICI term breaks the orthogonality of the sub-carriers. In an OFDM system, phase noise leads to increased received signal error floor and bit error ratio (BER) performance degradation [3].

Frequency offset also causes ICI in OFDM system. Base on frequency offset cancellation [4]-[6], many works in the literature have attempted to cancel out phase noise in an OFDM system. We focus on phase noise cancellation schemes which can be divided into two categories: the phase noise compensation and phase noise self-cancellation schemes. Phase noise compensation schemes employ, for example, the pilot inserting method, the cyclic prefix (CP) method, or the blind estimation method, to estimate phase noise and then compensate for the distortion of phase noise at the receiver. In the pilot inserting method, phase noise is estimated using pilots embedded in the OFDM symbols and then corrected by the receiver [3], [7]-[10]. However, the spectral efficiency (SE) of this method decreases due to the extra pilots for phase noise compensation. Thus, the CP method [11] and the blind estimation method [12] are proposed and shown to increase the SE due to extra pilots not being used. The CP method only employs the CP of the inter-symbol interference (ISI)-free symbols, and the blind estimation method employs the training sequence to mitigate the distortion of phase noise. Although these phase noise compensation schemes are efficient in canceling out phase noise, they suffer from a high computational complexity, which is always 3-4 times that of FFT. To reduce the computational complexity, a phase noise self-cancellation scheme[13]-[15] is proposed. In contrast to phase noise compensation schemes, in the phase noise self-cancellation scheme, each data symbol is transmitted using two adjacent sub-carriers. Thus, the phase noise coefficients of the received signals from two adjacent sub-carriers are approximately equal, and the received signals are combined to suppress ICI. This method has the advantage of low complexity; however, the method reduces the SE by half, and residual phase noise still exists because the phase noise coefficients are not exactly the same.

In contrast to the time and frequency domains, the polarized domain, as a new dimension for signal processing, is used to combat phase noise due to the redundancy from the orthogonality provided by polarization, which can maintain the computational complexity and SE. In optical communication, a phase noise suppression scheme with orthogonal polarization transmission in an OFDM system is proposed [16]. The independent orthogonally polarized signals provide the redundancy to further increase the performance of the phase noise cancellation. Considering a realistic wireless channel, our previous work proposes a CPE cancellation scheme with differential polarization shift keying [17], which multiplies two components of the polarized signal to cancel out the phase rotation of the received signal caused by phase noise. However, regardless, in optical and wireless communications, depolarization has not been investigated. In a realistic wireless channel, the polarized signals are distorted by depolarization, polarization-dependent loss (PDL), polarization mode dispersion (PMD) [18] and cross polarization discrimination (XPD) [19]. XPD is the power coupling between different polarizations, which can be canceled out by, for example, zero-forcing or pre-coding methods. OFDM is insensitive to PMD in the non multi-path channel. In contrast, the PDL effect causes power imbalance of the received signals with different polarizations, resulting in phase noise performance degradation. In this paper, we investigate the phase noise cancellation in the polarized domain in the ideal and PDL channel. Moreover, we further extend our works to phase noise cancellation including both CPE and ICI.

In this paper, we propose an orthogonal polarization based phase noise self-cancellation (OP-PNSC) scheme for OFDM systems. The OP-PNSC scheme obtains the redundant from the orthogonal polarization by orthogonal dual polarized antennas to provide the same phase noise coefficients. Based on these coefficients, adding together the orthogonal polarized signals at the receiver can cancel out the distortion of phase noise, including CPE and ICI. This paper first investigates the phase noise cancellation performance of the OP-PNSC scheme in the AWGN channel. Then, the distortion of PDL in the practical wireless channel is investigated, and a PDL pre-compensated OP-PNSC (PPC-OP-PNSC) scheme is proposed under perfect channel state information (CSI) conditions. In the performance evaluation, the signal-to-interference-plus-noise ratio (SINR) and SE performance are analyzed to evaluate the performance of the OP-PNSC scheme. Moreover, the comparison with the polarization diversity (PD) scheme combined with the OFDM system is discussed because both of them have similar antenna architectures. Finally, the simulation results show that the SINR and BER performances of the OP-PNSC scheme are better than those of the phase noise compensation and phase noise self-cancellation schemes. Comparing to the phase noise self-cancellation scheme, the orthogonality of the polarizations enables a situation such that the orthogonal polarized signals are independent and transmitted in the same frequency channel. Thus, half of the spectral resources are saved, and the SE performance is twice that of the phase noise self-cancellation scheme.

Section 2 briefly describes system model. Then, the OP-PNSC scheme is proposed, and the CPE and ICI cancellation performances are determined in the AWGN channel in Section 3. Moreover, the effect of PDL in the OP-PNSC scheme is determined, and the performance of the PPC-OP-PNSC scheme is presented. The OP-PNSC scheme is analyzed in Section4 in terms of the SINR and SE. A comparison with the polarization diverstiy scheme is discussed in this section. The numerical results and performance comparison of the different schemes are given in Section 5. Finally, conclusions are drawn in Section 6.

## 2. System Model

In this section, we introduce the considered OFDM system equipped ODPAs at the transceiver, and the channel model.

OFDM system equipped ODPAs is illustrated in **Fig. 1**. In **Fig. 1**, the two branches at the transceiver connect to the horizontal and vertical polarized antennas to process the polarized signals. Fortunately, ODPAs are widely used in wireless communications such as 5G [20].

At the transmitter, the data signals $S'[k]$, mapped by data information, are sent to the upper branch and lower branche as shown in **Fig. 1**. Let $S_x[k]$ and $S_y[k]$ be the frequency domain transmitted signals for $k$-th sub-carrier in the upper ($x$) and lower ($y$) branches, and $S_x[k]=S_y[k]=S'[k]$, to facilitate the analysis. In each branch the signals, $S_x[k]$ and $S_y[k]$, convert from the frequency domain to the time domain via OFDM modulator, including guide insertion (GI), serial to parallel (S/P), inverse fast Fourier transform (IFFT), parallel to serial

(P/S), cyclic prefix (CP) and digital to analog converter (DAC). After IFFT, the OFDM transmitted signals in the time domain can be written as follows:

$$\boldsymbol{s}[n]=\begin{bmatrix} s_x[n] \\ s_y[n] \end{bmatrix}=\begin{bmatrix} \frac{1}{\sqrt{N}}\sum_{k=0}^{N-1} S_x[k]e^{j\frac{2\pi nk}{N}} \\ \frac{1}{\sqrt{N}}\sum_{k=0}^{N-1} S_y[k]e^{j\frac{2\pi nk}{N}} \end{bmatrix} \tag{1}$$

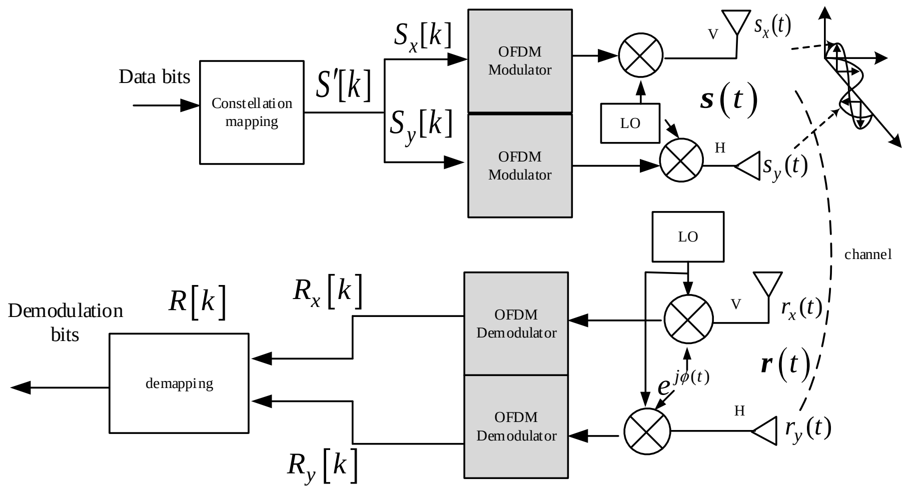


**Fig. 1.** the overview of OFDM system equipped ODPAs

where $N$ is the number of sub-carriers. To mitigate the ISI caused by multi-path, a CP of length $N_{CP}$ is prepended to $\boldsymbol{s}[n]$, where $N_{CP}$ is larger than or equal to the length of the channel impulse response (CIR). Then, the time-continuous analog signals $\boldsymbol{s}(t)=\left[\mathbf{P}_x s_x[t] \quad \mathbf{P}_y s_y[t]\right]^{\mathrm{T}}$ with orthogonal polarizations from $\boldsymbol{s}[n]$ are transmitted by ODPAs. The two polarizations are orthogonal if their inner product is zero, i.e., $\mathbf{P}_x$ is orthogonal to $\mathbf{P}_y$ iff $\mathbf{P}_x \cdot \mathbf{P}_y = 0$ [18]. Then, the orthogonal polarized time-continuous analog signals passed through the multi-path channel are disturbed by phase noise and AWGN which can be written as follows:

$$\boldsymbol{r}(t)=e^{j\phi(t)}\boldsymbol{h}(t)*\boldsymbol{s}(t)+\boldsymbol{w}(t)=e^{j\phi(t)}\boldsymbol{h}(t)*\begin{bmatrix} \mathbf{P}_x s_x(t) \\ \mathbf{P}_y s_y(t) \end{bmatrix}+\begin{bmatrix} w_x(t) \\ w_y(t) \end{bmatrix} \tag{2}$$

where $*$, $\phi(t)$ and $\boldsymbol{h}(t)$ denote the convolution operation, phase noise generated in the receiver LO and the depolarization channel, respectively. Note that we only consider phase noise caused by the LO at the receiver in this paper due to the transmitter is usually considered as the base station which can be well equipped and the performance of the LO at the transmitter can be assumed ideal. Thus, $\boldsymbol{w}(t)=[w_x(t) w_y(t)]^T$ are the AWGN vectors with independent and identically distributed (i.i.d.) entries for each polarization with zero mean and variance $\sigma^2$.

At the receiver, after polarization matching receiving by ODPAs, the polarized vector signals in each branch will be converted to the received signals $r_x(t)$ and $r_y(t)$. It is assumed that the received time domain signals are distorted by the same phase noise from one LO. After OFDM demodulator, including sampling, CP removing, and under the assumption of frequency and clock perfect synchronization, the time domain received signals are given by

$$\boldsymbol{r}[n] = \begin{bmatrix} r_x[n] \\ r_y[n] \end{bmatrix} = e^{j\phi(nT_s)} \begin{bmatrix} \boldsymbol{h}_x[n] \otimes \mathbf{P}_x s_x[n] \\ \boldsymbol{h}_y[n] \otimes \mathbf{P}_y s_y[n] \end{bmatrix} + \begin{bmatrix} w_x[n] \\ w_y[n] \end{bmatrix} \tag{3}$$

where $T_s$ is the sample time and $\otimes$ denotes the circular convolution. $\boldsymbol{h}_x[n]$ and $\boldsymbol{h}_y[n]$ are the CIR in the polarizations $x$ and $y$ after sampling, respectively. Let $a[k] = \frac{1}{N}\sum_{n=0}^{N-1} e^{j\phi(nT_s)}$. Then, the FFT is applied on $\boldsymbol{r}[n]$, $n = 0,1,...,N-1$ to obtain, $\boldsymbol{R}[k]$, $k = 0,1,...,N-1$, by

$$\begin{aligned} \boldsymbol{R}[k] &= a[k] \otimes \begin{bmatrix} \boldsymbol{H}_x[k] \\ \boldsymbol{H}_y[k] \end{bmatrix} \odot \begin{bmatrix} \mathbf{P}_x S_x[k] \\ \mathbf{P}_y S_y[k] \end{bmatrix} + \begin{bmatrix} W_x[k] \\ W_y[k] \end{bmatrix} = a[k] \otimes \boldsymbol{H}[k] \odot \boldsymbol{S}[k] + \boldsymbol{W}[k] \\ &= \underbrace{a[0]}_{\text{CPE}} \boldsymbol{H}[k] \odot \boldsymbol{S}[k] + \underbrace{\sum_{l=1}^{N-1} a[l] \boldsymbol{H}[(k-l)_N] \odot \boldsymbol{S}[(k-l)_N]}_{\text{ICI}} + \boldsymbol{W}[k], \end{aligned} \tag{4}$$

where $(k-l)_N$ represents $k-l \bmod N$ which stands for the interference of the $k$-th sub-carrier from the other sub-carriers and. $\odot$ denotes the Hadamard product. $\boldsymbol{H}_x[k]$ and $\boldsymbol{H}_y[k]$ are the channel frequency response (CFR) in the $x$ and $y$ polarization. $W_x[k]$ and $W_y[k]$ are the FFT version of the AWGN. Because the distortion of phase noise is multipli- cative, in this paper, the phase noise of the CPE and ICI terms are called the CPE coefficient and the ICI coefficient, respectively, and the FFT coefficients combined with phase noise are called phase noise coefficients.

In (4), we can seen that the CFR distorts the received signals. Thus, the channel model is introduced in the frequency domain. For a wireless transmitter which is equipped with the orthogonal dual polarized antennas, the depolarization channel in the frequency domain can be written as follows:

$$\tilde{\boldsymbol{H}}[k] = \begin{bmatrix} H_{HH}[k] & H_{HV}[k] \\ H_{VH}[k] & H_{VV}[k] \end{bmatrix} \tag{5}$$

where $H_{XY}(k)$ represents the channel gain between the $Y$ transmitted antenna and the $X$ received antenna in the $k$-th sub-carrier. Then $\tilde{\boldsymbol{H}}[k]$ can be decomposed using the singular value decomposition (SVD) as

$$\tilde{\boldsymbol{H}}[k]=\boldsymbol{U}\boldsymbol{\Sigma}\boldsymbol{V}^{*}=\boldsymbol{U}\begin{bmatrix}\sqrt{\lambda_{1,k}} & 0\\ 0 & \sqrt{\lambda_{2,k}}\end{bmatrix}\boldsymbol{V}^{*} \quad (6)$$

where $\boldsymbol{U}$ and $\boldsymbol{V}^*$ are $2\times2$ unitary matrices (i.e $\boldsymbol{U}\boldsymbol{U}^H=\boldsymbol{I}$ ) which cause the constellation rotation in the Poincare sphere, but do not change the orthogonality of the two PSs and the amplitude of the received signals in the two branches. Some methods can compensate this rotation [23]. $\boldsymbol{\Sigma}$ is the diagonal matrix. $\sqrt{\lambda_{1,k}}$ and $\sqrt{\lambda_{2,k}}$ $\left(\lambda_{1,k}\ge\lambda_{2,k}\ge0\right)$ are the maximum and minimum eigenvalues of the matrix $\boldsymbol{C}=\tilde{\boldsymbol{H}}[k]*\tilde{\boldsymbol{H}}^H[k]$, respectively. $(\cdot)^H$ represents the conjugated transpose of the matrix. PDL can be repressed as

$$PDL_k=10\lg(\lambda_{1,k}/\lambda_{2,k}) \quad (7)$$

It can be seen that the PDL value is related to $\lambda_{1,k}$ and $\lambda_{2,k}$ which are depend on $\boldsymbol{C}$ .

## 3. The Orthogonal Polarization based Phase Noise Self-Cancellation Scheme

To cancel out the distortion of phase noise in OFDM system, a novel orthogonal polarizations based phase noise self-cancellation scheme is proposed. The orthogonal polarizations provide the redundancy to cancel out phase noise with conjugation and adding processing. Thus, phase noise in the two branches generated within a group can be self-cancelled each other. In this section, the OP-PNSC scheme will be investigated and the phase noise cancellation performance is evaluated in terms of the CPE and ICI cancellation in the AWGN and PDL channel.

### 3.1 The OP-PNSC Scheme in the AWGN Channel

As mentioned in Section 2, the orthogonal polarizations can guarantee two signals from the two branches transmitted in the same frequency channel. However, at the receiver, the signals in the each branch are distorted by phase noise. So, the conjugation units are employed at the transceiver to guarantee that the desired signals in each branch are the same and the phase noise coefficients are conjugated at the receiver. The signals in the $y$ branch are conjugated.

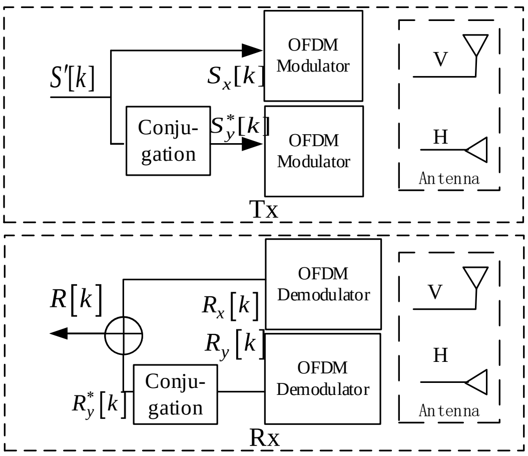


**Fig. 2.** the OP-PNSC scheme with conjugation and ODPAs at the transceiver

As shown in **Fig. 2**, the signals at the two branches from one resource are mutual conjugated at the transmitter, and the transmitted signals in the $y$ branch after mapping are conjugated, $S_y^*[k]=\left[S'[k]\right]^*$. $(\cdot)^*$ denotes the conjugation operation. Note that in the AWGN channel, the CFR is $\boldsymbol{H}_x[k]=\boldsymbol{H}_y[k]=\tilde{\boldsymbol{H}}[k]=\begin{bmatrix}1 & 0\\ 0 & 1\end{bmatrix}$. In the AWGN channel, we do not consider the depolarization. After matching receiving by ODPAs, only the received scale signals in the frequency domain according to (4) are given by

$$\boldsymbol{R}[k]=\begin{bmatrix}\sum_{l=0}^{N-1} a[l]S_x[(k-l)_N]\\ \sum_{l=0}^{N-1} a[l]S_y^*[(k-l)_N]\end{bmatrix}+\begin{bmatrix}W_x[k]\\ W_y[k]\end{bmatrix} \tag{8}$$

To recover the desired signals in the $y$ branch, the frequency domain received signals in the $y$ branch need to be conjugated,

$$R_y^*[k]=a^*[0]S_y[k]+\sum_{l=1}^{N-1} a^*[l]S_y[(k-l)_N]+W_y^*[k], \tag{9}$$

After conjugation, the desired signals in the $x$ and $y$ branches are the same. Then, the decision variables is obtained by adding processing as

$$R^{OP}[k]=\frac{1}{2}(R_x[k]+R_y^*[k]) \tag{10}$$

Then, based on the decision variables of the proposed scheme, the performance of phase noise cancellation of the OP-PNSC scheme can be derived in terms of CPE and ICI. Usually, phase noise is small in some communication systems and standards [21], [29]. Therefore, it is approximately equal to $e^{j\phi(nT_s)}=1+j\phi(nT_s)$, and the conjugated phase noise is equal to $e^{-j\phi(nT_s)}=1-j\phi(nT_s)$. According to (10), the CPE term in the $k$-th sub-carrier, $R_{CPE}^{OP}[k]$, can be expressed as follows:

$$\begin{aligned}R_{CPE}^{OP}[k]&=\frac{1}{2}\left(a[0]S_x[k]+a^*[0]S_y[k]\right)\\ &=\frac{1}{2N}\sum_{n=0}^{N-1}\left(1+1+(1-1)j\phi(nT_s)\right)S'[k]=S'[k]\end{aligned} \tag{11}$$

In addition, the ICI term in the $k$-th sub-carrier is

$$\begin{aligned} R_{ICI}^{OP}[k] &= \frac{1}{2}\sum_{l=1}^{N-1}\left(a[l]S_x[k-l]+a^*[l]S_y[k-l]\right) \\ &= \frac{1}{2N}\sum_{l=1}^{N-1}S'[k-l]\sum_{n=0}^{N-1}\left(e^{-j2\pi\frac{nl}{N}}+e^{j2\pi\frac{nl}{N}}+j\phi(nT_s)(e^{-j2\pi\frac{nl}{N}}-e^{j2\pi\frac{nl}{N}})\right) \\ &= \frac{1}{N}\sum_{l=1}^{N-1}S'[k-l]\sum_{n=0}^{N-1}\left(\phi(nT_s)\sin 2\pi\frac{nl}{N}\right) \end{aligned} \tag{12}$$

It can be seen that, the CPE term is canceled out to 0 and ICI term is degraded. The comparison of the OP-PNSC scheme and OFDM system with phase noise will be shown in Section 5.

### 3.2 The OP-PNSC Scheme in the PDL Channel

The phase noise cancellation performance of the OP-PNSC scheme which is efficient in cancelling out phase noise is derived in the AWGN channel. However, the OP signals transmitted in the wireless channel are distorted by depolarization effects. According to (11) and (12), the CPE and ICI cancellation are related to difference of CFR. Therefore, it is need to be investigated the OP-PNSC scheme in the depolarization channel.

Depolarization effects caused by the wireless channel are XPD, PMD and PDL [18]. XPD is the ability to separate the vertical and horizontal polarizations resulting in interference from one polarization to another. Pre-coding ( $\boldsymbol{UU}^H$ ) and post-coding ( $\boldsymbol{V}^*\boldsymbol{V}^{*H}$ ) obtain the diagonal matrix, $\boldsymbol{\Sigma}$ as in (6). And, XPD can be cancelled out [22]. PMD is caused by multi-path which can be mitigated in OFDM system. Therefore, we focus on the PDL effect, which is a serious problem distorting the phase noise cancellation in the frequency domain. Thus, we will derive the CPE and ICI cancellation performance in the PDL channel. Then, a PDL pre-compensation OP-PNSC (PPC-OP-PNSC) scheme is proposed and the comparisons of the performance before and after PDL compensation are given.

#### 3.2.1The Distortion of PDL in the OP-PNSC Scheme

The PDL effect is related to difference of the channel singular values [18]. In the PDL channel, the frequency domain received signals are obtain by substituting $\boldsymbol{\Sigma}$ from (6) into (8),

$$\boldsymbol{R}[k]=\begin{bmatrix}\sum_{l=0}^{N-1}a[l]\boldsymbol{\Sigma}_{x,k-l}[(k-l)_N]\mathbf{P}_x S_x[(k-l)_N] \\ \sum_{l=0}^{N-1}a[l]\boldsymbol{\Sigma}_{y,k-l}[(k-l)_N]\mathbf{P}_y S_y^*[(k-l)_N]\end{bmatrix}+\begin{bmatrix}W_x[k]\\W_y[k]\end{bmatrix} \tag{13}$$

The diagonal matrices from the same channel, $\boldsymbol{H}_x=\boldsymbol{H}_y$, in the two branches are equal: $\boldsymbol{\Sigma}_{x,k}=\boldsymbol{\Sigma}_{y,k}=\begin{bmatrix}\sqrt{\lambda_{1,k}} & 0\\ 0 & \sqrt{\lambda_{2,k}}\end{bmatrix}$ and $\mathbf{P}_m=\begin{bmatrix}\mathrm{P}_{m,H}\\ \mathrm{P}_{m,V}\end{bmatrix}$. Here, $\mathrm{P}_{m,H}$ and $\mathrm{P}_{x,V}$ are the horizontal and vertical components of the polarizations. For a given polarized signal, the power of polarized signals can be expressed by $\rho_m = \mathrm{P}_{m,H}^2 + \mathrm{P}_{m,V}^2$. Here, $\rho_m$ is the power of the polarized signal. By definition, PDL causes the power attenuation of the received frequency domain signals with different transmitted polarizations. At the receiver, the power of the signals in the *x* and *y* branches can be expressed as follows:

$$\rho_{x(y),k} = \mathbf{\Sigma}_{x(y),k} \begin{bmatrix} \mathrm{P}_{x(y),H} \\ \mathrm{P}_{x(y),V} \end{bmatrix} = \begin{bmatrix} \sqrt{\lambda_{1,k}} & 0 \\ 0 & \sqrt{\lambda_{2,k}} \end{bmatrix} \begin{bmatrix} \mathrm{P}_{x(y),H} \\ \mathrm{P}_{x(y),V} \end{bmatrix} = \lambda_{1,k} \mathrm{P}_{x(y),H}^2 + \lambda_{2,k} \mathrm{P}_{x(y),V}^2, \tag{14}$$

where $\mathrm{P}_{x(y),H}$ and $\mathrm{P}_{x(y),V}$ are the horizontal and vertical components for polarization $x(y)$. The polarization states of the transmitted signals are horizontal and vertical: $\mathbf{P}_H = \begin{bmatrix} 1 & 0 \end{bmatrix}^T$, $\mathbf{P}_V = \begin{bmatrix} 0 & 1 \end{bmatrix}^T$ using Jones vectors notation. Substituting (14) into (13), the frequency domain received signals distorted by PDL effect are expressed by

$$\boldsymbol{R}[k] = \begin{bmatrix} \sum_{l=0}^{N-1} a[l] \sqrt{\rho_{x,k-l}} S_x[(k-l)_N] \\ \sum_{l=0}^{N-1} a[l] \sqrt{\rho_{y,k-l}} S_y^*[(k-l)_N] \end{bmatrix} + \begin{bmatrix} W_x[k] \\ W_y[k] \end{bmatrix} \tag{15}$$

The signals powers in the two branches after polarization matching receiving by the ODPAs are $\rho_{x,k} = \lambda_{1,k}$ and $\rho_{y,k} = \lambda_{2,k}$. According to (14) and (15), the CPE and ICI terms of the OP-PNSC scheme distorted by PDL are given by

$$R_{CPE}^{OP}[k] = \frac{1}{2N} S'[k] \sum_{n=0}^{N-1} \left( \sqrt{\lambda_{1,k}} + \sqrt{\lambda_{2,k}} + (\sqrt{\lambda_{1,k}} - \sqrt{\lambda_{2,k}}) j\phi(nT_s) \right) \tag{16}$$

and

$$R_{ICI}^{OP}[k] = \frac{1}{2N} \sum_{l=1}^{N-1} S'[k-l] \sum_{n=0}^{N-1} j\phi(nT_s) (\sqrt{\lambda_{1,k-l}} e^{-j2\pi\frac{nl}{N}} - \sqrt{\lambda_{2,k-l}} e^{j2\pi\frac{nl}{N}}) \tag{17}$$

The phase noise coefficients of the OP-PNSC scheme in the PDL channel cannot be canceled out completely due to that residual phase noise still exists. When the PDL value is increased, the difference in the phase noise coefficient between the two branches at the receiver is increased, resulting in greater residual phase noise. In addition, the residual phase noise results in a degradation of the performance of the OP-PNSC scheme. Therefore, it can be obviously concluded that PDL reduces the efficiency of the phase noise cancellation of the OP-PNSC scheme.

### 3.2.2 The PPC-OP-PNSC Scheme

Based on the OP-PNSC scheme, the PPC-OP-PNSC scheme, which obtains the equivalent phase noise coefficients in the two branches by compensating the power in the *y* branch, is proposed.

The phase noise residual is related to the difference between the maximum and minimum singular values of the dual-polarized channel. Our goal is to reduce the phase noise residual by mitigating the power difference between the received orthogonal polarized signals in the two branches. To obtain a better SNR, we compensate for the received signals in the *y* branch as a result of the greater power attenuation. Under the perfect CSI assumption, the pre-compensated matrix, depending on the PDL estimated value, is multiplied by the transmitted signals in the *y* branch. Here, CSI can be estimated by inserting pilots at the transmitter described in, for example, [19]. The pre-compensated factor for the *k*-th sub-carrier is given by

$$\chi_k = \sqrt{\lambda_{1,k}} \Big/ \sqrt{\lambda_{2,k}} = \sqrt{10^{PDL_k/10}} \tag{18}$$

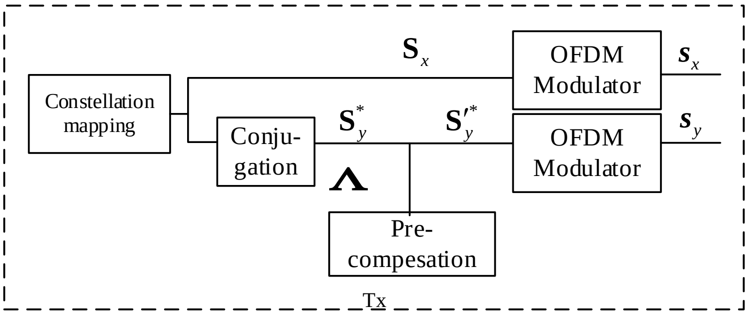


**Fig. 3.** Pre-compensation unit

An overview of the pre-compensation is shown in **Fig. 3.** The pre-compensation matrix $\mathbf{\Lambda}$, composed of $\chi_k$ , $k = 0,\ldots,N-1$ , can be written as follows:

$$\mathbf{\Lambda} = \begin{pmatrix} \chi_0 & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & \chi_{N-1} \end{pmatrix} \tag{19}$$

Using matrix notation, *N* parallel signals in the frequency domain can be written as $\mathbf{S}_y^* = [S_y^*[0]\cdots S_y^*[N-1]]^T$ , which is compensated by $\mathbf{\Lambda}$ to obtain the transmitted frequency domain signals in the *y* branch by

$$\mathbf{S}_y'^* = \mathbf{\Lambda}\mathbf{S}_y^* = [\chi_0 S_y^*[0],\cdots,\chi_{N-1}S_y^*[N-1]] \tag{20}$$

When the channel is flat fading, $\chi_0 = \chi_1 ... = \chi_{N-1}$ and the channel is frequency selective fading , $\chi_0 \neq \chi_1 ... \neq \chi_{N-1}$. After pre-compensation, the frequency domain received signals for *k*-th sub-carrier in the *y* branch can be expressed as follows:

$$\begin{aligned} R_y^*[k] &= \sum_{l=0}^{N-1} a^*[l]\chi_{k-l}\sqrt{\rho_{y,k-l}}S_y[(k-l)_N] + W_y^*[k] \\ &= a^*[0]\sqrt{\lambda_{1,k}}S_y[k] + \sum_{l=1}^{N-1} a^*[l]\sqrt{\lambda_{1,k-l}}S_y[(k-l)_N] + W_y^*[k], \end{aligned} \tag{21}$$

where $\chi_k\sqrt{\rho_{y,k}}=\chi_k\sqrt{\lambda_{2,k}}=\sqrt{\lambda_{1,k}}$ . According to (16) and (17), the CPE term of the decision variable is

$$R_{CPE}^{PPC}[k]=\frac{1}{2N}S'[k]\sum_{n=0}^{N-1}\left(\sqrt{\lambda_{1,k}}+\sqrt{\lambda_{1,k}}+(\sqrt{\lambda_{1,k}}-\sqrt{\lambda_{1,k}})j\phi(nT_s)\right)=\sqrt{\lambda_{1,k}}S'[k] \tag{22}$$

and the ICI term is

$$R_{ICI}^{PPC}[k]=\frac{1}{N}\sum_{l=1}^{N-1}\sqrt{\lambda_{1,k-l}}S'[k-l]\sum_{n=0}^{N-1}\phi(nT_s)\sin 2\pi\frac{nl}{N} \tag{23}$$

We can see that the distortion of phase noise in the OP-PNSC scheme is degraded after pre-compensation in the PDL channel. When PDL=0, which means that the power attenuations of the two branches are equal, pre-compensation is not needed. It is the upper bound of the phase noise cancellation performance for the OP-PNSC scheme in the PDL channel, which is only related to $\sqrt{\lambda_{1,k}}$ . In particular, the orthogonal polarized signals are only distorted by Gaussian noise in the AWGN channel ($\sqrt{\lambda_{1,k}}=1$). Comparing with the OP-PNSC scheme without pre-compensation, the PPC-OP-PNSC scheme demonstrates greater efficiency in canceling out phase noise.

## 4. Performance Analysis and Discussion

In this section, the SINR performance of the OP-PNSC scheme will be analyzed to verify the phase noise cancellation performance. The SE performance is also analyzed. Moreover, the comparison between the OP-PNSC scheme and the polarization diversity (PD) scheme is given.

### 4.1 Performance Analysis

The interference of the desired signals from other sub-carriers caused by ICI is indicated by the SINR. Thus, the SINR of the OP-PNSC scheme in the AWGN and PDL channel with and without pre-compensation are given. Moreover, comparing with the phase noise self-cancellation scheme, the OP-PNSC scheme can maintain the SE. Therefore, the SINR and SE will be analyzed to evaluate the performance of the OP-PNSC /PPC-OP- PNSC scheme.

#### 4.1.1 SINR

As mentioned in [13], in an OFDM system with phase noise, the interference is the ICI term caused by phase noise, and the power of the desired signals is equal to that of the CPE term. Thus, the SINR can be written as follows:

$$SINR=\frac{\mathrm{E}\left[\left|R_{CPE}\right|^2\right]}{\mathrm{E}\left[\left|R_{ICI}\right|^2\right]+\sigma^2} \tag{24}$$

In the AWGN channel, according to , the OP-PNSC scheme can cancel out the constellation rotation caused by CPE. Therefore, the received desired signal (the CPE term) power in the *k*-th sub-carrier can be represented as

$$\mathrm{E}\left[\left|R_{CPE}\right|^2\right]=\mathrm{E}\left[\left|S'[k]\right|^2\right] \tag{25}$$

The ICI term is also degraded in the OP-PNSC scheme according to . The interference power in the numerator of (24) in the *k*-th sub-carrier is given by

$$\mathrm{E}\left[\left|R_{ICI}\right|^2\right]=\mathrm{E}\left[\left|\frac{1}{N}\sum_{l=1}^{N-1}S'[k-l]\sum_{n=0}^{N-1}\left(\phi(nT_s)\sin 2\pi\frac{nl}{N}\right)\right|^2\right] \tag{26}$$

It is assumed that the transmitted signals have zero mean and are statistically independence [24]. The SINR of the OP-PNSC scheme in the AWGN channel can be expressed as follows:

$$SINR_{OP}=\frac{1}{\sum_{l=1}^{N-1}\mathrm{E}\left[\left|\frac{1}{N}\sum_{n=0}^{N-1}\left(\phi(nT_s)\sin 2\pi\frac{nl}{N}\right)\right|^2\right]+\sigma^2} \tag{27}$$

Now, we will analyze the SINR performance in the PDL channel. According to (16), the power of the CPE term in the PDL channel is given by

$$E\left[\left|R_{CPE}\right|^2\right]=\frac{1}{4}\mathrm{E}\left[\left|\left(\sqrt{\lambda_{1,k}}+\sqrt{\lambda_{2,k}}+(\sqrt{\lambda_{1,k}}-\sqrt{\lambda_{2,k}})\frac{1}{N}\sum_{n=0}^{N-1}j\phi(nT_s)\right)\right|^2\right]E\left[\left|S'[k]\right|^2\right] \tag{28}$$

The powers of the ICI term and the SINR are written as

$$R_{ICI}[k]=\frac{1}{4}\mathrm{E}\left[\left|\frac{1}{N}\sum_{l=1}^{N-1}S'[k-l]\sum_{n=0}^{N-1}\left(\sqrt{\lambda_{1,k-l}}-\sqrt{\lambda_{2,k-l}}\,j\phi(nT_s)\right)e^{-2\pi\frac{nl}{N}}\right|^2\right] \tag{29}$$

And the SINR in the PDL channel is

$$SINR_{PDL}=\frac{\mathrm{E}\left[\left|\left(\sqrt{\lambda_{1,k}}+\sqrt{\lambda_{2,k}}+(\sqrt{\lambda_{1,k}}-\sqrt{\lambda_{2,k}})\frac{1}{N}\sum_{n=0}^{N-1}j\phi(nT_s)\right)\right|^2\right]}{\mathrm{E}\left[\left|\frac{1}{N}\sum_{l=1}^{N-1}\sum_{n=0}^{N-1}j\phi(nT_s)(\sqrt{\lambda_{1,k-l}}e^{-j2\pi\frac{nl}{N}}-\sqrt{\lambda_{2,k-l}}e^{j2\pi\frac{nl}{N}})\right|^2\right]+\sigma^2} \tag{30}$$

The SINR of the OP-PNSC scheme in the PDL channel is related to the PDL value and is still distorted by the residual phase noise. As a comparison, the SINR of the PPC-OP-PNSC scheme is derived. After pre-compensation, the CPE term can be written as follows:

$$E\left[\left|R_{CPE}\right|^2\right]=\rho_{x,k}E\left[\left|S'[k]\right|^2\right]=\lambda_{1,k}E\left[\left|S'[k]\right|^2\right] \tag{31}$$

According to (23), the SINR of thePPC-OP-PNSC scheme can be expressed as follows:

$$SINR_{PPC} = \frac{N^2 \lambda_{1,k}}{\sum_{l=1}^{N-1} \lambda_{1,k-l} \left| \sum_{n=0}^{N-1} \phi(nT_s) \sin 2\pi \frac{nl}{N} \right|^2 + N^2 \sigma^2} \tag{32}$$

### 4.1.2 Spectral Efficiency

By employing the ODPAs to transmit the orthognal polarized signals in the same frequency, the SE of the OP-PNSC scheme more than that of the phase noise self-cancellation (PHNSC) scheme. Therefore, the spectral efficiency (SE) of the OP-PNSC scheme and the PHNSC scheme is analyzed. The SE is defined as the ratio of the information data rate that can be transmitted over a given bandwidth, which is related to the SINR, $\eta_{SE} = k/W = \log_2(1+SINR)$. It is assumed that the data rate of the OFDM system without phase noise is $k$ bit/s, and the bandwidth is $W$ Hz.

In [13]-[15], the PHNSC schemes transmit the same signals through two sub-carriers. To obtain the same SINR, $SINR_{self} = SINR_{OP}$, the required bandwidth of the PHNSC schemes is $2W$ when the data rate of the phase noise self-cancellation scheme is $k$ bit/s. The reason is that these schemes use two adjacent or conjugated sub-carriers to transmit the same data. Thus, the SE of the PHNSC scheme is given by

$$\eta_{self} = \frac{k}{2W} \tag{33}$$

From (33), the SE of the phase noise self-cancellation scheme is only half of that of the OFDM system without phase noise. Note that two ODPAs of the OP-PNSC scheme are employed, which are worked in the same frequency. Thus, the required bandwidth of the OP-PNSC scheme is $W$. Therefore, the SE of the OP-PNSC scheme can be written as follows:

$$\eta_{OP-PNSC} = \frac{k}{W} = \frac{1}{2}\eta_{self} \tag{34}$$

Comparing with the phase noise self-cancellation scheme, the SE of the OP-PNSC scheme remains unchanged because the orthogonal dual-polarized antennas are employed. These orthogonal dual-polarized antennas, which can be co-located, employ polarization isolation instead of space isolation, providing space and cost savings. Especially, the proposed scheme are useful in the bandwidth limited scenarios.

## 4.2 Discussion

Similar to the OP-PNSC scheme, the PD scheme combined with the OFDM system (OFDM-PD) in [25]-[27] employs orthogonal dual-polarized antennas to transmit the same signals. However, the OP-PNSC scheme employs phase-conjugated processing at the transceiver to make the phase noise coefficients of the two branches mutually conjugated, which is the foundation of phase noise self-cancellation. In other words, without conjugation and additional processing, the OFDM-PD scheme cannot cancel out phase noise. In this section, the distortion of phase noise in the OFDM-PD scheme will be discussed.

Polarization diversity employs orthogonality of H/V polarization in channel and obtains the performance gain via fading uncorrelation of the orthogonal polarized signals. Combining the polarization diversity and OFDM can be used to obtain a higher SE. Although the OFDM-PD scheme used orthogonal dual-polarized antennas at the transceiver to provide diversity gain using space-frequency block coding (SFBC) [28], it is sensitive to phase noise. Phase noise introduces ICI, including not only the interference among sub-carriers in an OFDM symbol but also that between transmitters, which will destroy the Alamouti-type SFBC structure, resulting in diminished diversity gains of the OFDM-PD scheme and significant performance degradation [28], [29].

According to [28] and [29], pairs of information symbols ( $S[k]$ , $S[k+1]$ ) are coded using SFBC, i.e., for sub-carrier *k*, $S[k]$ and $S[k+1]$ are transmitted from H and V antennas, respectively. For sub-carrier *k*+1, $-S^*[k+1]$ and $S^*[k]$ are transmitted from H and V antennas, respectively. The received signals in the *k*-th sub-carrier in the presence of phase noise are expressed in (35).

$$\begin{aligned}
R[k] &= \left(\left|H_{HH}[k]\right|^2 S[k] - H_{HH}^*[k]H_{HV}[k]S^*[k+1]\right)a[k] \\
&\quad + \left(\left|H_{VH}[k]\right|^2 S[k] - H_{VH}^*[k]H_{VV}[k]S^*[k+1]\right)a[k] \\
&\quad + \left(H_{HV}[k]H_{HH}^*[k]S^*[k+1] + \left|H_{HV}[k]\right|^2 S[k]\right)a^*[k+1] \\
&\quad + \left(H_{VV}[k]H_{VH}^*[k]S^*[k+1] + \left|H_{VV}[k]\right|^2 S[k]\right)a^*[k+1] \\
&= \left(\left(\left|H_{HH}[k]\right|^2 a[k] + \left|H_{VV}[k]\right|^2 a^*[k+1]\right) + \left(\left|H_{HV}[k]\right|^2 a^*[k+1] + \left|H_{VH}[k]\right|^2 a[k]\right)\right)S[k] \\
&\quad + \left(\left(H_{HV}[k]H_{HH}^*[k]a^*[k+1] - H_{HH}^*[k]H_{HV}[k]a[k]\right)\right. \\
&\quad \left. + \left(H_{VV}[k]H_{VH}^*[k]a^*[k+1] - H_{VH}^*[k]H_{VV}[k]a[k]\right)\right)S^*[k+1]
\end{aligned} \tag{35}$$

From (35), $S^*[k+1]$ term does not exist if phase noise is absent, and the received signals is equal to $\left(\left|H_{HH}[k]\right|^2 + \left|H_{VV}[k]\right|^2 + \left|H_{HV}[k]\right|^2 + \left|H_{VH}[k]\right|^2\right)S[k]$. However, the desired signals $S[k]$ in the *k*-th sub-carrier are distorted by phase noise, $a[k]$ and $a[k+1]$. In addition, $R[k]$ remains distorted by ICI caused by phase noise. More analysis of the phase noise effect in the SFBC scheme can be seen in [29]. Comparing with the OP-PNSC scheme, the SFBC structure is destroyed by phase noise, resulting in the reduced BER performance of the OFDM-PD scheme, which is simulated in Section 4.

The OP-PNSC scheme cancels out phase noise only via the conjugation and addition processing, which can be implemented based on hardware without a complex algorithm. Therefore, the computational complexity of the OP-PNSC scheme only depends on the FFT length without considering the channel. The complexity of phase noise cancellation is $O\left(2N\log_2 N\right)$ due to two branches with FFT employed. Moerover, considering the channel estimatin and compensation, SVD decomposition complexity is caculated. The SVD decomposition is cube of the size of channel matrix. In the PPC-OP-PNSC scheme, channel is a $2\times 2$ matrix. Thus, the computational complexity of SVD decomposition is $O(2^3 N) =$

$O(8N)$. Then, it is calculated that the total computational complexity of the PPC-OP-PNSC is $O(2N\log_2 N + 8N)$.

The OP-PNSC scheme can be readily implemented to MIMO-OFDM by employing the SFBC or V-BLAST coding, etc. For example, by empolying $M_t$ pairs of transmitted ODPAs and $M_r$ pairs of recieved ODPAs, the polarized signals are coding by Alamouti-type, etc., which can provided the diversity gain. In the high frequency communication, the OP-PNSC scheme is also can be implemented to deal with phase noise due to that the scheme is independent of the hardware equipment.

In this section, the SINR and SE performances of the OP-PNSC scheme are analyzed and found to demonstrate the superiority of this scheme. As a comparison, the performance of the OFDM-PD scheme in the presence of phase noise is determined.

## 5. Simulation and Numerical Results

In this section, the BER comparison with the conventional OFDM system and the DPOLSK scheme is presented. Then, the SINR and BER comparisons with the phase noise compensation (PHNC) and phase noise self-cancellation (PHNSC) scheme are illustrated in the AWGN and PDL channels to demonstrate that the OP- PNSC (PPC-OP-PNSC) scheme achieves a better phase noise cancellation performance under the same channel conditions.

Computer simulation results are presented in terms of the SINR and BER performances of the OP-PNSC scheme in both the AWGN channel and the PDL channel. The SINR and BERs are used to discuss the system performance when distorted by phase noise under the OFDM system with and without phase noise, the OP-PNSC scheme, the phase noise compensation scheme and the phase noise self-cancellation scheme. The BER versus the required transmitted signal-to-noise rate $E_b/N_0$ ($E_b$ is the transmitted signal power, and $N_0$ is the spectral density coefficient for the AWGN) are considered. 16QAM is adopted in each sub-carrier. The OFDM parameters are shown in **Table 1** and the phase noise parameters are shown in **Table 2**. Accordingly, the PSDs of the phase noise are -76, -80 and -85 dBc/Hz@1 MHz, respectively [31]. The channel matrix is generated as the dual-polarized channel modeled in [18], and it is assumed that the receiver has perfect channel state information. To transmit and receive the orthogonal polarized signals, the transceiver employs the orthogonal dual-polarized antennas with horizontal and vertical polarizations.

**Table 1.** OFDM parameters

| | |
|---|---|
| data length | 128 |
| FFT length | 256 |
| CP length | 32 |

**Table 2.** Phase noise parameters

| | |
|---|---|
| integrated phase noise power | -16/-20-24 dBc |
| noise floor | -120dBc |
| cut-off frequency | 1MHz |
| sampling frequency | 80MHz |

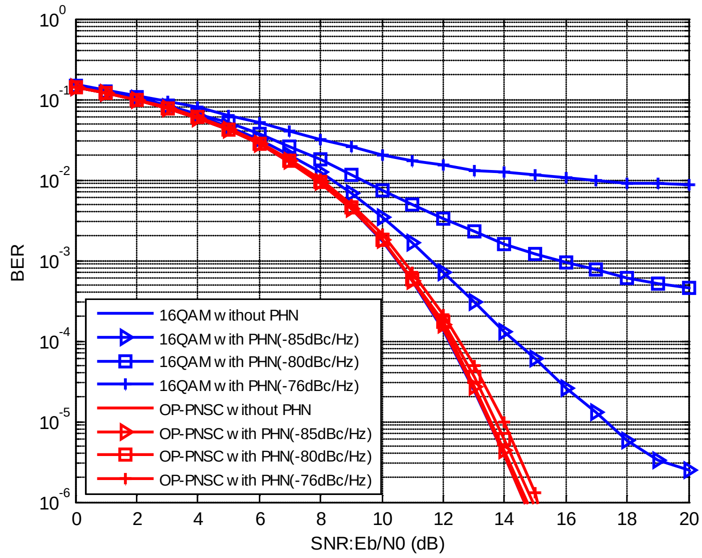


**Fig. 4.** BER comparison of the OP-PNSC scheme and OFDM in the presence of phase noise in the AWGN channel(16QAM)

The BER performance comparison between the OP-PNSC scheme and the OP-OFDM (OFDM system combing orthogonal polarization) system is shown in **Fig. 4**. As observed in **Fig. 4**, the BER performance of the OP-OFDM system with phase noise is worse than that of the system without phase noise, i.e., the BER performance of the OP-OFDM system with phase noise (the PSD is -80 dBc/Hz@1 MHz) is 5 dB worse than that of the system without phase noise when the BER is $10^{-3}$. Furthermore, when the SNR is more than 15 dB, the BER curve is almost horizontal. This is because phase noise is a multiplied effect such that phase noise cannot be suppressed by increasing the SNR. This phenomenon caused by phase noise is called the noise floor. By contrast, the BER performance of the OP-PNSC scheme is the same as that of the conventional scheme without phase noise. Comparing to the OP-OFDM system with phase noise, the BER performance of the OP-PNSC scheme (the PSD is -85 dBc/Hz@1 MHz) is 2 dB better when the BER is $10^{-4}$. Moreover, when the BER is $10^{-3}$, the BER performance gain is 5.2 dB (the PSD is -80 dBc/Hz@1 MHz). Consequently, the OP-PNSC scheme can effectively cancel out the distortion of phase noise in the OFDM system to improve the BER performance when the LO is non-ideal.

**Fig. 5** depicts the BER performance as a function of the SNR, including the OP-PNSC scheme with 16QAM and the 16DPOSLK proposed in [17]. From this figure, it is observed that the BER performance of the OP-PNSC scheme is apparently better than that of the DPOLSK scheme. This is because the DPOSLK scheme is only focused on CPE cancellation. Moreover, the power of the AWGN in the DPOLSK scheme is twice that in the OP-PNSC scheme because the two components of the DPOLSK signals are influenced by the same AWGN. Consequently, the BER performance of the OP-PNSC scheme is substantially better than that of the DPOLSK scheme.

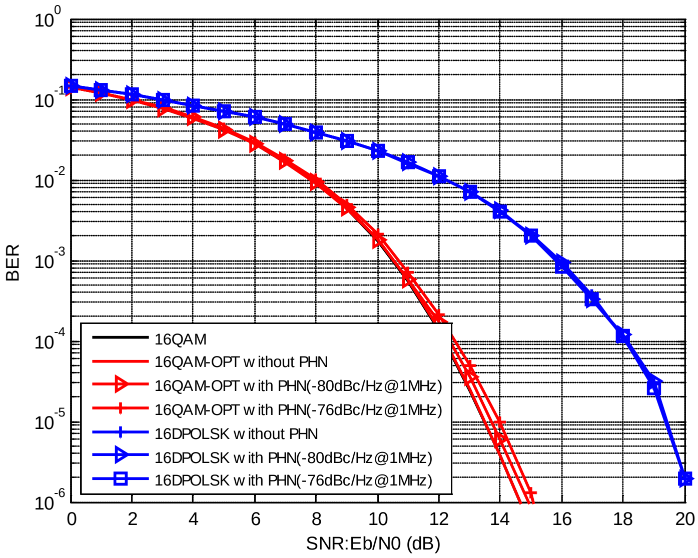


**Fig. 5.** BER comparison of the OP-PNSC scheme (16QAM) and the 16DPOLSK scheme

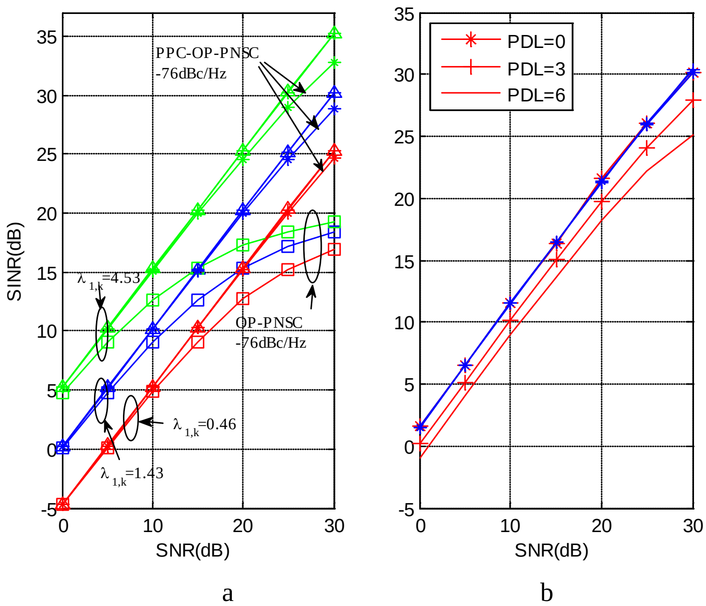


**Fig. 6.** The SINR performance of the OP-PNSC and PPC-OP-PNSC schemes. (a) Comparison between the OP-PNSC scheme and the OP-OFDM system. (b) Comparison between the OP-PNSC and PPC-OP-PNSC schemes with different PDL values: $\lambda_{1,k}=1.43$.

The SINR performances of both the OP-PNSC scheme and the PPC-OP-PNSC scheme in the PDL channel are illustrated in **Fig. 6**. The PSD of the phase noise is -76 dBc/Hz@1 MHz and PDL = 2 dB. As shown in **Fig. 6(a)**, the SINR comparison between the OP-OFDM system and the OP-PNSC scheme with and without phase noise is illustrated. It can be observed that the SINR increases with increasing $\lambda_{1,k}$. However, the SINR of the OP-OFDM system with phase noise is 5-15 dB worse than that of the system without phase noise. The SINR performance of the OP-PNSC scheme is substantially better than that of the OP-OFDM system in the presence of phase noise. For example, the OP-PNSC scheme can obtain a 2-13 dB gain in terms of the SINR for high SNR (more than 20 dB). **Fig. 6(b)** shows the SINR performance of the OP-PNSC scheme before and after PDL pre-compensation. It is clearly observed that the SINR performance of the OP-PNSC scheme in the PDL channel is decreased with increased PDL value, resulting from the ICI term power increase caused by residual phase noise. By contrast, the PPC-OP-PNSC scheme achieves a better SINR performance, and the SINR curves are almost coincident with those of the OP-PNSC scheme when PDL=0 dB (PDL=0 dB indicates no PDL effect in the channel).

The BER performance comparison between the PPC-OP-PNSC scheme (red curves) and the OP-PNSC scheme without PDL pre-compensation (black curves) with 16QAM in the PDL channel is shown in **Fig. 7**. As observed in **Fig. 7(a)**, due to the power imbalance caused by the PDL effect, the BER (black curves) is worse than that in the AWGN channel, and the residual phase noise will lead to increased bit error. In particular, the PSD of the phase noise is -76 dBc/Hz@1MHz, and the error floor is very obvious, resulting in a decrease in reliability. However, the OP-PNSC scheme after PDL pre-compensation achieves a better BER performance (red curves) in the presence of phase noise. It can be observed in **Fig. 7(a)** that, after pre-compensation, the BER curves are almost coincident with those of the scheme without phase noise. Comparing to the proposed scheme without pre-compensation, the BER performance is 6 dB better. The BER performance of **Fig. 7(b)** is used to illustrate the PDL vs. BER when SNR=12 dB and $\lambda_{1,k}=1.43$. The BER performance of the OP-PNSC scheme without pre-compensation is decreased with increased PDL value because the increased power imbalance results in increased residual phase noise. As a comparison, the BER performance of the PPC-OP-PNSC scheme is stable, which means that the power imbalance caused by PDL is compensated.

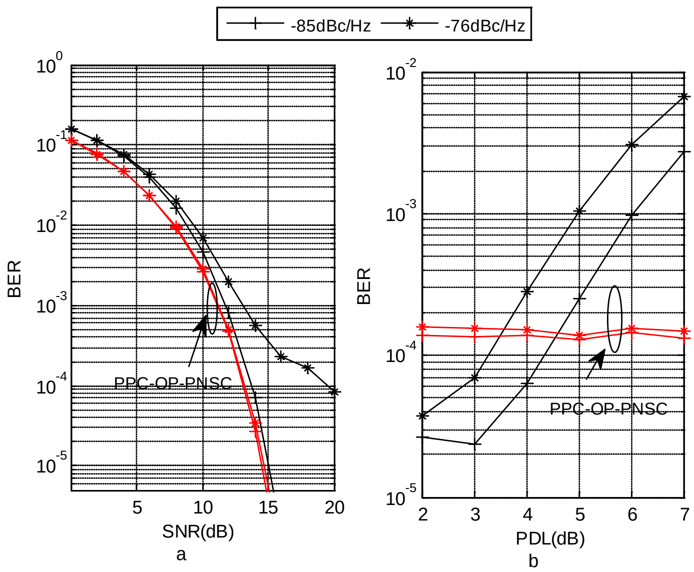


**Fig. 7.** BER performance of the PPC-OP-PNSC scheme in the PDL channel. (a) BER vs SNR, PDL=5dB, $\lambda_{1,k}=1.43$. (b)BER vs PDL, SNR=12dB, $\lambda_{1,k}=1.43$.

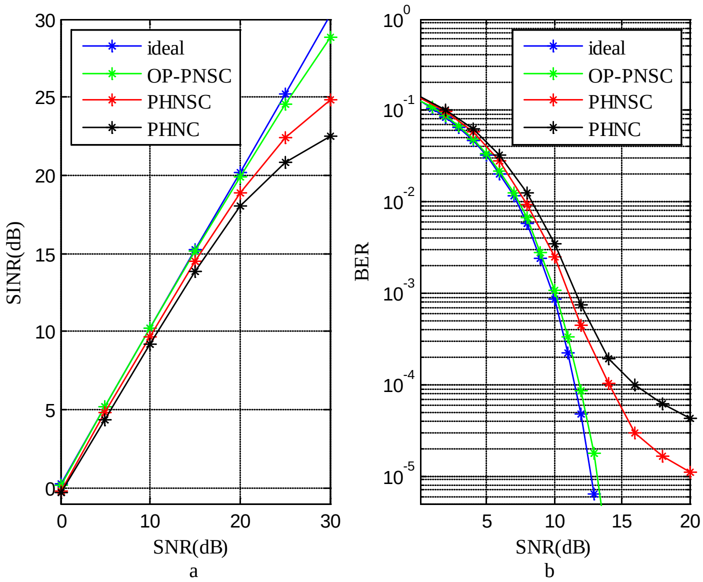


**Fig. 8.** BER and SINR performance comparison of the PPC-OP-PNSC, PHNC and PHNSC schemes. (a) SINR performance. (b) BER performance.

The comparison of the OP-OFDM, OP-PNSC, phase noise compensation (PHNC) and phase noise self-cancellation (PHNSC) schemes is illustrated in terms of the SINR and BER performances. The PSD of the phase noise is -76 dBc/Hz@1 MHz. The SINR performance is plotted in **Fig. 8(a)**, in which it is obvious that the performance of the OP-PNSC scheme is better than that of both the phase noise compensation scheme and the phase noise self-cancellation scheme. This is because the phase noise cancellation of the phase noise compensation scheme needs the estimated accuracy. However, the estimated accuracy is limited, resulting in a reduced phase noise cancellation performance. The phase noise self-cancellation scheme employs the adjacent sub-carriers to cancel out phase noise without estimation; thus, the performance of the phase noise cancellation scheme is better. However, the phase noise coefficient of the adjacent sub-carriers is not exactly equal; thus, residual phase noise remains. The OP-PNSC scheme can cancel out phase noise completely and is 4-6 dB better than the phase noise compensation and phase noise self-cancellation schemes. As shown in **Fig. 8(b)**, it can also be observed that the BER performance of the OP-PNSC scheme is the best. The phase noise compensation and phase noise self-cancellation schemes have an error floor in the high SNR region. However, the BER curve of the OP-PNSC scheme is coincident with that of the OP-OFDM system without phase noise.

**Fig. 9** shows a comparison of the OP-PNSC scheme and the OFDM-PD scheme. As mention in Section 3, these two schemes have the same hardware architecture. Although the OFDM-PD scheme can provide a diversity gain, the scheme is sensitive to phase noise, which can be observed in **Fig. 9**. In the presence of phase noise, the BER performance of the OP-PNSC scheme is 2 dB better than that of the OFDM-PD scheme. In particular, the SNR penalty is worse with increased PSD of phase noise. This is because phase noise causes the error floor to increase, resulting in a decreased BER performance of the OFDM-PD scheme. However, the OP-PNSC scheme achieves the better BER performance. It is clear that the BER curves of the OP-PNSC scheme (red curves) are 2-10 dB better than those of the OFDM-PD scheme and that the curves with phase noise are almost coincident with the curve without phase noise.

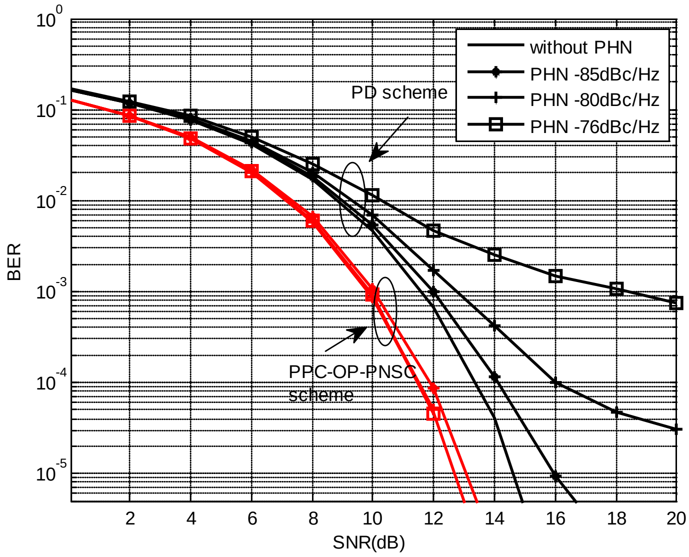


**Fig. 9.** BER comparison of the PPC-OP-PNSC and OFDM-PD schemes

## 6. Conclusion

In this paper, the OP-PNSC scheme, which can cancel out phase noise, including both CPE and ICI, effectively in the AWGN and realistic PDL channels, with increased SE, was proposed. With respect to the BER, the OP-PNSC scheme achieved a better performance compared with the phase noise compensation and phase noise self-cancellation schemes. Specifically, the OP-PNSC scheme brought about a more significant SE performance improvement compared to the phase noise self-cancellation scheme because the orthogonal polarized signals were transmitted in the same-frequency sub-carrier. Moreover, this paper also investigated the phase noise cancellation performance in the PDL channel, and a PDL pre-compensated OP-PNSC scheme was proposed. The simulation results show that the OP-PNSC scheme has an advantage in terms of BER performance in the AWGN and PDL channels. Therefore, the OP-PNSC scheme could be very useful for multi-carrier systems with high transmission quality.

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