Efficient ICI Self-Cancellation Scheme for OFDM Systems

Kyung-Hwa Kim and Bangwon Seo

In this paper, we present a new inter-carrier interference (ICI) self-cancellation scheme — namely, ISC scheme — for orthogonal frequency-division multiplexing systems to reduce the ICI generated from phase noise (PHN) and residual frequency offset (RFO). The proposed scheme comprises a new ICI cancellation mapping (ICM) scheme at the transmitter and an appropriate method of combining the received signals at the receiver. In the proposed scheme, the transmitted signal is transformed into a real signal through the new ICM using the real property of the transmitted signal; the fast-varying PHN and RFO are estimated and compensated. Therefore, the ICI caused by fast-varying PHN and RFO is significantly suppressed. We also derive the carrier-to-interference power ratio (CIR) of the proposed scheme by using the symmetric conjugate property of the ICI weighting function and then compare it with those of conventional schemes. Through simulation results, we show that the proposed ISC scheme has a higher CIR and better bit error rate performance than the conventional schemes.

Keywords: OFDM, phase noise, residual frequency offset, inter-carrier interference, ICI cancellation.

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I. Introduction

Recently, the orthogonal frequency-division multiplexing (OFDM) modulation scheme has been widely used in many communication systems, such as fourth-generation mobile communications [1] and wireless local area networks [2], owing to its high spectral efficiency and effectiveness in avoiding inter-symbol interference caused by multipath delays. However, orthogonality among subcarriers is required to utilize the advantages of OFDM. If residual frequency offset (RFO) or phase noise (PHN) is introduced by a voltage-controlled local oscillator, OFDM systems will suffer severely from inter-carrier interference (ICI). RFO and PHN destroy the orthogonality among subcarriers in OFDM systems, which causes ICI; thus severely degrading the performance of OFDM systems.

In [3]–[5], it was shown that the two main effects of PHN on OFDM systems, causing degradation of system performance, are common phase error (CPE) and ICI. Although CPE causes phase rotation for all subcarriers, it can be easily estimated (by using pilot symbols) and easily compensated. However, ICI — the interference between subcarriers — is difficult to completely eliminate in OFDM systems.

Many PHN compensation schemes have been proposed to reduce ICI [6]–[13]. Among them, ICI self-cancellation schemes — namely, ISC schemes — have received considerable attention because they can significantly reduce ICI by mapping data symbols (with a predefined ICI weighting function) onto a group of subcarriers that then transmit the outcome. By appropriately combining the signals within a group at a receiver, the ICI signals generated within the group can easily self-cancel one another [6]–[10]. In [10], a symmetric data-conjugated ISC scheme has been proposed.

This scheme was shown to have the best performance among conventional ISC schemes in terms of the carrier-to-interference ratio (CIR) and bit error rate (BER), because the scheme utilizes the symmetric conjugate property of the ICI weighting function. However, if there is an RFO or if the normalized 3 dB bandwidth (ρ) of the power spectral density (PSD) of PHN is large, the performance of the conventional symmetric data-conjugated ISC method is severely degraded.

In this paper, a single-input single-output (SISO) OFDM system is considered in which a transmitter having a single antenna transmits its data to a receiver with a single antenna. We propose a new ISC method to suppress the ICI generated from PHN and RFO — that is, when ρ is large or when an RFO exists. The proposed scheme comprises a new ICI cancellation mapping scheme — namely, ICM scheme — at the transmitter and an appropriate method of combining the received signals at the receiver. By using the new ICM scheme at the transmitter, the transmitted signal is transformed into a real signal, and the fast-varying PHN and RFO are estimated and compensated at the receiver using the real property of the transmitted signal. Therefore, the ICI generated by the fast-varying PHN and RFO is significantly suppressed. We derive the CIR of the proposed scheme using the symmetric conjugate property of the ICI weighting function and compare it with that of the conventional ISC scheme. The simulation results show that the proposed ISC scheme has a higher CIR and better BER performance than the conventional ISC scheme given in [10].

The rest of this paper is organized as follows: Section II describes a system model of the SISO OFDM system. In section III, we present a new ISC scheme for the SISO OFDM system with PHN and RFO. Section IV presents the simulation results, and section V concludes the paper.

II. System Model

In this paper, we consider a SISO OFDM system. Let X_m, $0 \le m \le N-1$, denote the data of the mth subcarrier, where N is the total number of subcarriers. We assume that

$$E[X_m X_k^*] = \begin{cases} E[|X_m|^2] = 1 & \text{if } m = k, \\ 0 & \text{if } m \ne k. \end{cases} \quad (1)$$

The transmitted time-domain signal x_n is expressed by

$$x_n = \frac{1}{N} \sum_{m=0}^{N-1} X_m \mathrm{e}^{j2\pi mn/N} \quad \text{for} \quad 0 \le n \le N-1. \quad (2)$$

We assume that there is a PHN, ϕ_n, at the local oscillator of the receiver. Therefore, the received time-domain signal affected by ϕ_n at the receiver can be written as

$$y_n = (h_n \otimes x_n)\mathrm{e}^{j\phi_n} + w_n \quad \text{for} \quad 0 \le n \le N-1, \quad (3)$$

where h_n and w_n represent the time-domain channel impulse response and additive white Gaussian noise (AWGN), respectively. The symbol $\otimes$ denotes circular convolution.

The PHN can be modeled by passing a white Gaussian noise sequence through a low-pass filter having a PSD given by [3]

$$L(f) = \begin{cases} 10^{-a} & 0 \le |f| \le \beta, \\ 10^{-a}(f/\beta)^{-b} & \beta \le |f| \le \gamma, \\ 10^{-c} & \gamma \le |f| \le f_s/2, \end{cases} \quad (4)$$

where β is the 3 dB bandwidth of PSD and f_s is the sampling rate. Here, a denotes the PHN level near the center frequency up to $\pm\beta$ and b represents the rate of decrease of PSD in the $\beta \le |f| \le \gamma$ frequency range. Further, c represents the noise floor.

In addition, PHN can be classified according to the normalized 3 dB bandwidth ρ as $\rho = \beta/\Delta F_s$, where ΔF_s is the subcarrier spacing. If $\rho < 1$, the PHN is considered to be slowly varying during an OFDM symbol period. Thus, in this case, the CPE component is the dominant factor that affects the detection performance at the receiver when compared with the ICI component. On the other hand, if $\rho \ge 1$, then the PHN is considered to be rapidly varying during an OFDM symbol period; thus, the ICI component dominates the CPE component [14].

The discrete Fourier transform (DFT) output of y_n is represented by

$$\begin{aligned} Y_k &= \sum_{n=0}^{N-1} y_n \mathrm{e}^{-j2\pi kn/N} \\ &= \frac{1}{N} \sum_{m=0}^{N-1} H_m X_m \sum_{n=0}^{N-1} \mathrm{e}^{j\phi_n} \mathrm{e}^{-j2\pi(k-m)n/N} + W_k, \quad (5) \\ &\text{for } 0 \le k \le N-1, \end{aligned}$$

where W_k is the DFT output of the AWGN and H_m is the channel coefficient of the mth subcarrier, which is represented by

$$H_m = \sum_{l=0}^{L-1} h_l \mathrm{e}^{-j2\pi ml/N}. \quad (6)$$

If we denote the ICI weighting function of the kth subcarrier by Θ_k, it can be written as

$$\Theta_k = \frac{1}{N} \sum_{n=0}^{N-1} \mathrm{e}^{j\phi_n} e^{-j2\pi kn/N}. \quad (7)$$

Using Θ_k, Y_k can be rewritten as

$$\begin{aligned} Y_k &= \sum_{m=0}^{N-1} H_m X_m \Theta_{k-m} + W_k \\ &= H_k X_k \Theta_0 + \sum_{m \ne k} H_m X_m \Theta_{k-m} + W_k, \quad 0 \le k \le N-1. \end{aligned} \quad (8)$$

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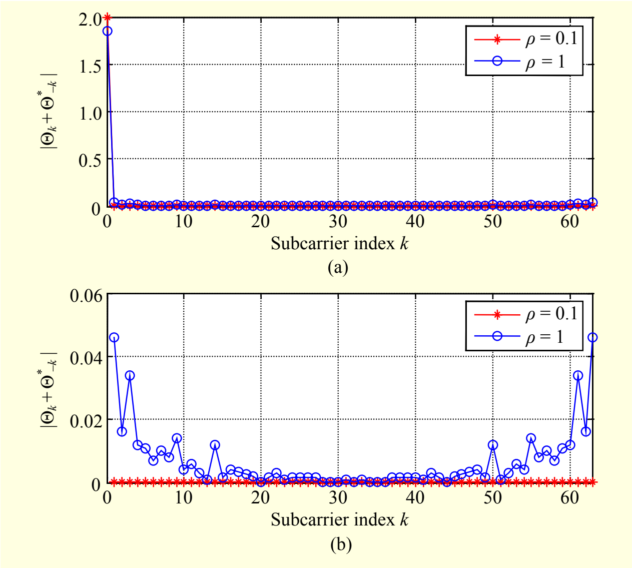

Fig. 1. Example of $|\Theta_k+\Theta^*_{-k}|$ for (a) $k=0, 1, \ldots, 63$ when ρ=0.1 and $\rho=1$ and (b) $k=1, 2, \ldots, 63$ when $\rho=0.1$ and $\rho=1$.

Here, Θ_0 represents the time average of the PHN process and it introduces a constant phase rotation of all the constellations. Therefore, it is usually referred to as CPE. On the other hand, the second term on the right-hand side of (8) represents ICI.

From (7), we know

$$\begin{aligned}\Theta_k+\Theta^*_{-k} &= \frac{1}{N}\sum_{l=0}^{N-1}(\mathrm{e}^{j\phi_n}+\mathrm{e}^{-j\phi_n})e^{-j2\pi kn/N}\\ &= \frac{2}{N}\sum_{l=0}^{N-1}\cos\phi_n e^{-j2\pi kn/N}.\end{aligned} \tag{9}$$

If we assume that the PHN ϕ_n has a zero mean and very small variance, then $\cos\phi_n$ can be approximated to be 1. Therefore, (9) can be rewritten as

$$\Theta_k+\Theta^*_{-k}\approx\frac{2}{N}\sum_{l=0}^{N-1}\mathrm{e}^{-j2\pi kn/N}=\begin{cases}2 & \text{for } k=0,\\ 0 & \text{otherwise.}\end{cases} \tag{10}$$

From (10), the ICI weighting function Θ_k clearly has a symmetric conjugate property [10].

The symmetric conjugate property can be clearly observed in Fig. 1. An example of $|\Theta_k+\Theta^*_{-k}|$ for $k=0, 1, \ldots, N-1$ (Where $N=64$ for both $\rho=0.1$ and $\rho=1$) is shown in Fig. 1(a). In Fig. 1(b), Θ_0 is not plotted, because its value is quite large when compared with Θ_k, for $k=1, 2, \ldots, 63$.

III. Proposed ISC Scheme for a SISO OFDM System with PHN and RFO

Figure 2 shows the block diagram of the proposed transmitter for the SISO OFDM system.

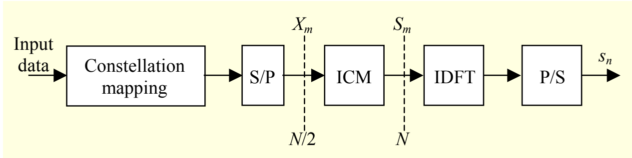

Fig. 2. Proposed transmitter structure with ISC for SISO OFDM system.

Let X_m, $0\le m\le(N/2)-1$, be the data of the mth subcarrier that satisfies the independence condition given in (1). The sequence S_m, $0\le m\le N-1$, represents the transmitted signal obtained from X_m by applying the ICM method. In the proposed scheme, the relationship between X_m and S_m is given by

$$S_m=\begin{cases}n_1\,\mathrm{Re}(X_0) & m=0,\\ n_2\,\mathrm{Im}(X_0) & m=N/2,\\ X_m & 1\le m\le(N/2)-1,\\ X^*_{N-m} & (N/2)+1\le m<N-1,\end{cases} \tag{11}$$

which can be rewritten in vector form as follows:

$$\begin{aligned}\mathbf{S}&=[S_0, S_1, \ldots, S_{(N/2)-2}, S_{(N/2)-1}, S_{N/2}, S_{(N/2)+1}, \ldots, S_{N-2}, S_{(N/2)-1}]\\ &=[n_1\,\mathrm{Re}(X_0), X_1, \ldots, X_{(N/2)-2}, X_{(N/2)-1},\\ &\qquad n_2\,\mathrm{Im}(X_0), X^*_{(N/2)-1}, \ldots, X^*_2, X^*_1],\end{aligned} \tag{12}$$

where n_1 and n_2 are the normalization constants to set $|S_m|=1$ at $m=0$ and $m=N/2$. By using the independence condition (1) and $E\{\mathrm{Re}[X_k]\mathrm{Im}[X^*_k]\}=0$ for a proper complex symbol X_k, the correlation between the signals after ICM is given by

$$E[S_mS^*_k]=\begin{cases}1 & m=k\\ 0 & m\ne k\end{cases}\quad\text{for}\quad 0\le m,\, k\le(N/2)-1. \tag{13}$$

The transmitted time-domain signal after IDFT is given by

$$s_n=\frac{1}{N}\sum_{m=0}^{N-1}S_m\mathrm{e}^{j2\pi mn/N}\quad\text{for}\quad 0\le n\le N-1. \tag{14}$$

Because the proposed ICM has a Hermitian symmetric property, such as

$$S_m=S^*_{(N-m)\bmod N}, \tag{15}$$

the time-domain signal s_n can be rewritten as

$$\begin{aligned}s_n&=\frac{1}{N}\sum_{m=0}^{(N/2)-1}(S_m\mathrm{e}^{j2\pi mn/N}+S_{N-m}\mathrm{e}^{j2\pi(N-m)n/N})\\ &=\frac{1}{N}\{n_1\,\mathrm{Re}(X_0)+n_2\,\mathrm{Im}(X_0)\}\\ &\quad+\frac{1}{N}\sum_{m=1}^{(N/2)-1}(X_m\mathrm{e}^{j2\pi mn/N}+X^*_m\mathrm{e}^{-j2\pi mn/N})\\ &=\frac{1}{N}\{n_1\,\mathrm{Re}(X_0)+n_2\,\mathrm{Im}(X_0)\}\\ &\quad+\frac{1}{N}\sum_{m=1}^{(N/2)-1}2\,\mathrm{Re}(X_m\mathrm{e}^{j2\pi mn/N}).\end{aligned} \tag{16}$$

From (16), we can see that s_n has a real value.

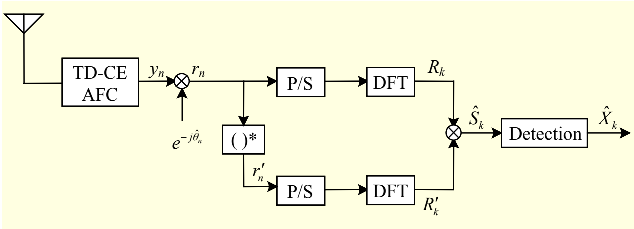

Fig. 3. Proposed receiver structure for SISO OFDM system.

Figure 3 shows the proposed receiver structure for the SISO OFDM system. The upper path is similar to the receiver structure of the conventional SISO OFDM, except that the phase offset compensation section is added to compensate for the fast-varying PHN and RFO. The lower path is added to suppress the ICI using the symmetric conjugate property of PHN, where $(\cdot)^*$ represents a conjugate operation.

After a frequency down-conversion at the receiver, the received signal passes through the time-domain channel estimator (TD-CE) and the automatic frequency controller (AFC). The joint estimation of carrier frequency offset (CFO) and channel transfer function for OFDM with PHN is discussed in [15]. Although the channel estimator estimates the Doppler frequency and channel impulse response and the AFC estimates and compensates the CFO, an RFO exists because of estimation error.

The received signal affected by PHN and RFO can be given by

$$y_n = s_n e^{j\theta_n} + w_n \quad \text{for} \quad 0 \le n \le N-1, \tag{17}$$

where θ_n is a phase offset denoted by $\theta_n = \phi_n + 2\pi n\varepsilon/N$. Here, ε is the normalized RFO and w_n is a complex AWGN with mean equal to zero and variance equal to N_0. Because s_n has a real value we can estimate θ_n from y_n as follows:

$$\theta'_n = \angle y_n, \tag{18}$$

where $\angle$ denotes the argument of a complex number.

We assume that $\theta_n << \pi/2$ because typically a local oscillator is phase locked and the initial frequency offset acquisition is achieved using the AFC. Therefore, the estimated phase offset ($\hat{\theta}_n$) is given by

$$\hat{\theta}_n = \begin{cases} \theta'_n & |\theta'| \le \pi/2, \\ \theta'_n - \pi & \pi/2 < \theta' < \pi, \\ \theta' + \pi & -\pi < \theta' < -\pi/2. \end{cases} \tag{19}$$

Because the symmetric conjugate property of the ICI weighting function is satisfied only when the phase error is small, the phase offset due to PHN and RFO must be compensated. The phase offset can be compensated from y_n using $\hat{\theta}_n$, and the phase-compensated signal can be written as follows:

$$r_n = y_n e^{-j\hat{\theta}_n} = s_n e^{j\tilde{\theta}_n} + w'_n, \quad 0 \le n \le N-1, \tag{20}$$

where $\tilde{\theta}_n = \theta_n - \hat{\theta}_n$ and $w'_n = w_n e^{-j\hat{\theta}_n}$. Using a notation similar to that in (7), the DFT output of r_n can be expressed as

$$R_k = \sum_{m=0}^{N-1} S_m \tilde{\Theta}_{k-m} + W_k, \quad 0 \le k \le N-1, \tag{21}$$

where the modified ICI weighting function is

$$\tilde{\Theta}_k = \frac{1}{N} \sum_{n=0}^{N-1} e^{j\tilde{\theta}_n} e^{-j2\pi kn/N}. \tag{22}$$

On the other hand, r'_n in the lower path of Fig. 3 is given by

$$r'_n = y_n^* e^{j\hat{\theta}_n} = s_n e^{-j\tilde{\theta}_n} \quad \text{for} \quad 0 \le n \le N-1, \tag{23}$$

and its frequency-domain signal R'_k can be expressed as

$$R'_k = \sum_{m=0}^{N-1} S_m \tilde{\Theta}^*_{-(k-m)} + W_k^* \quad \text{for} \quad 0 \le k \le N-1. \tag{24}$$

Consequently, the demodulated output is obtained as

$$\begin{aligned} \hat{S}_k &= \frac{R_k + R'_k}{2} \\ &= \frac{1}{2} S_k (\tilde{\Theta}_0 + \tilde{\Theta}_0^*) + \frac{1}{2} \sum_{m=0, m\ne k}^{N-1} S_m (\tilde{\Theta}_{k-m} + \tilde{\Theta}^*_{-(k-m)}) \\ &\quad + \mathrm{Re}\{W_k\}. \end{aligned} \tag{25}$$

From (25), we can see that the combining method of R_k and R'_k is similar to the optimal maximal-ratio combining method. The first term of (25) represents a desired signal rotated by CPE, and the second term is the ICI. The CPE can be reduced to zero using the symmetric conjugate property of the ICI weighting function (9), and the ICI term can be suppressed by property (10).

The data X_k can be estimated as follows:

$$\hat{X}_0 = \mathrm{Re}\{\hat{S}_0\} + j\,\mathrm{Re}\{\hat{S}_{N/2}\}, \tag{26}$$

$$\hat{X}_k = \frac{\hat{S}_k + \hat{S}^*_{N/2}}{2} \quad \text{for} \quad 1 \le k \le (N/2)-1. \tag{27}$$

The average CIR of the proposed scheme at the 0th subcarrier can be clearly shown to be

$$\begin{aligned} CIR_{\mathrm{pro}} &= \frac{E\left[|S_0(\tilde{\Theta}_0 + \tilde{\Theta}_0^*)|^2 \right]}{E\left[\left| \sum_{m=1}^{N-1} S_m (\tilde{\Theta}_{-m} + \tilde{\Theta}_m^*) \right|^2 \right]} \\ &= \frac{|\tilde{\Theta}_0 + \tilde{\Theta}_0^*|^2}{\sum_{m=1}^{N-1} |\tilde{\Theta}_{-m} + \tilde{\Theta}_m^*|^2}, \end{aligned} \tag{28}$$

where the second equality indicates that the sequence S_m, $m = 0, 1, \ldots, N-1$ satisfies the independence condition given by (13).

Considering the symmetric conjugate property of the ICI weighting function in (9) and (10), the numerator of (28) equals four and the denominator approximates to zero. Hence, the resulting CIR can be made to be sufficiently large so that the proposed scheme can effectively compensate for the ICI caused by PHN. For comparison, the average CIR for the 0th subcarrier of the conventional OFDM system without ISC is given in [7] as

$$CIR_{\mathrm{OFDM}} = \frac{E\left[|X_0 \Theta_{-\varepsilon}|^2 \right]}{E\left[\left| \sum_{m=1}^{N-1} X_m \Theta_{m-\varepsilon} \right|^2 \right]} = \frac{|\Theta_{-\varepsilon}|^2}{\sum_{m=1}^{N-1} |\Theta_{-m-\varepsilon}|^2}, \quad (29)$$

and the average CIR of the symmetric data conjugate method is given in [10] as

$$CIR_{\mathrm{symm}} = \frac{|\Theta_{-\varepsilon} + \Theta^{*}_{-\varepsilon}|^2}{\sum_{m=1}^{(N/2)-1} |\Theta_{-m-\varepsilon} + \Theta^{*}_{-m-\varepsilon}|^2 + \sum_{m=1}^{(N/2)-1} |\Theta_{-(N-1-m+\varepsilon)} + \Theta^{*}_{N-1-m-\varepsilon}|^2}. \quad (30)$$

When the normalized 3 dB bandwidth of the PHN is large or if an RFO exists, the denominator of (30) does not become zero. Therefore, we can obtain the following inequality relation:

$$CIR_{\mathrm{pro}} \geq CIR_{\mathrm{symm}} > CIR_{\mathrm{OFDM}}. \quad (31)$$

IV. Simulation Results

In this section, the simulation results are presented for verifying the efficiency of the proposed ISC. The average CIR and BER performance of the following three types of OFDM systems are compared:

1) SISO OFDM system without ISC: 32QAM modulation.
2) SISO OFDM system with the conventional ISC scheme [10]: 64QAM modulation.
3) SISO OFDM system with the proposed ISC scheme: 64QAM modulation.

Different types of quadrature amplitude modulation (QAM) modulations were used to obtain a spectral efficiency of 1 bit/Hz/s for the three schemes. The signal energy per information bit to noise PSD ratio (E_b/N_0) remains the same in all cases when used to examine the BER performance. Moreover, the PHN is randomly generated using independent and identically distributed Gaussian samples having a PSD given in (4) with the following parameters: $a = 6.5$, $b = 2$, $c = 12$, $\beta = 10$ kHz, and $\gamma = 100$ kHz.

The total number of subcarriers is 128, and the channel model is the Extended Pedestrian A (EPA) propagation channel model defined in the 3GPP TS36.101 standard with the parameters given in Table 1 [16]. The channel estimation is assumed to be perfect, and the channel is equalized using a zero-forcing algorithm.

Figure 4 shows the average CIRs of the different schemes as a function of the normalized 3 dB bandwidth (ρ) when the normalized RFO (ε) is equal to 0, 0.05, and 0.10, respectively. The average CIRs of the proposed ISC, conventional ISC [10], and OFDM system without ISC schemes are obtained from (28), (29), and (30), respectively. From this figure, we can see that the proposed ISC scheme has a better CIR performance than the conventional schemes and that the performance improves as ρ or ε increases. The conventional symmetric ISC scheme has a high CIR when $\rho < 1$ and $\varepsilon = 0$, but its CIR decreases as ρ and ε increase because the symmetric conjugate property of Θ_k is not satisfied for that case. The proposed ISC scheme achieves a high CIR regardless of the values of ρ and ε because the PHN and RFO can be estimated and compensated

Table 1. EPA channel model.

Tap	Relative delay (ns)	Average power (dB)
1	0	0
2	30	–1.0
3	70	–2.0
4	90	–3.0
5	110	–8.0
6	190	–17.2
7	410	–20.8

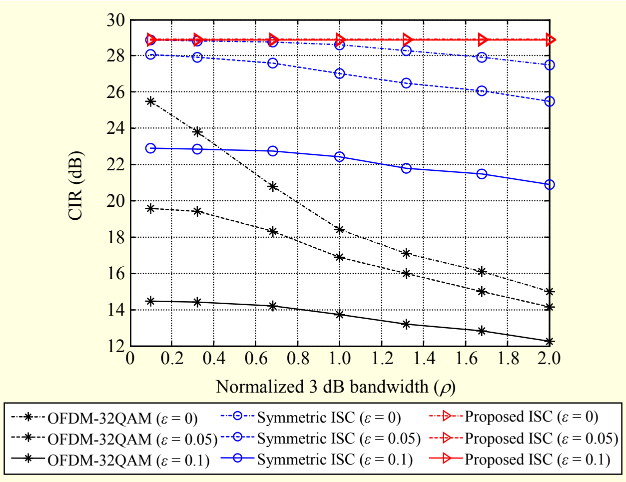

Fig. 4. Average CIR vs. normalized 3 dB bandwidth of PHN (ρ).

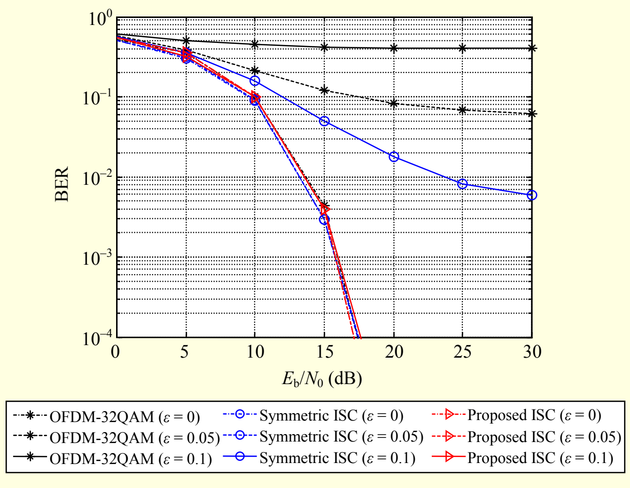

Fig. 5. BER vs. E_b/N_0 under AWGN channel when ρ = 0.1.

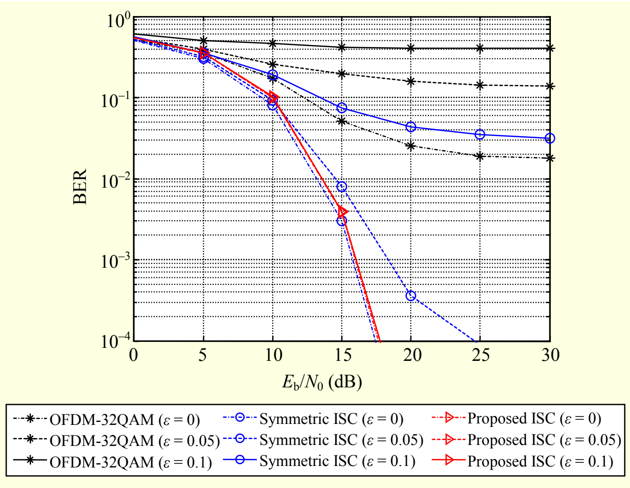

Fig. 6. BER vs. E_b/N_0 under AWGN channel when ρ = 1.

using the real property of the transmitted signal s_n.

Figures 5, 6, and 7 show the BER performance under the AWGN channel for values of ρ equal to 0.1, 1, and 2, respectively. When ρ = 0.1 and ε = 0, the BER performances of the conventional schemes are nearly identical to that of the proposed ISC scheme. However, the performance of the conventional symmetric ISC scheme degrades as ρ or ε increases. Especially when ρ = 2, the BER performance of the conventional symmetric ISC scheme is severely degraded — even in the absence of RFO. From Figs. 5, 6, and 7, we can observe that there is an error floor in the high E_b/N_0 region for the conventional schemes.

Figures 8, 9, and 10 show the BER performance under the EPA channel for ρ = 0.1, ρ = 1, and ρ = 2, respectively. From these figures, we can see that the proposed scheme achieves approximately a 1 dB to 2 dB gain at BER = 10^{-2} when compared with the conventional symmetric ISC scheme for ε = 0.05 and $\rho \geq 1$.

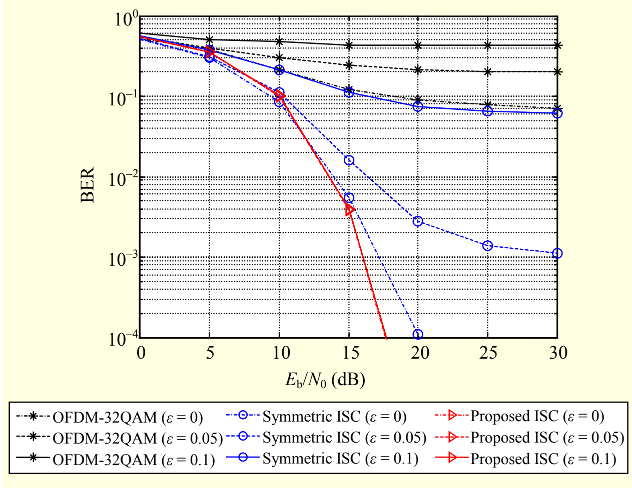

Fig. 7. BER vs. E_b/N_0 under AWGN channel when ρ = 2.

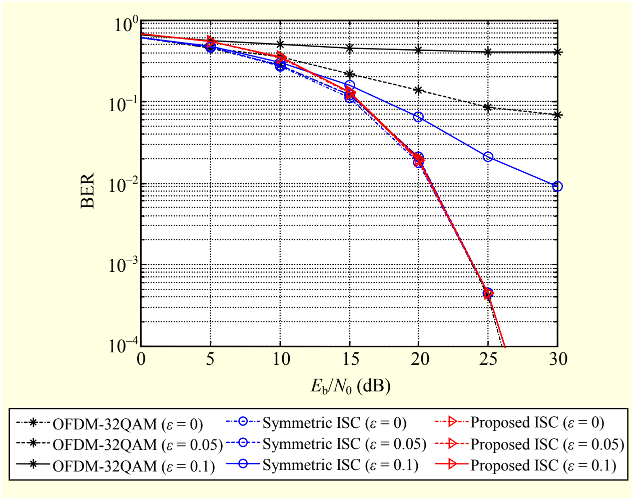

Fig. 8. BER vs. E_b/N_0 under EPA channel when ρ = 0.1.

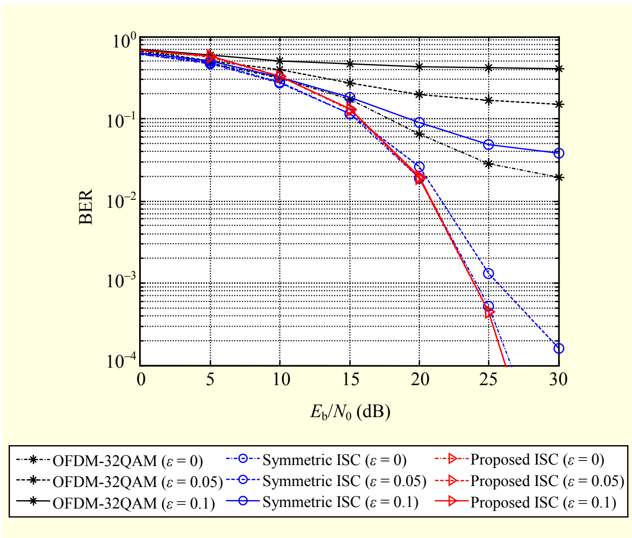

Fig. 9. BER vs. E_b/N_0 under EPA channel when ρ = 1.

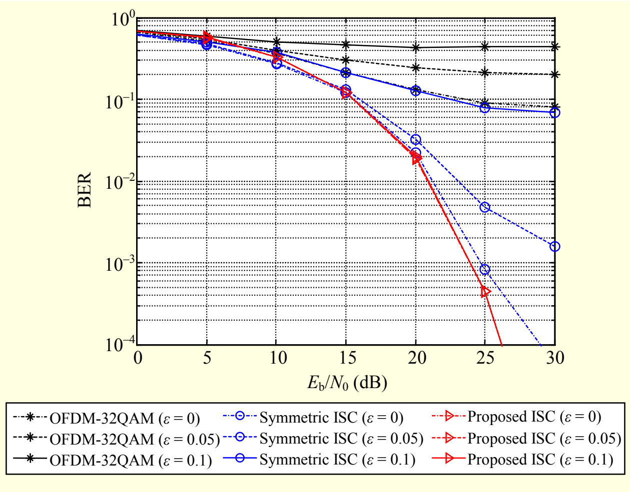

Fig. 10. BER vs. E_b/N_0 under EPA channel when $\rho = 2$.

V. Conclusion

In OFDM systems, PHN and RFO cause ICI in the received signal and are the main sources of performance degradation. In this paper, we proposed a novel ISC scheme to compensate the phase offset caused by the PHN and the RFO. In the proposed scheme, the transmitted signal is transformed into a real signal through a new ICM at the transmitter. By using the real property of the transmitted signal, PHN and RFO have been estimated and compensated at the receiver. Therefore, the ICI caused by the PHN and RFO can be significantly suppressed. Through the simulation results, we showed that the proposed scheme has a higher CIR and better BER performance than the conventional symmetric ISC scheme given in [10] and that the performance of the proposed scheme improves significantly as the normalized 3 dB bandwidth of PHN or RFO increases.

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