Phase Noise Estimation and Compensation Algorithms for 5G Systems

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Abstract. In the 5G system, orthogonal frequency division multiplexing (OFDM) waveform survived for its superior performance. As is well known, OFDM systems are sensitive to the phase noise introduced by local oscillators and it may get worse for likely higher carrier frequency in 5G systems. There are two aspects of the impact of phase noise, namely the common phase error (CPE) and the inter-carrier-interference (ICI). In this paper, first, we propose a more accurate way to estimate CPE. Then, we focus on ICI cancellation. To simplify the ICI model, we only consider the interference from adjacent sub-carriers. Based on the simplified model, we propose two schemes to estimate ICI. The performance of phase noise compensation algorithm we proposed is presented. Simulation results show that the algorithm we proposed can significantly reduce the impact of phase noise and improve the throughput of 5G systems.

Keywords: Phase noise · CPE estimation · ICI cancellation

1 Introduction

As a new generation of mobile communication technology, 5G has great innovations on data transmission rate, system bandwidth, carrier frequency, etc. Different from long term evolution (LTE), 5G new radio (NR) may support higher carrier frequency, such as 28 GHz and 40 GHz. Generally, the higher the carrier frequency, the stricter the requirements on the radio frequency (RF) circuits, and the more serious the phase noise. By default, for per decade increase in carrier frequency, power spectral density (PSD) increases by 20 dBc/Hz [1]. 3GPP working group clearly stipulates that when the carrier frequency is greater than 6 GHz, 5G systems must consider the impact of phase noise, and hence introduces a new phase tracking reference signal (PT-RS) for phase noise compensation [2].

Phase noise has always been a problem in high-frequency communication systems and there have been many studies. Reference [3] estimates phase noise and

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channel simultaneously. Although the channel and phase noise models have been simplified, the algorithm still requires a large number of pilots, and the iterative process is complex. Reference [4] selects orthogonal discrete cosine transform as the basis function to model the phase noise, and obtain the data and phase noise parameters iteratively according to the maximum likelihood criterion. There is a certain model error and multiple iterations are needed to obtain better results. Reference [5] uses a filter with a certain number of coefficients for extraction of phase noise and use scattered pilots to estimate coefficients. This method approximates the phase noise on condition that phase noise is very small, when the phase noise is large, this approximate method is not applicable. Reference [6] estimates the fast fourier transform (FFT) components of the current phase noise realization and suppress the inter-carrier-interference (ICI) by performing a deconvolution in the frequency domain. This method requires the correlation between ICI are known. Reference [7] proposes two phase noise cancellation schemes. The first one is based on linear interpolation of the common phase error (CPE) values over adjacent orthogonal frequency division multiplexing (OFDM) symbols. The second one improves the ICI estimation based on the methods presented in [8] by improving the accuracy of phase noise estimation at symbol boundaries using proper interpolation. Based on complex models and iterative operations, existing algorithms can eliminate the effects of phase noise including CPE and most of the ICI successfully. In practical systems, ICI introduced by phase noise is not very serious and does not need to be completely eliminated. Therefore, effective ICI cancellation schemes with lower complexity is needed.

In this paper, we propose an advanced CPE estimation method which takes advantage of a better merger scheme. Next, based on a simplified model which only consider adjacent sub-carrier interference, we propose an ICI cancellation algorithm. In this algorithm, we propose two solutions for ICI estimation and they are independent estimation and joint estimation respectively. Simulation results show that eliminating only the interference from adjacent subcarriers can greatly improve the performance of the system, especially when the phase noise is large. Simultaneously, the algorithm proposed in this paper is based on the pilot structure of the 5G standard and is easy to be extended and implemented in 5G systems.

This paper is organized as follows. Section 2 introduces the system model. Section 3 introduces a CPE estimation method and proposes an enhanced one. Section 4 explores the ICI cancellation algorithm with two solutions. Finally, simulation results are presented for comparison.

2 System Model

In OFDM systems, if only consider the phase noise at the receiving end, the baseband time domain receiving signal can be given as

$$y[n] = (x[n] \otimes h[n])\mathrm{e}^{j\phi[n]} + w[n], \tag{1}$$

where $x[n]$ denotes the time domain transmitting sequence, $h[n]$ indicates the time domain channel, $\mathrm{e}^{j\phi[n]}$ refers to the phase noise, $w[n]$ is the additive noise and $\otimes$ represents circular convolution. Denoting the FFT of transmitted signal, channel, phase noise and additive noise on subcarrier $k\,(0 \le k \le N-1)$ as X_k, H_k, Φ_k and W_k, the received signal in the frequency domain can be given as

$$Y_k = X_k H_k \underbrace{\Phi_0}_{CPE} + \underbrace{\sum_{i=0, i\neq k}^{N-1} X_i H_i \Phi_{(k-i)_N}}_{ICI} + W_k, \tag{2}$$

where

$$\Phi_k = \frac{1}{N}\sum_{n=0}^{N-1} e^{j\phi[n]} e^{-j\frac{2\pi}{N}nk}, \tag{3}$$

N denotes the number of system FFT points and $(\cdot)_N$ represents modulo N. As shown in (2), the phase noise has two main impacts: 1. CPE, which causes common phase rotation in constellations of received symbols. Because the magnitude of CPE is close to 1, we define it as $e^{j\theta}$; and 2. ICI, which breaks the orthogonality of OFDM waveform.

3 CPE Estimation

To estimate CPE, 5G systems introduce a new reference signal PT-RS [2]. An example of reference signals' structure in a resource block (RB) is illustrated in Fig. 1.

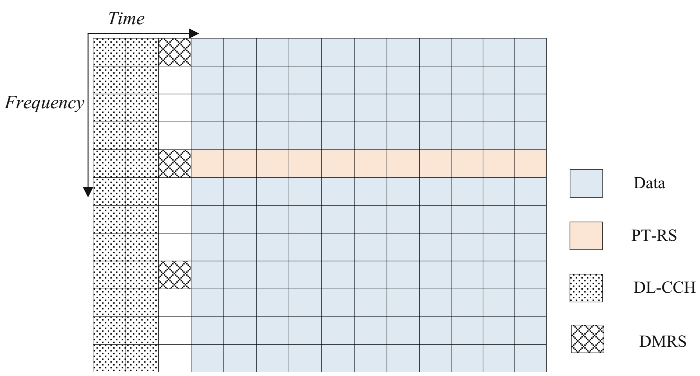


Fig. 1. Example of structure of reference signals.

In Fig. 1, dedicated demodulation reference signals (DM-RS) used for channel estimation are placed at the third OFDM symbol. Meanwhile, there are one PT-RS in each OFDM of one RB. Next, we will study CPE estimation scheme based on the structure of reference signals illustrated in Fig. 1.

According to (2) and assuming the channel is invariant during a slot, the received signals for sub-carrier k in the third and l-th symbol ($l > 3$) can be obtained as

$$Y_{k,3} = X_{k,3} H_{k,3} e^{j\theta_3} + W_{k,3} \tag{4}$$

$$Y_{k,l} = X_{k,l} H_{k,3} \mathrm{e}^{j\theta_l} + W_{k,l}, \tag{5}$$

where ICI is considered as a part of additive noise. Denoting the total number of PT-RS in frequency domain M, the difference of CPE in symbol 3 and l ($l > 3$) can be obtained as [9]

$$\theta_\tau = \frac{1}{M} \sum\nolimits_{k=1}^{M} \text{angle}\left\{Y_{k,3}^{*} Y_{k,l}\right\}, \tag{6}$$

where Y^* represents the conjugation of Y. Through rotating phase angle θ_τ based on $H_{k,3}\mathrm{e}^{j\theta_3}$ which obtained by channel estimation, we can get the channel affected by phase noise of all data symbols. In case the channel is additive white gaussian noise (AWGN), we ought to calculate CPE of every symbol alone.

In (6), each PT-RS in same symbols estimates one CPE difference and θ_τ is the average value of all. This combing method do not consider the amplitude of vector $Y_{k,3}^{*} Y_{k,l}$. In fact, the larger the amplitude of the vector, the smaller the impact of additive noise relatively and as a result, the CPE difference is more accuracy. Thus, we give a better scheme to obtain θ_τ

$$\theta_\tau = \text{angle}\left\{\sum\nolimits_{k=1}^{M} Y_{k,3}^{*} Y_{k,l}\right\}. \tag{7}$$

4 ICI Cancellation

When the phase noise is large, the influence of ICI cannot be ignored. Like ICI introduced by doppler shift, interference from adjacent sub-carriers is the maximum. Meanwhile, ICI introduced by distant sub-carriers is very small. Therefore, we just estimate adjacent two sub-carriers' interference and the interference from rest sub-carriers is considered as a part of additive noise. In order to estimate ICI using PT-RS accurately, as shown in Fig. 2, we introduce a new reference signal structure, in which subcarriers adjacent to PT-RS are left blank.

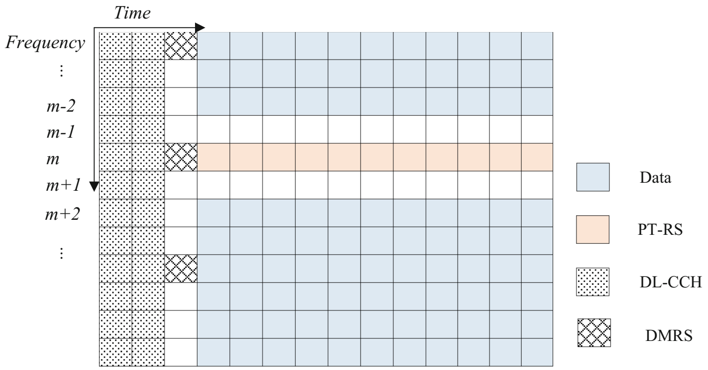


Fig. 2. Structure of reference signal for ICI reduction.

As shown in Fig. 2, we assume the PT-RS is placed at subcarrier m and the index of vacant subcarriers is $m-1$ and $m+1$ consequently. For the subcarrier $m-1$, since it has no data of its own, the signals it receives are composed of interference from subcarriers $m-2$, m and additive noise. So, according to (2), the received signals in vacant subcarriers in one OFDM symbol can be given as

$$Y_{m-1} = X_{m-2}H_{m-2}\Phi_1 + X_mH_m\Phi_{N-1} + W_{m-1}, \tag{8}$$

$$Y_{m+1} = X_mH_m\Phi_1 + X_{m+2}H_{m+2}\Phi_{N-1} + W_{m+1}. \tag{9}$$

When the channel is AWGN, Eqs. (8) and (9) becomes

$$Y_{m-1} = X_{m-2}\Phi_1 + X_m\Phi_{N-1} + W_{m-1}, \tag{10}$$

$$Y_{m+1} = X_m\Phi_1 + X_{m+2}\Phi_{N-1} + W_{m+1}. \tag{11}$$

To solve Φ_1 and Φ_{N-1}, in addition to the known PT-RS X_m, data X_{m-2} and X_{m+2} are also needed. Thus, in this paper, the ICI estimation and cancellation process are performed after the soft demodulation of receiver end. When an OFDM symbol does not configure PT-RS, both CPE and ICI of this symbol can be obtained by interpolation.

When the channel is not AWGN, through channel estimation and CPE compensation, we can obtain $H_{m-2}\Phi_0$, $H_m\Phi_0$ and $H_{m+2}\Phi_0$. Equations (8) and (9) now can be rewritten as

$$Y_{m-1} = X_{m-2}H_{m-2}\Phi_0\Phi_1/\Phi_0 + X_mH_m\Phi_0\Phi_{N-1}/\Phi_0 + W_{m-1}, \tag{12}$$

$$Y_{m+1} = X_mH_m\Phi_0\Phi_1/\Phi_0 + X_{m+2}H_{m+2}\Phi_0\Phi_{N-1}/\Phi_0 + W_{m+1}. \tag{13}$$

Thus, when considering real channel, just estimate Φ_1/Φ_0, Φ_{N-1}/Φ_0 instead of Φ_1, Φ_{N-1} and the remaining steps remain unchanged.

Of course, the structure of reference signals is not only one form of Fig. 2. Through double the number of PT-RS or just keep it like Fig. 1, ICI can also be calculated without leaving blank. The method of leaving blank increases the accuracy of ICI estimation at the expense of system throughput and others are the opposite. In this paper, we focus on how to estimate ICI more accurately, and two solutions under AWGN channel are introduced next in detail.

4.1 Calculate Φ_1 and Φ_{N-1} Independently

When there is only one subcarrier to place PT-RS in the frequency domain, namely $M = 1$, we define the received signals in (10) and (11) as $\mathbf{Y}$ and it can be obtained by

$$\begin{bmatrix} Y_{m-1} \\ Y_{m+1} \end{bmatrix} = \mathbf{A} \begin{bmatrix} \Phi_1 \\ \Phi_{N-1} \end{bmatrix} + \begin{bmatrix} W_{m-1} \\ W_{m+1} \end{bmatrix}, \tag{14}$$

where

$$\mathbf{A} = \begin{bmatrix} X_{m-2} & X_m \\ X_m & X_{m+2} \end{bmatrix}. \tag{15}$$

Here, we define the average power of W is σ^2 and assume Φ_1 and Φ_{N-1} are independent and their average power both are P. A standard minimum mean square error (MMSE) estimation can be utilized

$$\begin{bmatrix} \Phi_1 \\ \Phi_{N-1} \end{bmatrix} = \mathbf{V}_{\mathrm{MMSE}}\mathbf{Y}, \tag{16}$$

where

$$\mathbf{V}_{\mathrm{MMSE}} = \left(\mathbf{A}^{\mathrm{H}}\mathbf{A} + \frac{\sigma^2}{P}\mathbf{I}\right)^{-1}\mathbf{A}^{\mathrm{H}}. \tag{17}$$

Since the average power of PT-RS and data is both 1, the inverse of $\mathbf{A}^{\mathrm{H}}\mathbf{A}$ may not exist and this method effectively avoid this problem.

When $M > 1$, according to (16), we can obtain M pair of Φ_1 and Φ_{N-1}. Because averaging complex number directly makes no sense, it is not suggested to calculate ICI by each PT-RS separately and then average them as the final result. In this paper, we propose to merge items obtained by all PT-RS in the same OFDM symbol firstly and get one final result. Namely, the dimension of $\mathbf{A}$ and $\mathbf{Y}$ is extended to be $2M \times 2$ and $2M \times 1$. Since all PT-RS are merged, the value of $\frac{\sigma^2}{P}\mathbf{I}$ is very small relative to $\mathbf{A}^{\mathrm{H}}\mathbf{A}$ and the probability of $\mathbf{A}^{\mathrm{H}}\mathbf{A}$ irreversible is almost zero. In practical situations, Φ_1 and Φ_{N-1} are necessarily independent of each other. Therefore, we suggest use zero forcing (ZF) scheme in place of MMSE scheme in this condition, namely $\mathbf{V}_{\mathrm{ZF}} = \left(\mathbf{A}^{\mathrm{H}}\mathbf{A}\right)^{-1}\mathbf{A}^{\mathrm{H}}$. The subsequent simulations of this paper also use ZF scheme.

In this section, we propose a method to calculate ICI independently and discuss the situations under different PT-RS numbers. This method is simple and unaffected by the size of the phase noise.

4.2 Calculate Φ_1 and Φ_{N-1} Jointly

The actual phase noise consists of CPE and the variable part relative to it. Hence, the phase noise in an OFDM symbol can be represented as

$$\phi[n] = \mathrm{angle}(\Phi_0) + \Delta\phi[n], \tag{18}$$

and further we can get

$$\mathrm{e}^{j\phi[n]} = \mathrm{e}^{j(\mathrm{angle}(\Phi_0)+\Delta\phi[n])} \approx \Phi_0(1 + j\Delta\phi[n]). \tag{19}$$

This approximation method is more accurate than that in [5], which consider $\mathrm{e}^{i\phi[n]} \approx 1 + j\Delta\phi[n]$, especially when the phase noise is large. According to the definition of Φ, we can get

$$\begin{aligned} \Phi_1 &= \frac{1}{N}\sum\nolimits_{n=0}^{N-1} \Phi_0(1+j\Delta\phi[n])\,\mathrm{e}^{-j\frac{2\pi}{N}n} \\ &= \frac{1}{N}\sum\nolimits_{n=0}^{N-1} \Phi_0 \mathrm{e}^{-j\frac{2\pi}{N}n} + j\frac{1}{N}\Phi_0\sum\nolimits_{n=0}^{N-1}\Delta\phi[n]\mathrm{e}^{-j\frac{2\pi}{N}n} \\ &= j\frac{1}{N}\Phi_0\sum\nolimits_{n=0}^{N-1}\Delta\phi[n]e^{-j\frac{2\pi}{N}n}, \end{aligned} \tag{20}$$

$$\begin{aligned}\Phi_{N-1} &= \frac{1}{N}\sum\nolimits_{n=0}^{N-1}\Phi_0\left(1+j\Delta\phi[n]\right)\mathrm{e}^{j\frac{2\pi}{N}n} \\ &= \frac{1}{N}\sum\nolimits_{n=0}^{N-1}\Phi_0\mathrm{e}^{j\frac{2\pi}{N}n} + j\frac{1}{N}\Phi_0\sum\nolimits_{n=0}^{N-1}\Delta\phi[n]e^{j\frac{2\pi}{N}n} \\ &= j\frac{1}{N}\Phi_0\sum\nolimits_{n=0}^{N-1}\Delta\phi[n]e^{j\frac{2\pi}{N}n}. \end{aligned} \tag{21}$$

By comparing (20) and (21), the relationship between Φ_1 and Φ_{N-1} can be obtained as

$$\Phi_{N-1} = -{\Phi_0}^2 \cdot conj(\Phi_1). \tag{22}$$

Using the relationship between Φ_1 and Φ_{N-1}, Eqs. (10) and (11) can solve a result separately. In detail, when there is only one subcarrier to place PT-RS in the frequency domain, we define $\Phi_0 = c + dj$, $\Phi_1 = a + bj$, $\Phi_{N-1} = -\Phi_0^2(a - bj)$, $X_k = x_k^r + x_k^i j$, $Y_k = y_k^r + y_k^i j$ and $W_k = w_k^r + w_k^i j$. By solving (10), we get $\mathbf{Y}'$

$$\begin{bmatrix} y_{m-1}^r \\ y_{m-1}^i \\ y_{m+1}^r \\ y_{m+1}^i \end{bmatrix} = \mathbf{A}'\begin{bmatrix} a \\ b \end{bmatrix} + \begin{bmatrix} w_{m-1}^r \\ w_{m-1}^i \\ w_{m+1}^r \\ w_{m+1}^i \end{bmatrix}, \tag{23}$$

where

$$\mathbf{A}' = \begin{bmatrix} x_{m-2}^r - \left(c^2 - d^2\right)x_m^r + 2cdx_m^i & -x_{m-2}^i - \left(c^2 - d^2\right)x_m^i - 2cdx_m^r \\ x_{m-2}^i - \left(c^2 - d^2\right)x_m^i - 2cdx_m^r & x_{m-2}^r + \left(c^2 - d^2\right)x_m^r - 2cdx_m^i \\ x_m^r - \left(c^2 - d^2\right)x_{m+2}^r + 2cdx_{m+2}^i & -x_m^i - \left(c^2 - d^2\right)x_{m+2}^i - 2cdx_{m+2}^r \\ x_m^i - \left(c^2 - d^2\right)x_{m+2}^i - 2cdx_{m+2}^r & x_m^r + \left(c^2 - d^2\right)x_{m+2}^r - 2cdx_{m+2}^i \end{bmatrix}. \tag{24}$$

Same as Sect. 4.1, we can solve a and b by the MMSE criterion. When $M > 1$, also same as Sect. 4.1, merge all PT-RS firstly and the dimension of $\mathbf{A}'$ and $\mathbf{Y}'$ is extended to be $4M \times 2$ and $4M \times 1$ respectively.

In this section, by seeking the relationship between Φ_1 and Φ_{N-1}, we estimate ICI jointly. Compared with estimating ICI independently, the items can be used to merge is double. However, the relationship between Φ_1 and Φ_{N-1} relied on by the joint estimation is obtained by approximation. There is a certain model error when estimating ICI jointly, especially when the phase noise is large, and the independent estimation method has no such problem.

5 Simulations and Performance Analysis

In this section, we make simulations for comparison and analysis. Main assumptions are shown in Table 1.

Table 1. Simulation assumption.

Parameters	Assumptions
Carrier frequency	30 GHz
Subcarrier spacing	120 kHz
Allocated bandwidth	100e6 Hz
UE speed	0 km/h
Coding scheme	Turbo

5.1 CPE Estimation

In this part, we compare the CPE estimation scheme in [9] and proposed in this paper with scheduled bandwidth 32 RBs under CDL-B channel. The PSD of phase noise is −70, −70, −140 and −140 dBc/Hz in frequency 0, 10e3, 1e6 and 9e9 Hz corresponding. Specific definition of phase noise PSD can refer to [10]. The modulation method is 256QAM and the code rate is 0.75. With the PT-RS FD of every 4th RB and TD of every 1 symbol, the simulation results are given as follow.

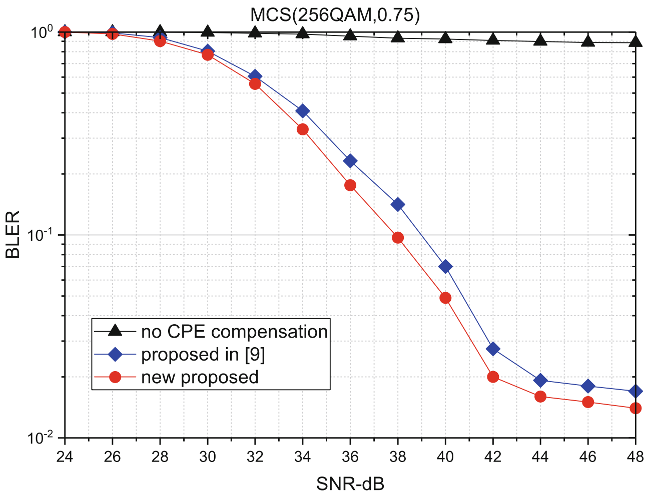


Fig. 3. Simulation results of CPE estimation.

In Fig. 3, no CPE compensation represents the system does not configure PT-PS. As can be seen from Fig. 3, the method we proposed can decrease the block error ratio (BLER) about 1 dB compared with the method in [9]. High-order modulation is very sensitive to phase rotation, so no CPE compensation leads to BLER remain high. The impact of ICI on high-order modulation systems is relatively large, and therefore, in Fig. 3, when the SNR is high, the BLER without ICI cancellation tends to be flat.

5.2 ICI Cancellation

In order to highlight the effect of ICI cancellation, we increase the ICI of the system by changing the PSD of phase noise. In detail, the frequencies are adjusted to 0, 10e4, 10e6, 9e9 Hz and 0, 50e3, 1e6, 9e9 Hz in AWGN and CDL-B channel respectively, and other configurations remain unchanged. When the channel is CDL-B, in order to exclude the effect of channel estimation error on the result of ICI cancellation, we perform simulations under both ideal channel estimation and exponential power delay profile channel estimation. Simulation results under AWGN and CDL-B channel are given as follows.

As can be seen from the Figs. 4 and 5, the two ICI estimation solutions proposed in this paper both are effective, and the second solution is about 2 dB better than the first in AWGN channel. The results indicate that, compared with estimating ICI independently, though estimating ICI jointly has model error, double the number of items using to merge make the ICI estimation more accurate and ultimately lead to a better result.

In Figs. 4 and 5(b), when the SNR is high, due to the impact of the remaining ICI, BLER tends to be flat. When the modulation mode changes from 256QAM to 64QAM, as shown in Fig. 5(a), BLER can be reduced to 0 and smaller performance gains from ICI elimination. Therefore, it can be concluded that when the modulation order is high, only eliminating adjacent subcarriers' interference can significantly improve the system performance, but the remaining ICI still impair performance. When the modulation order is low, the impact of ICI is relatively small and the effects of distant ICI can be ignored. According to several experiments, increasing the number of iterations does not make the results better, so only one iteration is sufficient.

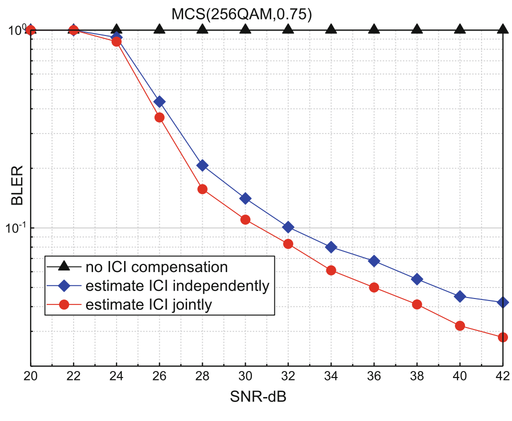
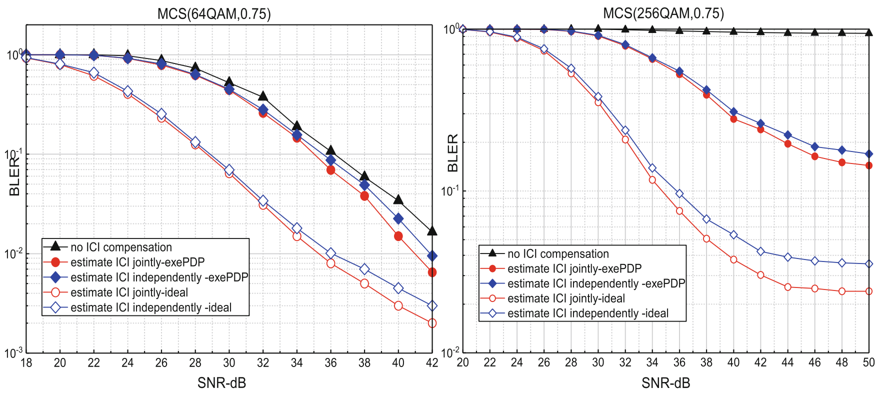


Fig. 4. Simulation of ICI cancellation under AWGN with code rate 0.75.



(a) 64QAM (b) 256QAM

Fig. 5. Simulations of ICI cancellation under CDL-B channel with code rate 0.75.

6 Conclusion

In this paper, based on the method proposed in [9], we proposed a new method to estimate CPE. Simulation Results indicate that when there are multiple subcarriers in the frequency domain placing PT-RS, considering the influence of vector's amplitude makes CPE estimation more accurate.

Based on the CPE estimation, we propose an ICI cancellation algorithm with low complexity. In this algorithm, we use a simplified ICI model which only consider the interference from adjacent subcarriers and propose two solutions to estimation ICI. Simulations show that make use of the relationship between ICI is better than estimate them independently and hence, it can be concluded that the number of the items used to merge has a great influence on the ICI estimation. Meanwhile, the simulations show that the performance of the algorithm we proposed are related to the PSD of phase noise and MCS. When the MCS is low, only eliminating the adjacent subcarriers' interference is sufficient. When the MCS is high, the remaining ICI will affect the performance of the system.

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