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Decoupled estimation of frequency-dependent IQI and channel for OFDM systems with direct-conversion transceivers

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Abstract: The in-phase and quadrature-phase imbalance (IQI) is one of the major radio frequency impairments existing in orthogonal frequency division multiplexing (OFDM) systems with direct-conversion transceivers. During the transmission of the communication signal, the impact of IQI is coupled with channel impulse responses (CIR), which makes the traditional channel estimation schemes ineffective. A decoupled estimation scheme is proposed to separately estimate the frequency-dependent IQI and wireless channel. Firstly, the generalized channel model is built to separate the parameters of IQI and wireless channel. Then an iterative estimation scheme of frequency-dependent IQI is designed at the initial stage of communication. Finally, based on the estimation result of IQI, the least square algorithm is utilized to estimate the channel-related parameters at each time of channel variation. Compared with the joint estimation schemes of IQI and channel, the proposed decoupled estimation scheme requires much lower training overhead at each time of channel variation. Simulation results demonstrate the good estimation performance of the proposed scheme.

Keywords: direct-conversion transceivers, frequency-dependent in-phase and quadrature-phase imbalance (IQI), orthogonal frequency division multiplexing (OFDM), decoupled estimation.

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1. Introduction

Compared to the classic super-heterodyne structure, the structure of direct conversion radio has advantages of low cost, small size, low power consumption, and flexibility to various frequency bands and wireless standards [1,2]. Thus the direct-conversion transmitters and receivers (transceivers) have attracted great attention in recent years and are assumed as a good candidate for massive multiple input multiple output (MIMO) systems [3], which is one of the key technologies in 5G.

However, due to the structure limitation and non-ideal properties of analog components, the direct-conversion transceivers are prone to suffer from in-phase and quadrature-phase imbalance (IQI) [4]. This imbalance is the mismatch between the in-phase and quadrature-phase branches in amplitude, phase shift, and cascaded filters, which can be present at both the transmitter and the receiver. In orthogonal frequency division multiplexing (OFDM) systems, IQI would destroy the orthogonality of subcarriers and lead to an overall degradation in system performance. The problem of IQI has received significant research attention. The effect of IQI on OFDM receivers was studied and system-level algorithms to compensate for the distortions were proposed in [4]. In [5], a recursive compensation technique was proposed to compensate for both frequency-independent and frequency-dependent IQIs in direct conversion transmitters. In [6], a widely linear digital self-interference cancellation processing was proposed in direct-conversion full-duplex transceivers. In [7], the effect of IQI on MIMO beamforming systems was investigated and the linear throughput-maximizing digital baseband joint beamforming and equalization scheme was proposed. A power-based compensation method was proposed for wideband frequency-dependent IQI in direct-conversion transmitters in [8].

Effective compensation of IQI requires accurate estimates of channel impulse responses (CIR) and IQI parameters. During the transmission of signal, the impact of IQI is coupled with CIR, which makes the traditional channel estimation schemes inappropriate. Considering the coupled status of IQI and CIR, several joint estimation schemes of IQI and channel were investigated in [9 – 13]. A least square (LS) algorithm for IQI and channel estimation was proposed in [9] with the aid of nearly twenty OFDM training symbols. A Gaussian elimination (GE) scheme was proposed in [10] for joint estimation of frequency-

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dependent IQI and channel, which fully exploited the time-domain property of channel and required two training symbols per OFDM frame. In [11], a joint IQI and channel estimation method in the time-domain was developed for OFDM systems operating over doubly selective channels. An expectation-maximization (EM) based iterative algorithm to jointly estimate the CIR and IQI was proposed for Alamouti OFDM systems in [12]. In [13], an iterative shrinkage algorithm was adopted to estimate the IQI parameters and sparse channel for OFDM systems.

In the aforementioned joint estimation schemes, both the parameters of IQI and wireless channel are estimated together. However, the joint estimation of IQI and wireless channel is not an economical solution when the wireless channel is time-varying. The variable characteristic of channel would lead to increasing number of channel parameters to be estimated over time. However, the change rate of IQI parameters is extremely slow and one estimation of IQI could be effective for a long communication period. Thus a decoupled scheme to separately estimate IQI and channel parameters is desirable, which could effectively reduce the training overhead and redundancy of estimation.

A decoupled estimation scheme was proposed in [14] to separately estimate the frequency-independent IQI and channel. However, the frequency-dependent IQI, which usually exists in wideband direct-conversion transceivers, had not been discussed in [14]. The parameters of frequency-dependent IQI vary at each OFDM subcarrier. Thus the total number of unknown frequency-dependent IQI parameters is N times of frequency-independent IQI, with N denoting the number of OFDM subcarriers. It makes the decoupled estimation of frequency-dependent IQI and channel a challenging task.

In this paper, a decoupled IQI and channel estimation scheme is proposed for OFDM systems with frequency-dependent IQIs at both transmitter and receiver. This paper is distinct from the related literature in the following aspects: (i) Different from the previous joint estimation of IQI and channel [9 – 13], this paper considers decoupled estimation of IQI and channel to reduce the training overhead at each time of channel variation. (ii) Different from [14] which discussed the decoupled estimation of frequency-independent IQI and channel, this paper proposes a decoupled estimation scheme for frequency-dependent IQI and channel, which resolves a more common existing but challenging problem that could not be solved by [14]. (iii) This paper also devises a new training pattern for the Gaussian elimination scheme in [10].

This paper is organized as follows. A generalized channel model is built in Section 2, upon which the parameters of frequency-dependent IQI and wireless channel could be disjunctively processed. Then a decoupled estimation scheme is proposed in Section 3. The frequency-dependent IQI parameters are estimated iteratively with specially designed training groups at the initial stage of communication. Next, the estimates of channel-related parameters are acquired with just one OFDM training symbol at each time of channel change, which successfully reduces the training overhead. In Section 4, simulations verify the effectiveness of the proposed scheme. Finally, the paper is concluded in Section 5.

2. System model

An OFDM system with frequency-dependent IQI at both transmitter and receiver is considered in following. The transmitted OFDM symbol is denoted by

$$\boldsymbol{S} = [\boldsymbol{S}(1) \quad \boldsymbol{S}(2) \quad \cdots \quad \boldsymbol{S}(N)]^{\mathrm{T}} \tag{1}$$

where N is the number of OFDM subcarriers. Each transmitted OFDM symbol is passed through the inverse discrete Fourier transform (IDFT) operation as

$$\boldsymbol{s} = \boldsymbol{F}^{\mathrm{H}}\boldsymbol{S} \tag{2}$$

where $\boldsymbol{s}$ is the signal vector in time-domain, $\boldsymbol{F}$ is the normalized DFT matrix as

$$\boldsymbol{F}(m,n) = \frac{1}{\sqrt{N}}\exp\left(\frac{-\mathrm{j}2\pi(m-1)(n-1)}{N}\right),$$
$$1 \leqslant m,n \leqslant N.$$

Then a cyclic prefix (CP) is added to each transformed block of data. Due to the impact of IQI at the transmitter, the distorted transmitted signal is given by

$$\boldsymbol{s}_d = \boldsymbol{\lambda}_t\boldsymbol{s} + \boldsymbol{\varphi}_t\boldsymbol{s}^* \tag{3}$$

where $\boldsymbol{\lambda}_t$ and $\boldsymbol{\varphi}_t$ are frequency-dependent IQI parameters at the transmitter caused by the mismatches between I/Q branches in amplitude (α_t), phase shift (θ_t), and imbalance filters ($\boldsymbol{\xi}_t^I$ and $\boldsymbol{\xi}_t^Q$):

$$\boldsymbol{\lambda}_t = ((\mu_t + \nu_t)\boldsymbol{\xi}_t^I + (\mu_t - \nu_t)\boldsymbol{\xi}_t^Q)/2$$
$$\boldsymbol{\varphi}_t = ((\mu_t + \nu_t)\boldsymbol{\xi}_t^I - (\mu_t - \nu_t)\boldsymbol{\xi}_t^Q)/2$$
$$\mu_t = \cos(\theta_t/2) - \mathrm{j}\alpha_t\sin(\theta_t/2)$$
$$\nu_t = \alpha_t\cos(\theta_t/2) - \mathrm{j}\sin(\theta_t/2).$$

Then $\boldsymbol{s}_d$ is transmitted through the wireless channel. After discarding the received CPs, the received baseband signal is given by

$$\boldsymbol{r} = \boldsymbol{h}\boldsymbol{s}_d + \boldsymbol{n} \tag{4}$$

where $\boldsymbol{h}$ is the channel impulse response (CIR) and $\boldsymbol{n}$ is the additive white Gaussian noise (AWGN). After experiencing receiver side IQI, the received signal becomes

$$\boldsymbol{r}_d = \boldsymbol{\lambda}_r\boldsymbol{r} + \boldsymbol{\varphi}_r\boldsymbol{r}^* \tag{5}$$

where $\boldsymbol{\lambda}_r$ and $\boldsymbol{\varphi}_r$ are frequency-dependent IQI parameters at the receiver:

$$\boldsymbol{\lambda}_r = ((\mu_r + \nu_r^*)\boldsymbol{\xi}_r^I + (\mu_r - \nu_r^*)\boldsymbol{\xi}_r^Q)/2$$
$$\boldsymbol{\varphi}_r = ((\mu_r^* + \nu_r)\boldsymbol{\xi}_r^I - (\mu_r^* - \nu_r)\boldsymbol{\xi}_r^Q)/2$$
$$\mu_r = \cos(\theta_r/2) + \mathrm{j}\alpha_r \sin(\theta_r/2)$$
$$\nu_r = \alpha_r \cos(\theta_r/2) - \mathrm{j}\sin(\theta_r/2).$$

By substituting (3) and (4) into (5) and making the DFT operation, the relationship between the transmitted and received OFDM symbol at the kth subcarrier can be described as

$$\boldsymbol{R}(k) = \boldsymbol{H}_a(k)\boldsymbol{S}(k) + \boldsymbol{H}_b(k)\boldsymbol{S}^{\#}(k) + \boldsymbol{W}(k) \tag{6}$$

where $\boldsymbol{S}^{\#}$ denotes the mirror interference of $\boldsymbol{S}$, which is caused by IQI with the following structure:

$$\begin{aligned}\boldsymbol{S}^{\#} = [&\boldsymbol{S}^*(1) \quad \boldsymbol{S}^*(N) \quad \cdots \quad \boldsymbol{S}^*(N/2+2)\\ &\boldsymbol{S}^*(N/2+1) \quad \boldsymbol{S}^*(N/2) \quad \cdots \quad \boldsymbol{S}^*(2)]^{\mathrm{T}}\end{aligned} \tag{7}$$

$\boldsymbol{H}_a$ and $\boldsymbol{H}_b$ are generalized channels composed of the coupled IQI and channel parameters as

$$\boldsymbol{H}_a(k) = \boldsymbol{\Lambda}_r(k)\boldsymbol{H}(k)\boldsymbol{\Lambda}_t(k) + \boldsymbol{\Phi}_r(k)\boldsymbol{H}^{\#}(k)\boldsymbol{\Phi}_t^{\#}(k) \tag{8}$$

$$\boldsymbol{H}_b(k) = \boldsymbol{\Lambda}_r(k)\boldsymbol{H}(k)\boldsymbol{\Phi}_t(k) + \boldsymbol{\Phi}_r(k)\boldsymbol{H}^{\#}(k)\boldsymbol{\Lambda}_t^{\#}(k) \tag{9}$$

where $\boldsymbol{\Lambda} = \boldsymbol{F}\boldsymbol{\lambda}$ and $\boldsymbol{\Phi} = \boldsymbol{F}\boldsymbol{\varphi}$, which are the frequency-domain frequency-dependent IQI parameters, with the subscripts r and t representing the receiver and transmitter, respectively. $\boldsymbol{H} = \boldsymbol{F}\boldsymbol{h}$, which is the N-dimensional channel vector in the frequency domain. $\boldsymbol{W}$ is the frequency-domain received noise which is distorted by the receiver IQI as

$$\boldsymbol{W}(k) = \boldsymbol{\Lambda}_r(k)\boldsymbol{N}(k) + \boldsymbol{\Phi}_r(k)\boldsymbol{N}^{\#}(k) \tag{10}$$

where $\boldsymbol{N}$ is an additive white Gaussian noise (AWGN) vector in the frequency-domain.

It can be found in (6) that at the receiver, the transmitted signal $\boldsymbol{S}$ is interfered by its mirror interference $\boldsymbol{S}^{\#}$. Considering the interferences within a pair of mirror subcarriers, the system model can be re-written in a series of independent 2×2 equations as

$$\begin{aligned}\begin{bmatrix}\boldsymbol{R}(k)\\ \boldsymbol{R}^{\#}(k)\end{bmatrix} = \begin{bmatrix}\boldsymbol{H}_a(k) & \boldsymbol{H}_b(k)\\ \boldsymbol{H}_b^{\#}(k) & \boldsymbol{H}_a^{\#}(k)\end{bmatrix} \times\\ \begin{bmatrix}\boldsymbol{S}(k)\\ \boldsymbol{S}^{\#}(k)\end{bmatrix} + \begin{bmatrix}\boldsymbol{W}(k)\\ \boldsymbol{W}^{\#}(k)\end{bmatrix}.\end{aligned} \tag{11}$$

In order to compensate the IQI in (11), information of $\boldsymbol{H}_a$ and $\boldsymbol{H}_b$ need to be known. If a joint estimation scheme is used, the number of estimation parameters is doubled compared to a conventional channel estimation problem in OFDM systems with ideal IQ branches. These doubled estimation parameters lead to double training overhead and computational amount. In order to avoid this problem, we reorganize the generalized channel models in (8) and (9) as

$$\boldsymbol{H}_a(k) = \boldsymbol{C}(k) + \boldsymbol{A}_t^{\#}(k)\boldsymbol{A}_r(k)\boldsymbol{C}^{\#}(k) \tag{12}$$

$$\boldsymbol{H}_b(k) = \boldsymbol{A}_t(k)\boldsymbol{C}(k) + \boldsymbol{A}_r(k)\boldsymbol{C}^{\#}(k) \tag{13}$$

where the wireless channel related information is only concentrated in $\boldsymbol{C}$ as

$$\boldsymbol{C}(k) = \boldsymbol{\Lambda}_t(k)\boldsymbol{\Lambda}_r(k)\boldsymbol{H}(k) \tag{14}$$

while $\boldsymbol{A}_t$ and $\boldsymbol{A}_r$ are just constructed by IQI parameters as

$$\boldsymbol{A}_t(k) = \boldsymbol{\Phi}_t(k)/\boldsymbol{\Lambda}_t(k) \tag{15}$$

$$\boldsymbol{A}_r(k) = \boldsymbol{\Phi}_r(k)/\boldsymbol{\Lambda}_r^{\#}(k). \tag{16}$$

Then the system model in (11) is rewritten to decouple the channel-related information $\boldsymbol{C}$ from the IQI parameters of $\boldsymbol{A}_t$ and $\boldsymbol{A}_r$ as

$$\underbrace{\begin{bmatrix}\boldsymbol{R}(k)\\ \boldsymbol{R}^{\#}(k)\end{bmatrix}}_{\boldsymbol{R}_k} = \underbrace{\begin{bmatrix}\mathbf{P}_{k,1} & \mathbf{P}_{k,2}\\ \mathbf{P}_{k,3} & \mathbf{P}_{k,4}\end{bmatrix}}_{\mathbf{P}_k}\underbrace{\begin{bmatrix}\boldsymbol{S}(k)\\ \boldsymbol{S}^{\#}(k)\end{bmatrix}}_{\boldsymbol{S}_k} + \underbrace{\begin{bmatrix}\boldsymbol{W}(k)\\ \boldsymbol{W}^{\#}(k)\end{bmatrix}}_{\boldsymbol{W}_k} \tag{17}$$

where

$$\boldsymbol{P}_{k,1} = \boldsymbol{C}(k) + \boldsymbol{A}_t^{\#}(k)\boldsymbol{A}_r(k)\boldsymbol{C}^{\#}(k)$$
$$\boldsymbol{P}_{k,2} = \boldsymbol{A}_t(k)\boldsymbol{C}(k) + \boldsymbol{A}_r(k)\boldsymbol{C}^{\#}(k)$$
$$\boldsymbol{P}_{k,3} = \boldsymbol{A}_t^{\#}(k)\boldsymbol{C}^{\#}(k) + \boldsymbol{A}_r^{\#}(k)\boldsymbol{C}(k)$$
$$\boldsymbol{P}_{k,4} = \boldsymbol{C}^{\#}(k) + \boldsymbol{A}_t(k)\boldsymbol{A}_r^{\#}(k)\boldsymbol{C}(k).$$

Block fading of channel is assumed in the following discussion, which means that both $\boldsymbol{H}$ and $\boldsymbol{C}$ are variable from frame to frame but remain static within each frame. While considering the extremely slow variation speed of IQI, the parameters of $\boldsymbol{A}_t$ and $\boldsymbol{A}_r$ are assumed to be constant during the whole communication period. If the prior knowledge of $\boldsymbol{A}_t$ and $\boldsymbol{A}_r$ is known, only $\boldsymbol{C}$ is unknown in (17) and requires estimation. By this means, the number of estimation parameters could be reduced by half compared to that of (11). This is the motivation for us to investigate the decoupled estimation scheme of frequency-dependent IQI and channel in the next section.

3. Decoupled estimation scheme of frequency-dependent IQI and channel

In this section, a decoupled estimation scheme is proposed for frequency-dependent IQI and channel based on the system model in (17). Primarily, an iterative method to estimate the IQI parameters of $\boldsymbol{A}_t$ and $\boldsymbol{A}_r$ is developed.

Next, the channel-related parameter $\boldsymbol{C}$ is estimated in each OFDM frame with the aid of one training symbol, which reduces the redundancy of estimation and training overhead.

3.1 Iterative estimation of IQI parameters

In order to obtain the information of IQI at the initial stage of communication, an estimation scheme to iteratively estimate $\boldsymbol{A}_t$ and $\boldsymbol{A}_r$ is proposed as follows.

Firstly, we devise the frequency-domain training group of two contiguous OFDM symbols as

$$\boldsymbol{S}_{\mathrm{I}} = \boldsymbol{S}_{Tr} \tag{18}$$

$$\boldsymbol{S}_{\mathrm{II}} = \mathrm{j}\boldsymbol{S}_{Tr} \tag{19}$$

where $\boldsymbol{S}_{Tr}$ is an arbitrary N-dimensional OFDM training symbol, the subscripts I and II denote the first and second contiguous OFDM training symbols respectively.

With the aid of OFDM training groups, we could obtain the joint estimation results of $\boldsymbol{H}_a$ and $\boldsymbol{H}_b$ at two independent channel realizations respectively. Assuming that the wireless channel is constant within each training group, $\boldsymbol{H}_a$ and $\boldsymbol{H}_b$ are estimated with the GE approach [10] as

$$\boldsymbol{H}_a(k) = \frac{\boldsymbol{R}_{\mathrm{I}}(k) - \mathrm{j}\boldsymbol{R}_{\mathrm{II}}(k)}{2\boldsymbol{S}_{Tr}(k)} \tag{20}$$

$$\boldsymbol{H}_b(k) = \frac{\boldsymbol{R}_{\mathrm{I}}(k) + \mathrm{j}\boldsymbol{R}_{\mathrm{II}}(k)}{2\boldsymbol{S}^{\#}_{Tr}(k)} \tag{21}$$

where $\boldsymbol{R}_{\mathrm{I}}$ and $\boldsymbol{R}_{\mathrm{II}}$ are the received training symbols corresponding to $\boldsymbol{S}_{\mathrm{I}}$ and $\boldsymbol{S}_{\mathrm{II}}$, respectively.

Then (13) can be reformulated as

$$\underbrace{\begin{bmatrix} \boldsymbol{H}_{b1}(k) \\ \boldsymbol{H}_{b2}(k) \end{bmatrix}}_{\boldsymbol{H}_{bk}} = \underbrace{\begin{bmatrix} \boldsymbol{C}_1(k) & \boldsymbol{C}^{\#}_1(k) \\ \boldsymbol{C}_2(k) & \boldsymbol{C}^{\#}_2(k) \end{bmatrix}}_{\widetilde{\boldsymbol{C}}_k} \times \underbrace{\begin{bmatrix} \boldsymbol{A}_t(k) \\ \boldsymbol{A}_r(k) \end{bmatrix}}_{\boldsymbol{A}_k} \tag{22}$$

where the subscripts 1 and 2 denote two different channel realizations respectively. If the channel-related parameter $\boldsymbol{C}$ is known, $\boldsymbol{A}_t$ and $\boldsymbol{A}_r$ can be estimated by

$$\widehat{\boldsymbol{A}}_k = (\delta\boldsymbol{I} + \widetilde{\boldsymbol{C}}^{\mathrm{H}}_k\widetilde{\boldsymbol{C}}_k)^{-1}\widetilde{\boldsymbol{C}}^{\mathrm{H}}_k\boldsymbol{H}_{bk} \tag{23}$$

where the parameter $\delta > 0$ is added for regularization.

To start the iterative estimation procedure, we make an approximation that $\boldsymbol{A}^{\#}_t(k)\boldsymbol{A}_r(k) = 0$, based on the fact that $|\boldsymbol{\Phi}_t(k)| \ll |\boldsymbol{A}_t(k)|$ and $|\boldsymbol{\Phi}_r(k)| \ll |\boldsymbol{A}^{\#}_r(k)|$. Hence $\widetilde{\boldsymbol{C}}_k$ can be initialized as $\boldsymbol{C}_1(k) = \boldsymbol{H}_{a1}(k)$ and $\boldsymbol{C}_2(k) = \boldsymbol{H}_{a2}(k)$ by substituting $\boldsymbol{A}^{\#}_t(k)\boldsymbol{A}_r(k) = 0$ into (12). Then the first estimation of $\boldsymbol{A}_t$ and $\boldsymbol{A}_r$ can be obtained with (23).

Given the IQI parameters of $\boldsymbol{A}_t$ and $\boldsymbol{A}_r$, the model to estimate $\boldsymbol{C}(k)$ can be established as

$$\underbrace{\begin{bmatrix} \boldsymbol{H}_a(k) \\ \boldsymbol{H}^{\#}_a(k) \end{bmatrix}}_{\boldsymbol{H}_{ak}} = \underbrace{\begin{bmatrix} 1 & \boldsymbol{A}^{\#}_t(k)\boldsymbol{A}_r(k) \\ \boldsymbol{A}_t(k)\boldsymbol{A}^{\#}_r(k) & 1 \end{bmatrix}}_{\widetilde{\boldsymbol{A}}_k} \times \underbrace{\begin{bmatrix} \boldsymbol{C}(k) \\ \boldsymbol{C}^{\#}(k) \end{bmatrix}}_{\boldsymbol{C}_k}. \tag{24}$$

The estimation result is given by

$$\widehat{\boldsymbol{C}}_k = (\delta\boldsymbol{I} + \widetilde{\boldsymbol{A}}^{\mathrm{H}}_k\widetilde{\boldsymbol{A}}_k)^{-1}\widetilde{\boldsymbol{A}}^{\mathrm{H}}_k\boldsymbol{H}_{ak}. \tag{25}$$

Then we can use (23) and (25) to iteratively update the estimates of $\boldsymbol{A}_t$, $\boldsymbol{A}_r$ and $\boldsymbol{C}_1, \boldsymbol{C}_2$ until a predefined number of iterations I_t is performed. As the estimation of $\boldsymbol{A}_t$ and $\boldsymbol{A}_r$ is performed at the initial communication phase for just one time, the training overhead at this stage of four OFDM training symbols is negligible from the systematic point of view.

3.2 Estimation of channel-related parameter

With the knowledge of $\boldsymbol{A}_t$ and $\boldsymbol{A}_r$ as the estimation result of our first step, only $\boldsymbol{C}$ is unknown at each time of channel variation. Hence the training overhead would be reduced by half compared to the joint estimation schemes. In order to estimate the varying channel parameter $\boldsymbol{C}$ in the whole communication period, the system model of (17) is reformulated as

$$\underbrace{\begin{bmatrix} \boldsymbol{R}(k) \\ \boldsymbol{R}^{\#}(k) \end{bmatrix}}_{\boldsymbol{R}_k} = \underbrace{\begin{bmatrix} \boldsymbol{M}_{k,1} & \boldsymbol{M}_{k,2} \\ \boldsymbol{M}_{k,3} & \boldsymbol{M}_{k,4} \end{bmatrix}}_{\boldsymbol{M}_k} \times \underbrace{\begin{bmatrix} \boldsymbol{C}(k) \\ \boldsymbol{C}^{\#}(k) \end{bmatrix}}_{\boldsymbol{C}_k} + \underbrace{\begin{bmatrix} \boldsymbol{W}(k) \\ \boldsymbol{W}^{\#}(k) \end{bmatrix}}_{\boldsymbol{W}_k} \tag{26}$$

where

$$\boldsymbol{M}_{k,1} = \boldsymbol{S}_p(k) + \widehat{\boldsymbol{A}}_t(k)\boldsymbol{S}^{\#}_p(k)$$

$$\boldsymbol{M}_{k,2} = \widehat{\boldsymbol{A}}^{\#}_t(k)\widehat{\boldsymbol{A}}_r(k)\boldsymbol{S}_p(k) + \widehat{\boldsymbol{A}}_r(k)\boldsymbol{S}^{\#}_p(k)$$

$$\boldsymbol{M}_{k,3} = \widehat{\boldsymbol{A}}_t(k)\widehat{\boldsymbol{A}}^{\#}_r(k)\boldsymbol{S}^{\#}_p(k) + \widehat{\boldsymbol{A}}^{\#}_r(k)\boldsymbol{S}_p(k)$$

$$\boldsymbol{M}_{k,4} = \boldsymbol{S}^{\#}_p(k) + \widehat{\boldsymbol{A}}^{\#}_t(k)\boldsymbol{S}_p(k)$$

and $\boldsymbol{S}_p$ is an individual OFDM training symbol per OFDM frame, $\widehat{\boldsymbol{A}}_t$ and $\widehat{\boldsymbol{A}}_r$ are the estimation results of the iterative estimation approach proposed in Section 3.1. Thus the time-varying channel-related parameter of $\boldsymbol{C}$ can be estimated in each OFDM frame by

$$\widehat{\boldsymbol{C}}_k = (\delta\boldsymbol{I} + \boldsymbol{M}^{\mathrm{H}}_k\boldsymbol{M}_k)^{-1}\boldsymbol{M}^{\mathrm{H}}_k\boldsymbol{R}_k. \tag{27}$$

To further improve the accuracy of estimation, we exploit the time-domain property of $\boldsymbol{C}$ and find that the effective length of the impulse responses is $L + 2K - 2$, with L denoting the length of wireless channel impulse responses and K denoting the taps of imbalance filters at both transmitter and receiver. This effective length of distorted channel impulse responses is shorter than N under normal conditions. Thus the estimation noise could be suppressed by limiting the effective impulse responses length of $\boldsymbol{C}$ as

$$\widehat{\widehat{\boldsymbol{C}}}_k = \boldsymbol{F}\begin{pmatrix} \boldsymbol{U}\boldsymbol{F}^H\widehat{\boldsymbol{C}}_k \\ \boldsymbol{0}_{N-L-2K+2} \end{pmatrix} \tag{28}$$

where $\boldsymbol{U} = [\boldsymbol{I}_{L+2K-2} \quad \boldsymbol{0}_{(L+2K-2)\times(N-L-2K+2)}]$.

Finally, the LS detection results could be obtained by substituting $\widehat{\boldsymbol{A}}_t$, $\widehat{\boldsymbol{A}}_r$ and $\widehat{\widehat{\boldsymbol{C}}}$ into (17):

$$\widehat{\boldsymbol{S}}_k = (\delta\boldsymbol{I} + \boldsymbol{P}_k^{\mathrm{H}}\boldsymbol{P}_k)^{-1}\boldsymbol{P}_k^{\mathrm{H}}\boldsymbol{R}_k. \tag{29}$$

Evidently, with the estimation result of the proposed scheme, the mirror interference of $\boldsymbol{S}^{\#}$ can be canceled in (29).

In summary, the proposed decoupled estimation scheme of frequency-dependent IQI and channel is listed as follows.

(i) Initial communication stage

Input Two OFDM training groups

Use (20), (21) to estimate $\boldsymbol{H}_{a1}$, $\boldsymbol{H}_{b1}$ and $\boldsymbol{H}_{a2}$, $\boldsymbol{H}_{b2}$.

Initialize $\boldsymbol{C}$ with $\boldsymbol{C}_1(k) = \boldsymbol{H}_{a1}(k)$ and $\boldsymbol{C}_2(k) = \boldsymbol{H}_{a2}(k)$.

Iteration

While number of iterations $< I_t$, **do**

Estimate $\boldsymbol{A}_t$ and $\boldsymbol{A}_r$ with (23).

Update $\boldsymbol{C}$ with (25).

end while

Output Estimation of $\boldsymbol{A}_t$ and $\boldsymbol{A}_r$

(ii) Normal communication stage

Input Estimation results of $\boldsymbol{A}_t$, $\boldsymbol{A}_r$ and one OFDM training symbol

Iteration

While channel changes **do**

Estimate $\boldsymbol{C}$ with (27) and (28).

end while

Output Estimation result of $\boldsymbol{C}$.

4. Simulation results and analysis

In this section, simulation results are presented for the proposed decoupled estimation scheme. System parameters are chosen as follows: OFDM subcarrier is $N = 128$, the signal bandwidth is 5 MHz, the carrier frequency is 2 GHz, and the digital modulation is 16 quadrature amplitude modulation (QAM). Typical urban channel with maximum path delay of 2 μs is employed as in [15]. The IQI parameters are chosen as follows: the amplitude mismatch is 1 dB in transmitter and 1.5 dB in receiver, the phase shift mismatch is 2° in transmitter and 3° in receiver, the imbalance filters are $\boldsymbol{I}_t = [0.993\ 7, 0.112\ 5, 0]$, $\boldsymbol{Q}_t = [0.948\ 7, 0.282\ 8, 0.141\ 4]$, $\boldsymbol{I}_r = [0.993\ 7, -0.112\ 5, 0]$, and $\boldsymbol{Q}_r = [1, 0, 0]$.

Fig. 1 compares the uncoded bit error rate (BER) performance of the proposed scheme with the LS scheme in [9] and the GE estimation in [10], where the curve of "No compensation" refers to a system with IQI but no IQI estimation and compensation, the curve of "Ideal channel est" refers to a system with ideal knowledge of IQI and channel, T_s refers to the number of training symbols required by the corresponding estimation scheme at each time of channel variation. As can be seen by the curve of "No compensation", the degradation in BER performance due to the IQI is significant. For the other curves, the LS-based IQI compensation scheme in [9] is adopt to compensate the negative impact of IQI with the corresponding estimation results of IQI and channel. It is shown that the BER performance from both GE and the proposed estimation scheme follows the ideal channel estimation well. The BER performance from the LS scheme with even much more training symbols is relatively poor.

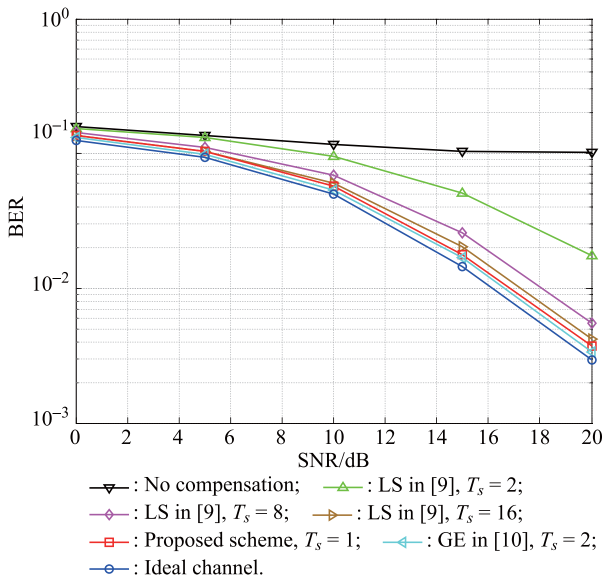

Fig. 1 BER performance of the proposed estimation scheme, LS, and GE

Fig. 2 illustrates the normalized mean square error (NMSE) performance of the IQI and channel estimation schemes. It is also verified by Fig. 2 that the proposed estimation scheme outperforms LS scheme and approaches the performance of GE. Meanwhile, the proposed estimation scheme requires only half training overhead of the GE at each time of channel variation.

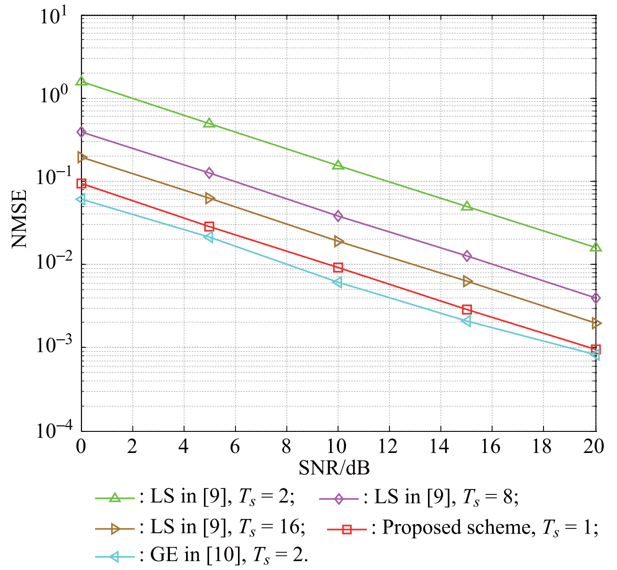

Fig. 2 NMSE performance of the proposed estimation scheme, LS, and GE

Fig. 3 shows the NMSE performance versus the number of iterations of the proposed iterative estimation method of $\boldsymbol{A}_t$ and $\boldsymbol{A}_r$. It can be found from this figure that the estimation of $\boldsymbol{A}_t$ and $\boldsymbol{A}_r$ converges after only four iterations. Thus the number of iterations is set to four in our simulations.

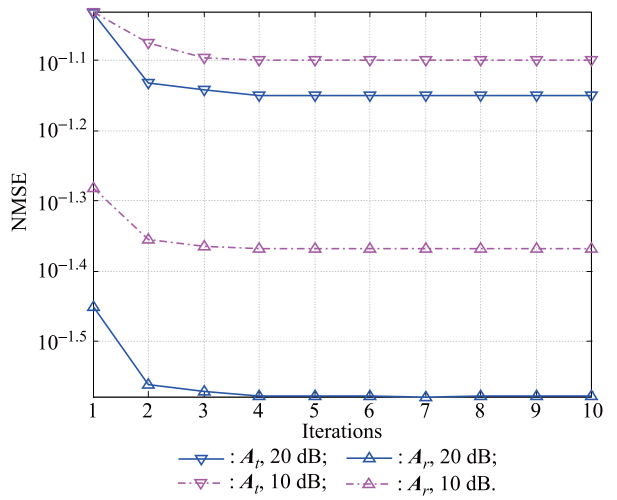

Fig. 3 NMSE performance v.s. iterations of estimation of $\boldsymbol{A}_t$ and $\boldsymbol{A}_r$

The computational complexity (complex multiplications) of our proposed scheme at each time of channel variation (estimation of $\boldsymbol{C}$) is $(\log_2 N + 47/3) \times N$, which is lower than the LS scheme with $T_s = 2$ s of $280/3 \times N$, and is comparable to the complexity of GE, which is $2(\log_2 N + 3) \times N$. The estimation complexity for estimating $\boldsymbol{A}_t$ and $\boldsymbol{A}_r$ is $4(\log_2 N + 1)N + (155/3)NI_t$, where I_t denotes the number of iterations. Considering the estimation of $\boldsymbol{A}_t$ and $\boldsymbol{A}_r$ is just performed at the initial communication stage once, it will not remarkably impact the whole computational complexity of the proposed scheme.

5. Conclusions

In this paper, a decoupled IQI and channel estimation scheme is proposed for the OFDM systems with frequency-dependent IQIs. The IQI parameters are iteratively estimated at the initial communication stage. Then only the channel-related parameters need to be re-estimated at each time of channel variation. The simulation results show that the proposed scheme demonstrates good estimation performance with much lower training overhead than the LS and GE.

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