A Novel MIMO-OFDM Adaptive Receiver in Time-Varying Channels

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Abstract: This paper proposes an advanced receiver with joint inter-carrier interference (ICI) self cancellation and channel equalization for multiple-input multiple-output orthogonal frequency division multiplexing (MIMO-OFDM) systems over rapidly time-varying channel environment. The existing schemes that deal with the ICI cancellation and channel equalization can't provide satisfactory performance over time-varying channels. In term of error rate performance and low computational complexity, ICI self cancellation is the best choice. So, this paper proposes an adaptive receiver to deal with the problem of joint ICI self cancellation and channel equalization. We employ the adaptive phase rotations in the receiver to effectively track the CFO variations without feeding back the CFO estimate to the transmitter as required in traditional existing scheme. We also give some simulations to verify the proposed scheme.

Key words: MIMO-OFDM, inter-carrier interference, ICI self cancellation, channel estimation.

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1. Introduction

The MIMO transmission system is employed to increase the transmitted data rate and improve the communication link quality compared with single-antenna system[1]-[3]. Furthermore, the Space-Time Block codes provides temporal/spatial multiplexing and have been employed to enhance the reliability of MIMO wireless communication systems and improve the performance of MIMO communications [4]. The OFDM technique can effectively combat the problem of frequency selective fading and can provide the advantages of high-data-rate transmission and high spectral efficiency [5], [6]. Many wireless standards have employed the OFDM technique [7], [8]. The main problem of the MIMO-OFDM system is that the system is highly sensitive to the CFO.

In order to effectively deal with the problem of the CFO, the estimation the CFO must be derived and the CFO compensation is then performed that is according to the estimated results [9]-[11]. For OFDM communication systems, based on the assumption that the channel variation is with approximate linearity in an OFDM symbol, [12]-[14] proposed the ICI cancellation schemes that use training sequences and pilot symbols to perform frequency-domain channel estimation/equalization over fast fading multipath channels. The conjugate cancellation (CC) [15] and phase rotated conjugate cancellation (PRCC) [16] are the two most important ICI self-cancellation schemes. For the CC scheme, the first path transmits the original OFDM symbol, and the second path transmits the conjugate version of the OFDM symbol. Based on the configuration, we find that the ICI in the first path is restrained by the second path. For the PRCC scheme, by applying a phase rotation to the transmitter in the CC scheme, we find that the ICI caused by both paths can be mutually cancelled more effectively. In this paper, we propose an adaptive receiver (adaptive modified PRCC scheme) that can overcome the above mentioned problem. Furthermore, the existing two-stage IQ imbalance scheme [17] uses a two-stage procedure to estimate and compensate the CFO and the IQ imbalance for MIMO-OFDM communications. The phase error problem is also not considered in the two-stage IQ imbalance scheme. So the two-stage IQ imbalance scheme can't provide satisfactory performance under environments with large CFOs. We propose an adaptive receiver for ICI self-cancellation by taking advantage of conjugate transmission [18] for the OFDM communications. Unlike the existing PRCC method, the proposed scheme uses two phase rotations on both paths at the receiver to cope with the fast fading channel. Furthermore, in order to obtain the formula of between the CFO and the phase rotation, we also derive the optimal phase rotations based in the maximization of the carrier-to-interference ratio (CIR), and also propose an adaptive process employing the normalized block least mean-squared (BLMS) algorithm [19]-[21]. [22] proposed a precoding-based blind channel estimation for MIMO-OFDM systems. This scheme doesn't solve the problem of the CFO and can't be applied to the time-varying channel. [23] proposed a decision-directed channel estimation scheme to deal with the shortage of pilot for the MIMO-OFDM systems [23]. To practically analyze and effectively solve the CFO problem of MIMO-OFDM systems over time-varying channels, this paper joint considers the channel estimation/equalization based on the Kalman algorithm [24], [25], minimum mean square error (MMSE) equalization [26] and the CFO compensation that employs an adaptive modified PRCC receiver [18]. Based on the signal subspace of the channel samples' correlation matrix, the estimation of channel parameters can be translated into an unconstrained minimization problem. Then, in order to solve this optimization problem, a subspace tracking by Kalman filter is carried out which is characterized in the state equation and the measurement equation.

2. System Model

The considered MIMO-OFDM transmitter is described as Fig. 1. The proposed receiver is described as Fig. 2. The time-domain received signals of the two paths can be expressed as follows,

$$\mathbf{y}^{(1)}_{\alpha,l,n} = \sum_{\beta=1}^{N_t}\sum_{p=0}^{L-1} d_{\beta,(1-p),n} h_{\beta,\alpha,p,n} e^{j\frac{2\pi}{N} l\varepsilon_{\alpha,n}} + w^{(1)}_{\alpha,l,n} \quad l = 0,1,\ldots,N-1 \tag{1}$$

$$\mathbf{y}^{(2)}_{\alpha,l,n} = \sum_{\beta=1}^{N_t}\sum_{p=0}^{L-1} d^{*}_{\beta,(1-p),n} h_{\beta,\alpha,p,n} e^{j\frac{2\pi}{N} l\varepsilon_{\alpha,n}} + w^{(2)}_{\alpha,l,n} \quad l = 0,1,\ldots,N-1 \tag{2}$$

where L is the number of channel tap, $\mathbf{y}^{(i)}_{\alpha,l,n}$ and $w^{(i)}_{\alpha,l,n}$ correspond to the received symbol and noise at the α -th receiving antenna, l -th subcarrier in the i -th path within the n -th transmission block before the FFT demodulation, respectively. $h_{\beta,\alpha,p,n}$ corresponds to the channel tap at the β -th transmitting antenna, α -th receiving antenna and p-th path within the n -th transmission block, and then $\beta = 1,2,\ldots,N_t$ and $\alpha = 1,2,\ldots,N_r$, N_t is the total number of transmitting antenna, N_r is the total number of receiving antenna. If we employ 2×2 STBC encoder that $d_{1,l,n}$ is the 2x2 STBC encoder output at the first antenna in the l-th subcarrier within the n -th transmission block, we obtain

$$\begin{bmatrix} d_{1,l,n} \\ d_{2,l,n} \end{bmatrix} = \begin{bmatrix} x_0 & -x_1^* & \dots & x_{N-2} & -x_{N-1}^* \\ x_1 & x_0^* & \dots & x_{N-1} & x_{N-2}^* \end{bmatrix} \tag{3}$$

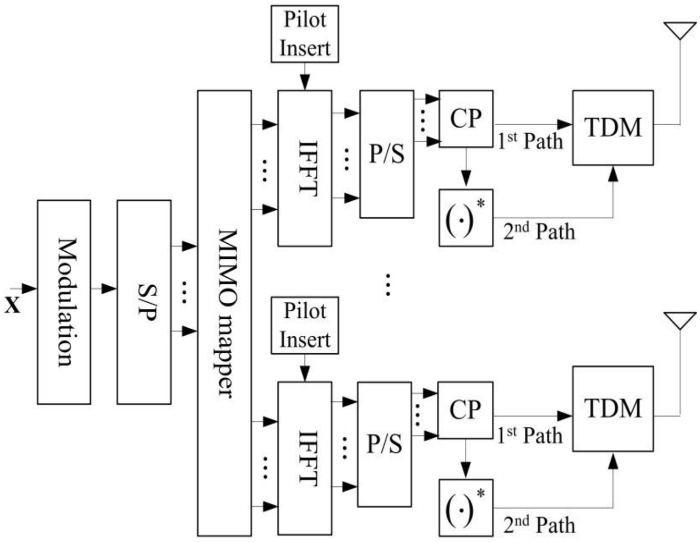

Fig. 1. The transmitter.

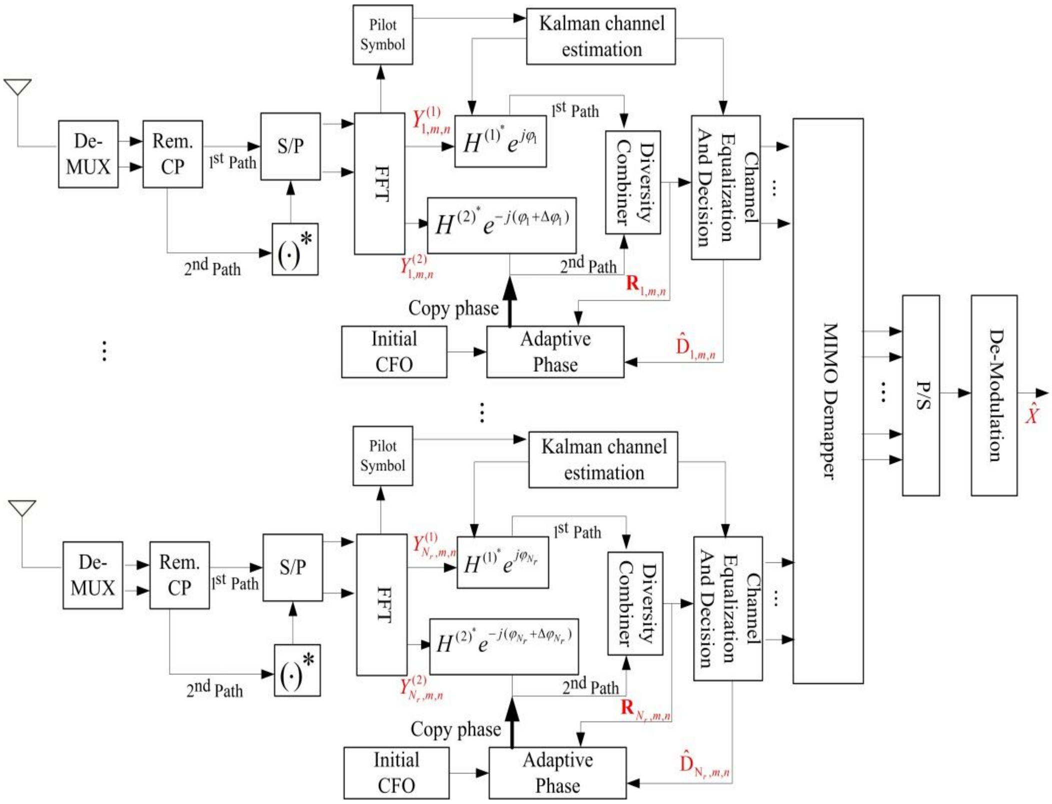

Fig. 2. The proposed receiver.

The frequency-domain received signals of these two paths can be written as follows,

$$\mathbf{Y}^{(1)}_{\alpha,m,n} = \sum_{\beta=1}^{N_t} D_{\beta,m,n} H_{\beta,\alpha,m,n} C(-\varepsilon_{\alpha,n}) + \sum_{\beta=1}^{N_t} \sum_{\substack{i=0 \\ i\neq m}}^{N-1} D_{\beta,\alpha,m,n} H_{\beta,m,n} C(m-i-\varepsilon_{\alpha,n}) + W^{(1)}_{\alpha,l,n} \quad m=0,\dots,N-1 \tag{4}$$

$$\mathbf{Y}^{(2)}_{\alpha,m,n}=\sum_{\beta=1}^{N_t}D_{\beta,m,n}H_{\beta,\alpha,m,n}C(-\varepsilon_{\alpha,n})+\sum_{\beta=1}^{N_t}\sum_{\substack{i=0\\i\neq m}}^{N-1}D_{\beta,m,n}H_{\beta,\alpha,m,n}C(m-i-\varepsilon_{\alpha,n})+W^{(2)}_{\alpha,l,n}\quad m=0,\ldots,N-1 \tag{5}$$

where $D_{\beta,m,n}$ is the FFT of 2×2 STBC encoder output. By combining and averaging the signal processed in (4) and (5), we obtain

$$\mathbf{R}_{\alpha,m,n}=\frac{1}{2}\left(\mathbf{Y}^{(1)}_{\alpha,m,n}+\mathbf{Y}^{(2)}_{\alpha,m,n}\right) \tag{6}$$

The CIR of the CC scheme in 2×2 MIMO-OFDM can be defined as

$$CIR_{\alpha,CC}=\frac{\left|C(-\varepsilon_{\alpha,n})+C(\varepsilon_{\alpha,n})\right|^2}{\sum_{i=1}^{N-1}\left|C(i-\varepsilon_{\alpha,n})+C(i+\varepsilon_{\alpha,n})\right|^2} \tag{7}$$

For the PRCC scheme, the first path transmits the original MIMO-OFDM signals with phase rotation $\varphi_{\alpha,n}$, while the second path transmits the conjugate version of the original MIMO-OFDM signals with phase rotation $-\varphi_{\alpha,n}$, i.e., the $d_{\beta,l,n}e^{j\varphi_{\alpha,n}}$ signal is transmitted in the first path and the $\left\{d_{\beta,l,n}e^{-j\varphi_{\alpha,n}}\right\}^*$ is transmitted in the second path, respectively. Therefore, the time-domain received signals of the two paths can be expressed as follows,

$$\mathbf{y}^{(1)}_{\alpha,l,n}=\sum_{\beta=1}^{N_t}\sum_{p=0}^{L-1}d_{\beta,(1-p),n}h_{\beta,\alpha,p,n}e^{j\varphi_{\alpha,n}}e^{j\frac{2\pi}{N}l\varepsilon_{\alpha,n}}+w^{(1)}_{\alpha,l,n}\,,\qquad l=0,1,\ldots,N-1 \tag{8}$$

$$\mathbf{y}^{(2)}_{\alpha,l,n}=\sum_{\beta=1}^{N_t}\sum_{p=0}^{L-1}d^*_{\beta,(1-p),n}h_{\beta,\alpha,p,n}e^{j\frac{2\pi}{N}l\varepsilon_{\alpha,n}}+w^{(2)}_{\alpha,l,n}\,,\qquad l=0,1,\ldots,N-1 \tag{9}$$

And the frequency-domain received signals of the two paths can be written as follows,

$$\mathbf{Y}^{(1)}_{\alpha,m,n}=\sum_{\beta=1}^{N_t}D_{\beta,m,n}H_{\beta,m,n}e^{j\varphi_{\alpha,n}}C(-\varepsilon_{\alpha,n})+\sum_{\beta=1}^{N_t}\sum_{\substack{i=0\\i\neq m}}^{N-1}D_{\beta,m,n}H_{\beta,m,n}e^{j\varphi_{\alpha,n}}C(m-i-\varepsilon_{\alpha,n})+W^{(1)}_{\alpha,m,n}$$
$$m=0,\ldots,N-1 \tag{10}$$

$$\mathbf{Y}^{(2)}_{\alpha,m,n}=\sum_{\beta=1}^{N_t}D_{\beta,m,n}H_{\beta,m,n}e^{-j\varphi_{\alpha,n}}C(\varepsilon_{\alpha,n})+\sum_{\beta=1}^{N_t}\sum_{\substack{i=0\\i\neq m}}^{N-1}D_{\beta,m,n}H_{\beta,m,n}e^{-j\varphi_{\alpha,n}}C(m-i+\varepsilon_{\alpha,n})+W^{(2)}_{\alpha,m,n}\,,$$
$$m=0,\ldots,N-1 \tag{11}$$

By combining and averaging the signal processed in (10) and (11), we obtain

$$\mathbf{R}_{\alpha,m,n}=\frac{1}{2}\left(\mathbf{Y}^{(1)}_{\alpha,m,n}+\mathbf{Y}^{(2)}_{\alpha,m,n}\right), \tag{12}$$

From (12), the CIR of the PRCC scheme in 2×2 MIMO-OFDM can be defined as

$$CIR_{\alpha,PRCC} = \frac{\left|e^{j\varphi_{\alpha,n}}C(-\varepsilon_{\alpha,n}) + e^{-j\varphi_{\alpha,n}}C(\varepsilon_{\alpha,n})\right|^2}{\sum_{i=1}^{N-1}\left|e^{j\varphi_{\alpha,n}}C(i-\varepsilon_{\alpha,n}) + e^{-j\varphi_{\alpha,n}}C(i+\varepsilon_{\alpha,n})\right|^2} \tag{13}$$

We can optimize $\varphi_{\alpha,n}$ in terms of maximizing the CIR in (13),

$$\varphi_{\alpha,n}^{(opt)} = \arg\max CIR_{PRCC}\left(\varepsilon_{\alpha,n}, \varphi_{\alpha,n}\right) \tag{14}$$

The optimal phase rotation can be express as follows

$$\varphi_{\alpha,n}^{(opt)} = -\pi\varepsilon_{\alpha,n}\frac{N-1}{N} \tag{15}$$

3. The Proposed Receiver

The adaptive receiver of the proposed modified PRCC receiver for MIMO-OFDM is shown as Fig. 2 With the presence of frequency offset $\varepsilon_{\alpha,n}$, the time-domain received signals from the two transmission paths can be described as follows,

$$\mathbf{y}_{\alpha,l,n}^{(1)} = \sum_{\beta=1}^{N_t}\sum_{p=0}^{L-1} d_{\beta,(1-p),n} h_{\beta,\alpha,p,n} e^{j\frac{2\pi}{N}l\varepsilon_{\alpha,n}} + w_{\alpha,l,n}^{(1)}, \qquad l = 0,1,\ldots,N-1 \tag{16}$$

$$\mathbf{y}_{\alpha,l,n}^{(2)} = \sum_{\beta=1}^{N_t}\sum_{p=0}^{L-1} d_{\beta,(1-p),n}^{*} h_{\beta,\alpha,p,n} e^{j\frac{2\pi}{N}l(\varepsilon_{\alpha,n}+\Delta\varepsilon_{\alpha,n})} + w_{\alpha,l,n}^{(2)} \qquad l = 0,1,\ldots,N-1 \tag{17}$$

After performing FFT operation, we obtain the following frequency-domain signals as follows,

$$\mathbf{Y}_{\alpha,m,n}^{(1)} = \sum_{\beta=1}^{N_t} D_{\beta,m,n} H_{\beta,m,n} C(-\varepsilon_{\alpha,n}) + \sum_{\beta=1}^{N_t}\sum_{\substack{i=0\\ i\neq m}}^{N-1} D_{\beta,m,n} H_{\beta,\alpha,i,n} C(m-i-\varepsilon_{\alpha,n}) + W_{\alpha,m,n}^{(1)}, \tag{18}$$

$$\mathbf{Y}_{\alpha,m,n}^{(2)} = \sum_{\beta=1}^{N_t} D_{\beta,m,n} H_{\beta,\alpha,m,n} C(\varepsilon_{\alpha,n}+\Delta\varepsilon_{\alpha,n}) + \sum_{\beta=1}^{N_t}\sum_{\substack{i=0\\ i\neq m}}^{N-1} D_{\beta,m,n} H_{\beta,\alpha,i,n} C(m-i+\varepsilon_{\alpha,n}+\Delta\varepsilon_{\alpha,n}) + W_{\alpha,m,n}^{(2)}, \tag{19}$$

where $\mathbf{Y}_{\alpha,m,n}^{(i)}$, $H_{\beta,\alpha,m,n}$ and $w_{\alpha,m,n}^{(i)}$ correspond to the received symbol, channel response and noise at the β-th transmitting antenna, the α-th receiving antenna, the *m*-th subcarrier in the *i*-th path within the *n*-th transmission after FFT modulation, respectively. $\varepsilon_{\alpha,n}$ and $\varepsilon_{\alpha,n}+\Delta\varepsilon_{\alpha,n}$ denote the CFOs in the two paths at the α-th receiving antenna within the *n*-th transmission, respectively. After performing the individual phase rotations and the weighting coefficients, the signals in both paths can be written as follows,

$$\mathbf{R}_{\alpha,m,n}^{(1)} = e^{j\varphi_{\alpha,n}}\mathbf{H}_{\beta,\alpha,m,n}^{(1)*}\mathbf{Y}_{\alpha,m,n}^{(1)} \tag{20}$$

$$\mathbf{R}^{(2)}_{\alpha,m,n} = e^{j(\varphi_{\alpha,n}+\Delta\varphi_{\alpha,n})}\mathbf{H}^{(2)*}_{\beta,\alpha,m,n}\mathbf{Y}^{(2)}_{\alpha,m,n} \tag{21}$$

where $\varphi_{\alpha,n}$ and $\varphi_{\alpha,n}+\Delta\varphi_{\alpha,n}$ denote the designed phase rotations in the two paths at the α -th receiving antenna within the *n*-th transmission, respectively. Finally, the output signal can be expressed as follows,

$$\mathbf{R}_{\alpha,m,n} = \frac{1}{2}\left(\mathbf{Y}^{(1)}_{\alpha,m,n} + \mathbf{Y}^{(2)}_{\alpha,m,n}\right) \tag{22}$$

From (22), the CIR of the proposed adaptive modified PRCC scheme can be defined as follows,

$$\begin{aligned} &CIR_{\alpha,proposed}\left(\varepsilon_{\alpha,n},\Delta\varepsilon_{\alpha,n},\phi_{\alpha,n},\Delta\phi_{\alpha,n}\right)= \\ &\frac{\left|e^{j\varphi_{\alpha,n}}C(-\varepsilon_{\alpha,n})+e^{-j(\varphi_{\alpha,n}+\Delta\varphi_{\alpha,n})}C(-\varepsilon_{\alpha,n}+\Delta\varepsilon_{\alpha,n})\right|^2}{\sum_{l=1}^{N-1}\left|e^{j\varphi_{\alpha,n}}C(-\varepsilon_{\alpha,n})+e^{-j(\varphi_{\alpha,n}+\Delta\varphi_{\alpha,n})}C(-\varepsilon_{\alpha,n}+\Delta\varepsilon_{\alpha,n})\right|^2} \end{aligned} \tag{23}$$

Given $\varepsilon_{\alpha,n}$ and $\Delta\varepsilon_{\alpha,n}$, the optimal phase rotations can be determined by maximizing the CIR,

$$\left(\varphi^{(opt)}_{\alpha,n},\Delta\varphi^{(opt)}_{\alpha,n}\right)=\arg\max CIR_{\alpha,PRCC}\left(\varepsilon_{\alpha,n},\Delta\varepsilon_{\alpha,n},\phi_{\alpha,n},\Delta\phi_{\alpha,n}\right) \tag{24}$$

By solving (24), we can derive the following sufficient condition for the optimal phase rotations,

$$2\varphi^{(opt)}_{\alpha,n}+\Delta\varphi^{(opt)}_{\alpha,n} = -\pi\frac{\left(2\varepsilon_{\alpha,n}+\Delta\varepsilon_{\alpha,n}\right)(N-1)}{N} \tag{25}$$

So the corresponding solution is as follows,

$$\varphi^{(opt)}_{\alpha,n} = -\pi\frac{\varepsilon_{\alpha,n}(N-1)}{N} \tag{26}$$

$$\Delta\varphi^{(opt)}_{\alpha,n} = -\pi\frac{\Delta\varepsilon_{\alpha,n}(N-1)}{N} \tag{27}$$

Assuming that the phase rotations $\varphi_{\alpha,n}$ and $\Delta\varphi_{\alpha,n}$ have approached the optimal values after sufficient adaptations, the CFOs can be estimated as follows,

$$\hat{\varepsilon}_{\alpha,n} = -\varphi_{\alpha,n}\frac{N}{\pi(N-1)} \tag{28}$$

$$\Delta\hat{\varepsilon}_{\alpha,n} = -\Delta\varphi_{\alpha,n}\frac{N}{\pi(N-1)} \tag{29}$$

The equalization output based on the MMSE criteria can be expressed as follows,

$$\hat{\mathbf{D}}_{\alpha,m,n} = \left(\left(\mathbf{H}_{\beta,\alpha,m,n}\mathbf{H}^{H}_{\beta,\alpha,m,n} + \mathbf{I}\sigma^2\right)^{-1}\mathbf{H}^{H}_{\beta,\alpha,m,n}\right)\mathbf{R}_{\alpha,m,n} \quad (30)$$

Then, the desired signal can be written as

$$\mathbf{dr}_{\alpha,m,n} = e^{j\varphi_{\alpha,n}}\hat{\mathbf{D}}_{\alpha,m,n}e^{j2\pi(-\hat{\varepsilon}_{\alpha,n})n/N} \quad (31)$$

Therefore, we can define the error signal as follows,

$$\mathbf{e}_{\alpha,m,n} = \mathbf{dr}_{\alpha,m,n} - \mathbf{R}_{\alpha,m,n} \quad (32)$$

If we average the squared error signals of all the subcarriers, the cost function used in the receiver can be defined as follows,

$$J_{\alpha,n}\left(\phi_{\alpha,n}, \Delta\phi_{\alpha,n}, \hat{\varepsilon}_{\alpha,n}, \Delta\hat{\varepsilon}_{\alpha,n}\right) = E\left\{\left|\mathbf{e}_{\alpha,m,n}\right|^2\right\} = \frac{1}{N}\sum_{m=0}^{N-1}\left|\mathbf{e}_{\alpha,m,n}\right|^2 \quad (33)$$

The phase rotation update equation employing normalized block least mean squared (BLMS) algorithm can be described as follows,

$$\begin{bmatrix}\phi_{\alpha,n+1}\\ \Delta\phi_{\alpha,n+1}\end{bmatrix} = \begin{bmatrix}\phi_{\alpha,n}\\ \Delta\phi_{\alpha,n}\end{bmatrix} + \begin{bmatrix}\mu_1 & 0\\ 0 & \mu_2\end{bmatrix}\frac{\nabla J\left(\phi_{\alpha,n}, \Delta\phi_{\alpha,n}, \hat{\varepsilon}_{\alpha,n}, \Delta\hat{\varepsilon}_{\alpha,n}\right)}{\zeta + E\left\{\mathbf{Y}^{(1)H}_{\alpha,n}\mathbf{Y}^{(1)}_{\alpha,n} + \mathbf{Y}^{(2)H}_{\alpha,n}\mathbf{Y}^{(2)}_{\alpha,n}\right\}} \quad (34)$$

where $\mathbf{Y}^{(i)}_{\alpha,n} = \left[Y^{(i)}_{\alpha,1,n} \quad Y^{(i)}_{\alpha,2,n} \quad \ldots \quad Y^{(i)}_{\alpha,N,n}\right]^T$, μ_i is the step size employed in the i-th paths, ζ is a small positive constant.

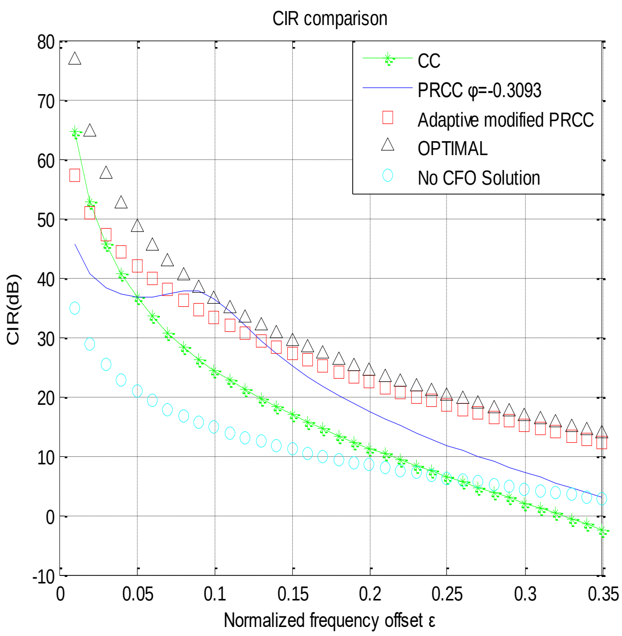

Fig. 3. CIR comparison of existing methods and the proposed adaptive modified PRCC.

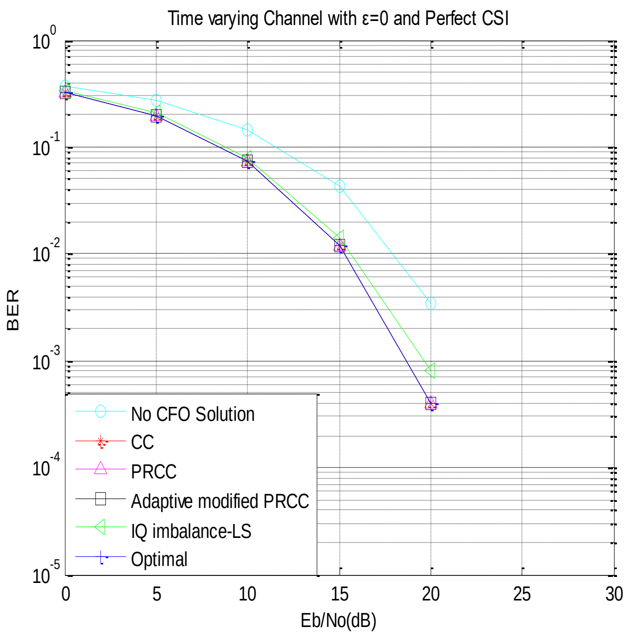

Fig. 4. BER comparison of existing methods and the proposed adaptive

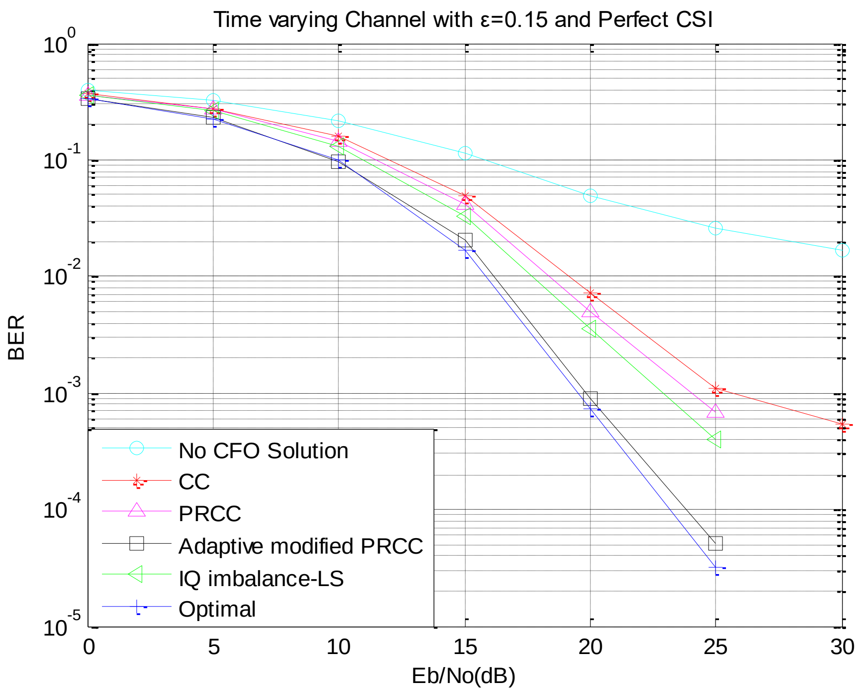

Fig. 5. BER comparison of existing methods and the proposed adaptive modified PRCC with QPSK modulation, ε=0.15 under ITU Channel A of 100km/hr.

4. Simulation Results

In this section, we compare the performances of CC, PRCC, Two stage IQ-Imbalance schemes, optimal solution and the proposed adaptive modified PRCC schemes that is jointly designed with MMSE equalization and Kalmna channel estimation for MIMO-OFDM systems. The adopted channel models are multipath Rayleigh fading and ITU channel models defined in the IEEE 802.11 Working Group [27]. We simulate those schemes in the multipath Rayleigh fading channel model and ITU channel model for vehicular test environments with 6 path taps. In these systems, we use QPSK modulation and 16QAM modulation for MIMO-OFDM systems, and then we apply two antennas at transmitter, and two antennas at receiver. The simulated MIMO-OFDM systems assume there are 256 subcarriers (i.e., FFT size=256), 20 cyclic prefixes (i.e., CP=20), difference of speed(i.e., V=5, 60, 100, 200, 300 km/hr), and different frequency offset (i.e., $\varepsilon = 0, 0.15, 0.25$). Fig. 3 shows the CIR comparisons of no CFO solution, CC, PRCC, the proposed adaptive modified PRCC and the optimal schemes for MIMO-OFDM system. From the result, we can see that the CIR performance of the proposed adaptive modified PRCC scheme is close to that of the optimal solution. Additionally, the CC scheme cannot provide good enough CIR when CFOs are large, and the PRCC cannot provide the widest tolerable range of CFOs with phase rotations equal to -0.3093, which is optimal only for $\varepsilon = 0.1$. Fig. 4 shows the BER comparison of the no CFO solution, CC, PRCC, IQ Imbalance, the optimal and the proposed adaptive modified PRCC schemes with QPSK modulation, frequency offset ε=0 in ITU Channel A with speed 100km/hr is applied. It demonstrates that the performances of the proposed adaptive modified PRCC, PRCC, and CC schemes are all approaches to that of the optimal scheme. This is because the proposed adaptive modified PRCC and PRCC schemes are designed for ICI self-cancellation by taking advantage of conjugate cancellation scheme which is a good method to mitigate the problem of frequency offset in small CFO. The IQ imbalance scheme is poorer than those of other schemes in such situation. Fig. 5 shows the BER comparison of the no CFO Solution, CC, PRCC, IQ Imbalance, the optimal scheme and the proposed adaptive modified PRCC schemes with QPSK modulation and ε=0.15 under ITU

Channel A of 100km/hr**.** It demonstrates that the performance of the proposed adaptive modified PRCC scheme is better than those of CC, PRCC and IQ imbalance. This is because that we reduce the developed speed of convergence by a factor of *N* and increase the computational advantage of the normalized BLMS algorithm, so the proposed adaptive receiver can track the CFOs and update the phase rotations based on the estimated CFOs from two the paths in fast fading channel. Fig. 6 shows the BER comparison of the no CFO solution, CC, PRCC, IQ Imbalance, the optimal scheme and the proposed adaptive modified PRCC scheme with QPSK modulation, ε=0.25 in ITU Channel A with speed 100km/hr is applied. It demonstrates that the performance of the proposed adaptive modified PRCC scheme is better than those of CC, PRCC and IQ imbalance. This is because the developed adaptive receiver can track CFOs and update the phase rotations based on the estimated CFOs from the two paths in fast fading channel. Fig. 7 shows the BER comparison of the no CFO solution, CC, PRCC, IQ Imbalance, the optimal scheme and the proposed adaptive modified PRCC schemes with 16QAM modulation, ε=0.15 in ITU Channel A with speed 100km/hr is applied. It demonstrates that the performance of the proposed adaptive modified PRCC scheme is better than those of CC, PRCC and IQ imbalance. This is because the proposed adaptive receiver can track CFOs and update the phase rotations based on the estimated CFOs from the two paths in fast fading channel. Additionally, we can see that the BER of the adaptive modified PRCC scheme with 16QAM modulation or high order modulation are close to those of other schemes. Fig. 8 shows the BER comparison of the no CFO solution, CC, PRCC, IQ Imbalance, the optimal scheme and the proposed adaptive modified PRCC schemes with 16QAM modulation, ε=0.25 in ITU Channel A with speed 100km/hr is applied. It demonstrates that the performance of the proposed adaptive modified PRCC scheme is better than those of CC, PRCC and IQ imbalance.

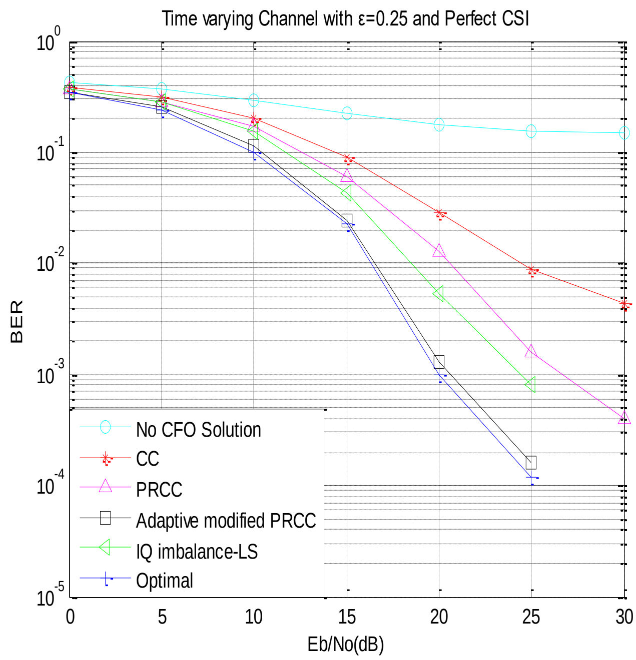

Fig. 6. BER comparison of existing methods and the proposed adaptive modified PRCC with QPSK modulation, ε=0.25 under ITU Channel A of 100km/hr.

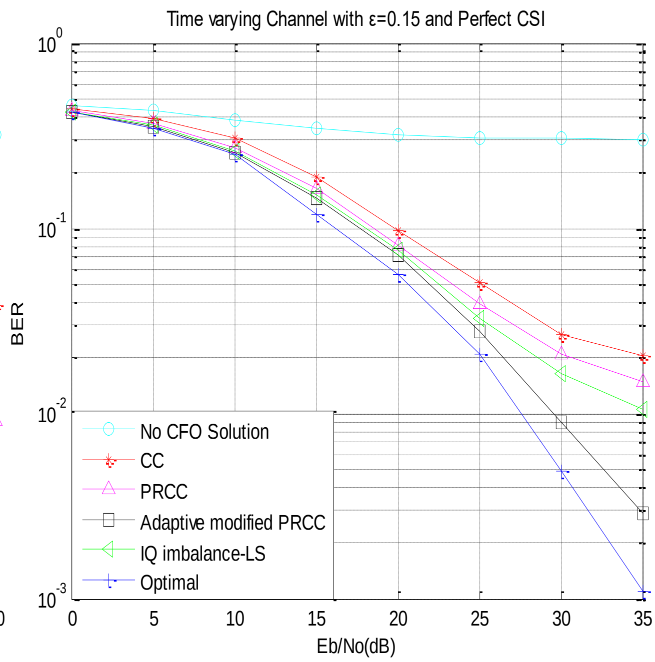

Fig. 7. BER comparison of existing methods and the proposed adaptive modified PRCC with 16QAM modulation, ε=0.15 under ITU Channel A of 100km/hr.

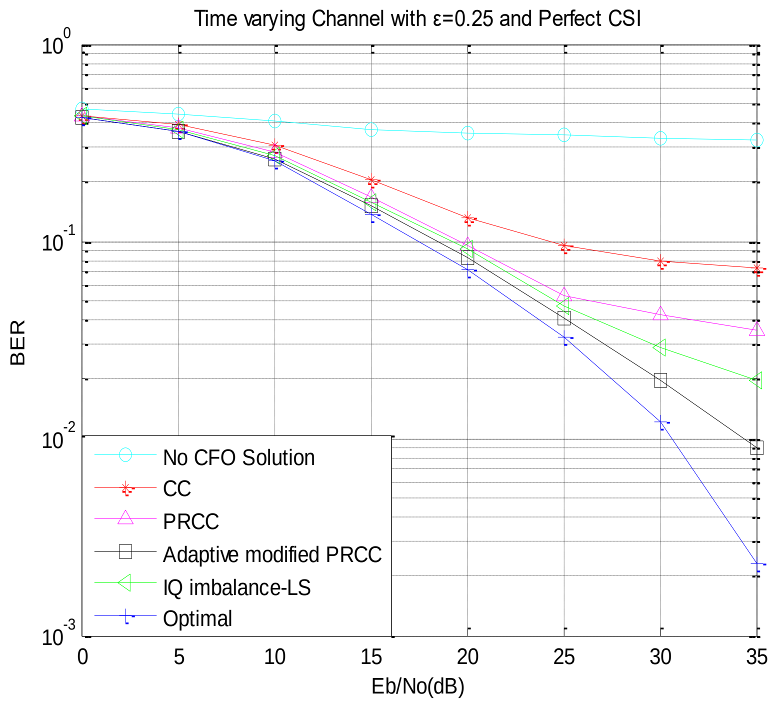

Fig. 8. BER comparison of existing methods and the proposed adaptive modified PRCC with 16QAM modulation, ε=0.25 under ITU Channel A of 100km/hr.

5. Conclusion

This paper proposes an advanced receiver with ICI self cancellation, Kalman channel estimation and equalization for MIMO-OFDM systems in time-varying channels. ICI self cancellation is an efficient technique in terms of bit error rate and computational complexity. We employ an adaptive receiver and construct two-path conjugate transmission that is based on the PRCC concept. Additionally, the Kalman filter is used to trace the time-varying channels of the MIMO-OFDM systems. We also derive the optimal phase rotations using the criterion of maximizing the carrier-to-interference ratio, and then develop an adaptive normalized block least mean-squared algorithm to approach the optimal solutions. With such adaptive phase rotations in the receiver, the CFO variations due to the channel effect and the mismatch between oscillators at the transmitter and receiver can effectively be tracked without feeding back the CFO estimate to the transmitter as required in conventional PRCC scheme.

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