Journal of Southeast University (English Edition) Vol. 30, No. 4, pp. 416 - 421 Dec. 2014 ISSN 1003—7985

High-performance channel estimation and compensation scheme for OFDM receivers with IQ imbalances

Shu Feng[1,2,3] Tong Juanjuan[1] Li Jun[1] Wang Jin[1] Gu Chen[1] Lu Jinhui[1]

([1]School of Electronic and Optical Engineering, Nanjing University of Science and Technology, Nanjing 210094, China)
([2]National Mobile Communications Research Laboratory, Southeast University, Nanjing 210096, China)
([3]Ministerial Key Laboratory of JGMT, Nanjing University of Science and Technology, Nanjing 210094, China)

Abstract: A pilot pattern across two orthogonal frequency division multiplexing (OFDM) symbols with a special structure is designed for the channel estimation of OFDM systems with inphase and quadrature (IQ) imbalances at the receiver. A high-efficiency time-domain (TD) least square (LS) channel estimator and a low-complexity frequency-domain Gaussian elimination (GE) equalizer are proposed to eliminate IQ distortion. The former estimator can significantly suppress channel noise by a factor $N/(L+1)$ over the existing frequency-domain (FD) LS, where N and $L+1$ are the total number of subcarriers and the length of cyclic prefix, and the proposed GE requires only $2N$ complex multiplications per OFDM symbol. Simulation results show that by exploiting the TD property of the channel, the proposed TD-LS channel estimator obtains a significant signal-to-noise ratio gain over the existing FD-LS one, whereas the proposed low-complexity GE compensation achieves the same bit error rate (BER) performance as the existing LS one.

Key words: inphase and quadrature (IQ) imbalance; equalizer; channel estimation; time domain; frequency domain; least square

doi: 10.3969/j. issn. 1003 - 7985. 2014. 04. 003

Orthogonal frequency division multiplexing (OFDM) has been adopted by several standards such as LTE/LTE-advanced, wireless local area network (IEEE 802.11a, g and n), wireless metropolitan area network (IEEE 802.16d, e and m), digital audio broadcasting, DVB-T/C, digital radio mondiale and digital video broadcasting. Compared with the heterodyne receiver, the receiving RF architecture of direct conversion has been recently reconsidered as a promising solution in OFDM systems to reduce the cost and power consumption of the receiver[1-3]. However, the receiving RF architecture of direct conversion is severely distorted by gain and phase imbalances between the I and Q paths due to imperfections of the analog component design[1-3]. This will seriously damage the orthogonality among subcarriers in OFDM systems and yield inter-carrier interference, which will form a large bit error rate (BER) floor. Therefore, the estimation and compensation of IQ imbalance in the direct conversion receivers are extremely important for improving system performance.

The schemes for cancelling IQ imbalance have been investigated by several scholars. In Ref. [1], the authors derived the signal-to-noise ratio (SNR) loss of IQ imbalances in OFDM receivers and proposed several frequency-domain (FD) and time-domain (TD) compensation methods, such as post-FFT least-squares, adaptive least mean square (LMS) and pre-FFT TD compensation to eliminate IQ distortions. These methods were extended to IQ imbalances at both the transmitter and the receiver[4]. Blind estimation and compensation schemes in the time domain were also proposed[5]. The joint estimation of the IQ imbalance and several other impairments such as phase noise and frequency offset were investigated in Refs. [6 - 10]. In Ref. [6], a finite impulse response (FIR) filter followed by an asymmetric phase compensator was proposed to correct both frequency dependent and frequency independent IQ imbalances. In Ref. [8], a differential filter was employed to estimate both frequency offset and IQ imbalance. A compensation method based on the subcarrier allocation of OFDM signals was proposed in Ref. [9]. Feigin et al. [11] extended the research of Tx/Rx IQ imbalances to the case of packet-switched systems. Narasimhan et al. [12-13] focused on pilot designs and reduced complexity compensation in MIMO-OFDM systems with IQ imbalances. Gregorio et al. [14] proposed a MIMO-PD (pre-distorter) that compensates both crosstalk and IQ imbalance in MIMO-OFDM systems. In Ref. [15], the digital compensation of both the transmitter and the receiver IQ imbalances in MIMO-OFDM transmission over doubly selective channels was studied, and two receiver schemes were proposed to simultaneously mitigate the IQ imbalance and channel time variation effects by using a novel IQ formulation. In Ref. [16], the authors

Received 2014-04-20.

Biography: Shu Feng (1973—), male, doctor, professor, shufeng@njust. edu. cn.

Foundation items: The Open Research Fund of National Mobile Communications Research Laboratory of Southeast University (No. 2013D02), the Fundamental Research Funds for the Central Universities (No. 30920130122004), the National Natural Science Foundation of China (No. 61271230, 61472190).

Citation: Shu Feng, Tong Juanjuan, Li Jun, et al. High-performance channel estimation and compensation scheme for OFDM receivers with IQ imbalances[J]. Journal of Southeast University (English Edition), 2014, 30(4): 416 - 421. [doi: 10.3969/j. issn. 1003 - 7985. 2014. 04. 003]

proposed two pre-compensation algorithms with and without IQ imbalance parameters in order to compensate for the channel non-reciprocity caused by a frequency-independent IQ imbalance. In Ref. [17], authors utilized convex optimization methods to optimize the power of all active subcarriers, and employed the adaptive Markov chain Monte Carlo model to select a training sequence to minimize the mean square error (MSE) of the channel estimator while suppressing the effect of the IQ imbalance. Additionally, a joint estimation of IQ imbalance and channel impulse response was proposed in Ref. [18] for MIMO-OFDM systems.

Unfortunately, in Ref. [1], the FD-LS channel estimation did not exploit the TD property of the channel with IQ imbalance. Thus, it required more than twenty training OFDM symbols to achieve the BER performance of ideal channel knowledge with no IQ imbalances (abbreviated as ideal IQ below). Clearly, this scheme had a low bandwidth efficiency. To solve this problem, an LS channel estimator has been designed, which fully exploits the TD property of channel parameters and significantly reduces the impact of channel noise. Also, it requires only two OFDM symbols to obtain the BER performance of the system without IQ imbalances. It should be noted that "equalizer" is equivalent to "compensation" in this paper.

1 System Model

In the OFDM systems with IQ imbalance as shown in Fig. 1, an OFDM symbol has N subcarriers, and the transmitted block of N data symbols over N subcarriers is denoted as

$$\boldsymbol{s} = [s(1), s(2), \ldots, s(N)]^{\mathrm{T}} \tag{1}$$

Taking IDFT operation on Eq. (1) yields

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

where

$$\boldsymbol{F}(m, n) = \frac{1}{\sqrt{N}}\exp\left(\frac{-\mathrm{j}2\pi(n-1)(m-1)}{N}\right)$$
$$m, n \in \{1, 2, \ldots, N\} \tag{3}$$

Then, the received OFDM symbol before being distorted by IQ imbalance is expressed as

$$\bar{\boldsymbol{y}} = \boldsymbol{F}^{\mathrm{H}}\boldsymbol{\Lambda}\boldsymbol{F}\bar{s} + \bar{\boldsymbol{w}} \tag{4}$$

where $\boldsymbol{\Lambda} = \mathrm{diag}\{\boldsymbol{H}\}$, and

$$\boldsymbol{H} = \boldsymbol{F}\begin{bmatrix} \boldsymbol{h} \\ \boldsymbol{0}_{(N-L-1)\times 1} \end{bmatrix} \tag{5}$$

where $\boldsymbol{h} = [h(1), h(2), \ldots, h(L+1)]^{\mathrm{T}}$ is the channel impulse response (CIR). The received OFDM symbol after being distorted by IQ imbalance is written as

$$\bar{z} = \mu\bar{\boldsymbol{y}} + \nu\bar{\boldsymbol{y}}^{*} \tag{6}$$

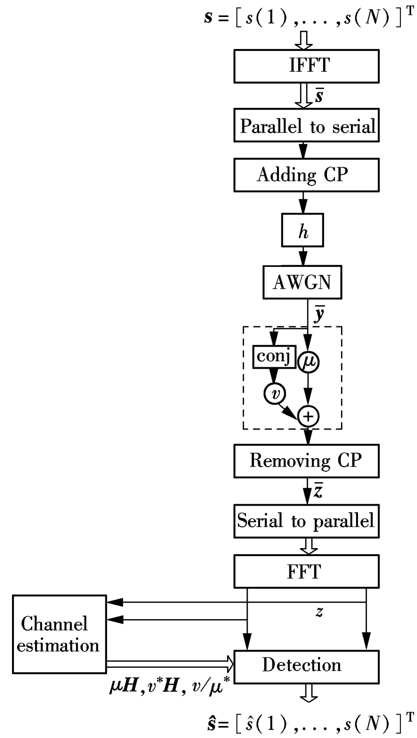


Fig. 1 Discrete baseband OFDM systems with an IQ imbalance at the receiver

where $\mu = \cos(\theta/2) + \mathrm{j}\alpha\sin(\theta/2)$ and $\nu = \alpha\cos(\theta/2) - \mathrm{j}\sin(\theta/2)$ with θ and α being phase and amplitude imbalances between I and Q branches[1], respectively. Taking FFT operation on Eq. (6) gives

$$z = \mu\mathrm{diag}\{\boldsymbol{H}\}\boldsymbol{s} + \nu\mathrm{diag}\{\boldsymbol{H}^{\#}\}\boldsymbol{s}^{\#} + \boldsymbol{w} \tag{7}$$

where the operation # is defined as [1]

$$\boldsymbol{X}^{\#} = [X^{*}(1), X^{*}(N), \ldots, X^{*}(N/2+2), X^{*}(N/2+1),$$
$$X^{*}(N/2), \ldots, X^{*}(2)]^{\mathrm{T}} \tag{8}$$

with

$$\boldsymbol{X} = [X(1), X(2), \ldots, X(N/2), X(N/2+1),$$
$$X(N/2+2), \ldots, X(N)]^{\mathrm{T}} \tag{9}$$

From Ref. [1], if $\boldsymbol{X} = \boldsymbol{F}\boldsymbol{x}$, then

$$\boldsymbol{X}^{\#} = (\boldsymbol{F}\boldsymbol{x})^{\#} = \boldsymbol{F}\boldsymbol{x}^{*} \tag{10}$$

Thus, we obtain the following identity:

$$(\boldsymbol{X}^{\#})^{\#} = (\boldsymbol{F}\boldsymbol{x}^{*})^{\#} = \boldsymbol{F}\boldsymbol{x}^{**} = \boldsymbol{X} \tag{11}$$

In the following, channel $\boldsymbol{h}$ is assumed to be constant during one frame and changes from one frame to another.

2 Proposed Pilot Pattern, Channel Estimator and Equalizer

In the following, a low-complexity Gaussian elimination is proposed to cancel the IQ distortion by using operation #. Then, a particular training pattern using two adjacent OFDM symbols is designed and a TD-LS channel estimation is presented to provide a high-precision estimation of the channel parameters ν/μ^{*}, $\mu\boldsymbol{H}$ and $\nu^{*}\boldsymbol{H}$. Additionally, the estimation of μH and $\nu^{*}H$ by the FD-LS in Ref. [1] is also transformed to the time domain and again back to the frequency domain to improve the accuracy of the estimation.

2.1 Gaussian elimination equalizer in FD

Since $(\mathrm{diag}\{\boldsymbol{H}\}\boldsymbol{s})^{\#} = \mathrm{diag}\{\boldsymbol{H}^{\#}\}\boldsymbol{s}^{\#}$ and $(\mathrm{diag}\{\boldsymbol{H}^{\#}\}\boldsymbol{s}^{\#})^{\#} = \mathrm{diag}\{\boldsymbol{H}\}\boldsymbol{s}$, making a mathematical operation # on Eq. (7) yields

$$z^{\#} = \nu^{*}\,\mathrm{diag}\{\boldsymbol{H}\}\boldsymbol{s} + \mu^{*}\,\mathrm{diag}\{\boldsymbol{H}\}^{\#}\boldsymbol{s}^{\#} + \boldsymbol{w}^{\#} \tag{12}$$

Defining $\kappa = \nu/\mu^{*}$ which can be estimated by existing channel estimators, based on Eqs. (7) and (12), we construct the following formula:

$$z - \kappa z^{\#} = (\mu - \kappa\nu^{*})\,\mathrm{diag}\{\boldsymbol{H}\}\boldsymbol{s} + \boldsymbol{w} - \kappa\boldsymbol{w}^{\#} \tag{13}$$

which has removed the IQ distortion in Eq. (7) and it is represented as

$$z - \kappa z^{\#} = (\mathrm{diag}\{\mu\boldsymbol{H}\} - \kappa\,\mathrm{diag}\{\nu^{*}\boldsymbol{H}\})\boldsymbol{s} + \boldsymbol{w} - \kappa\boldsymbol{w}^{\#} \tag{14}$$

Eq. (14) yields the following detector as

$$\hat{\boldsymbol{s}} = \{\mathrm{diag}\{\mu\boldsymbol{H}\} - \kappa\,\mathrm{diag}\{\nu^{*}\boldsymbol{H}\}\}^{-1}(z - \kappa z^{\#}) \tag{15}$$

which can be simplified as

$$\hat{\boldsymbol{s}}(k) = \frac{z(k) - \kappa z^{\#}(k)}{\mu\boldsymbol{H}(k) - \kappa\nu^{*}\boldsymbol{H}(k)} \tag{16}$$

To complete the detection of $\boldsymbol{s}$ in Eqs. (15) or (16), we need to estimate parameters κ, $\mu\boldsymbol{H}$, and $\nu^{*}\boldsymbol{H}$ in advance, where

$$\mu\boldsymbol{H} = \boldsymbol{F}\begin{bmatrix} \mu\boldsymbol{h} \\ \boldsymbol{0}_{(N-L-1)\times 1} \end{bmatrix} \tag{17}$$

$$\nu^{*}\boldsymbol{H} = \boldsymbol{F}\begin{bmatrix} \nu^{*}\boldsymbol{h} \\ \boldsymbol{0}_{(N-L-1)\times 1} \end{bmatrix} \tag{18}$$

It should be noted that Eqs. (17) and (18) are the building blocks of channel estimation in Section 1.

Similar to Ref. [1], the SNR loss of the error variance given by Eq. (16) over the error variance $\sigma_w^2/|\boldsymbol{H}(k)|^2$ (SNR without IQ imbalance) is defined as

$$g_{\mathrm{loss}} = 10\log\left(\frac{1 + |\kappa|^2}{|\mu|^2 - 2\mathrm{Re}(\kappa\nu^{*}\mu) + |\kappa|^2|\nu|^2}\right) \tag{19}$$

where $\mathrm{Re}(x)$ denotes the real part of x.

2.2 Pilot pattern design and TD-LS estimation of channel parameters

Let us devise the frequency-domain pilot vectors of two pilot OFDM symbols at the start of a frame as

$$\boldsymbol{s}_1 = \begin{bmatrix} \eta \\ s_{\mathrm{p}} \\ \eta \\ \boldsymbol{0}_{(N/2-1)\times 1} \end{bmatrix}, \qquad \boldsymbol{s}_2 = \begin{bmatrix} \mathrm{j}\eta \\ \boldsymbol{0}_{(N/2-1)\times 1} \\ \mathrm{j}\eta \\ \boldsymbol{s}_{\mathrm{p}} \end{bmatrix}$$

where $\eta = 2\overline{P}_{\mathrm{s}}$ with $\overline{P}_{\mathrm{s}}$ being the average transmit power for signal constellation, and $\boldsymbol{s}_{\mathrm{p}}$ is an $N/2-1$ dimensional column pilot vector with $\mathrm{tr}\{\boldsymbol{E}(\boldsymbol{s}_{\mathrm{p}}\boldsymbol{s}_{\mathrm{p}}^{\mathrm{H}})\} = (N-1)\overline{P}_{\mathrm{s}}$. After $\boldsymbol{s}_{\mathrm{p}}$ experiences the multipath channel, we obtain the following received training vectors on the frequency domain as

$$z_1(2:N/2) = \mu\,\mathrm{diag}\{\boldsymbol{H}(2:N/2)\}\boldsymbol{s}_{\mathrm{p}} + \boldsymbol{w}_1(2:N/2) \tag{20}$$

$$z_1(N/2+2:N) = \nu\,\mathrm{diag}\{\boldsymbol{H}^{\#}(N/2+2:N)\}\dot{\boldsymbol{s}}_{\mathrm{p}} + \boldsymbol{w}_1(N/2+2:N) \tag{21}$$

$$z_2(2:N/2) = \nu\,\mathrm{diag}\{\boldsymbol{H}^{\#}(2:N/2)\}\dot{\boldsymbol{s}}_{\mathrm{p}} + \boldsymbol{w}_2(2:N/2) \tag{22}$$

$$z_2(N/2+2:N) = \mu\,\mathrm{diag}\{\boldsymbol{H}(N/2+2:N)\}\boldsymbol{s}_{\mathrm{p}} + \boldsymbol{w}_2(N/2+2:N) \tag{23}$$

and

$$z_1(1) = \mu H(1)\eta + \nu H(1)^{*}\eta + \boldsymbol{w}_1(1) \tag{24}$$

$$z_1(N/2+1) = \mu H(N/2+1)\eta + \nu H(N/2+1)^{*}\eta + \boldsymbol{w}_1(N/2+1) \tag{25}$$

$$z_2(1) = \mathrm{j}\mu H(1)\eta - \mathrm{j}\nu H(1)^{*}\eta + \boldsymbol{w}_2(1) \tag{26}$$

$$z_2(N/2+1) = \mathrm{j}\mu H(N/2+1)\eta - \mathrm{j}\nu H(N/2+1)^{*}\eta + \boldsymbol{w}_2(N/2+1) \tag{27}$$

Then, combining Eqs. (24) to (27) forms the following equations:

$$-0.5\mathrm{j}z_2(1) + 0.5z_1(1) = \mu H(1)\eta - 0.5\mathrm{j}\boldsymbol{w}_2(1) + 0.5\boldsymbol{w}_1(1) \tag{28}$$

$$0.5\mathrm{j}z_2(1) + 0.5z_1(1) = \nu H(1)^{*}\eta + 0.5\mathrm{j}\boldsymbol{w}_2(1) + 0.5\boldsymbol{w}_1(1) \tag{29}$$

$$-0.5\mathrm{j}z_2(N/2+1) + 0.5z_1(N/2+1) = \mu H(N/2+1)\eta - 0.5\mathrm{j}\boldsymbol{w}_2(N/2+1) + 0.5\boldsymbol{w}_1(N/2+1) \tag{30}$$

$$0.5\mathrm{j}z_2(N/2+1) + 0.5z_1(N/2+1) = \nu H(N/2+1)^{*}\eta + 0.5\mathrm{j}\boldsymbol{w}_2(N/2+1) + 0.5\boldsymbol{w}_1(N/2+1) \tag{31}$$

where $\dot{\boldsymbol{s}}_{\mathrm{p}} = \boldsymbol{s}_1^{\#}(N/2+2:N)$.

Stacking Eqs. (20), (23), (28) and (30) gives a large matrix-vector form as

$$\tilde{z}_{\mathrm{a}} = \begin{bmatrix} -0.5\mathrm{j}z_2(1) + 0.5z_1(1) \\ z_1(2:N/2) \\ -0.5\mathrm{j}z_2(N/2+1) + 0.5z_1(N/2+1) \\ z_2(N/2+2:N) \end{bmatrix} = \mathrm{diag}\{\mu\boldsymbol{H}\}\begin{bmatrix} \eta \\ \boldsymbol{s}_{\mathrm{p}} \\ \eta \\ \boldsymbol{s}_{\mathrm{p}} \end{bmatrix} + \underbrace{\begin{bmatrix} -0.5\mathrm{j}\boldsymbol{w}_2(1) + 0.5\boldsymbol{w}_1(1) \\ \boldsymbol{w}_1(2:N/2) \\ -0.5\mathrm{j}\boldsymbol{w}_2(N/2+1) + 0.5\boldsymbol{w}_1(N/2+1) \\ \boldsymbol{w}_2(N/2+2:N) \end{bmatrix}}_{\tilde{\boldsymbol{w}}_{\mathrm{a}}} =$$

$$\mathrm{diag}\underbrace{\begin{bmatrix}\eta\\ s_{\mathrm{p}}\\ \eta\\ s_{\mathrm{p}}\end{bmatrix}}_{\tilde{s}_{\mathrm{p}}}(\mu\boldsymbol{H})+\tilde{\boldsymbol{w}}_{\mathrm{a}}=\mathrm{diag}\{\tilde{s}_{\mathrm{p}}\}\boldsymbol{F}\begin{bmatrix}\mu\boldsymbol{h}\\ \boldsymbol{0}_{(N-L-1)\times 1}\end{bmatrix}+\tilde{\boldsymbol{w}}_{\mathrm{a}} \tag{32}$$

Thus, the TD-LS estimate of $\mu\boldsymbol{h}$ is given as

$$\mu\hat{\boldsymbol{h}}_{\text{TD-LS}}=\boldsymbol{P}\boldsymbol{F}^{\mathrm{H}}\mathrm{diag}\{\tilde{s}_{\mathrm{p}}\}^{-1}\tilde{z}_{\mathrm{a}}=\mu\boldsymbol{h}+\boldsymbol{P}\boldsymbol{F}^{\mathrm{H}}\mathrm{diag}\{\tilde{s}_{\mathrm{p}}\}^{-1}\tilde{\boldsymbol{w}}_{\mathrm{a}} \tag{33}$$

where $\boldsymbol{P}=[\boldsymbol{I}_{L+1}\quad \boldsymbol{0}_{(L+1)\times(N-L-1)}]$. Then, we have the estimate of $\mu\boldsymbol{H}$ as

$$\mu\hat{\boldsymbol{H}}_{\text{TD-LS}}=\boldsymbol{F}\begin{bmatrix}\mu\hat{\boldsymbol{h}}_{\text{TD-LS}}\\ \boldsymbol{0}_{(N-L-1)\times 1}\end{bmatrix} \tag{34}$$

In the same way, we combine Eqs. (21), (22), (29) and (31) into a large matrix-vector form

$$\tilde{z}_{\mathrm{b}}^{\#}=\begin{bmatrix}0.5\mathrm{j}z_2(1)+0.5z_1(1)\\ z_2(2:N/2)\\ 0.5\mathrm{j}z_2(N/2+1)+0.5z_1(N/2+1)\\ z_1(N/2+2:N)\end{bmatrix}=$$

$$\mathrm{diag}\{\nu\boldsymbol{H}^{\#}\}\underbrace{\begin{bmatrix}\eta\\ \dot{s}_{\mathrm{p}}\\ \eta\\ \dot{s}_{\mathrm{p}}\end{bmatrix}}_{\tilde{s}_{\mathrm{p}}}+\underbrace{\begin{bmatrix}0.5\mathrm{j}\boldsymbol{w}_2(1)+0.5\boldsymbol{w}_1(1)\\ \boldsymbol{w}_2(2:N/2)\\ 0.5\mathrm{j}\boldsymbol{w}_2(N/2+1)+0.5w_1(N/2+1)\\ \boldsymbol{w}_1(N/2+2:N)\end{bmatrix}}_{\tilde{\boldsymbol{w}}_{\mathrm{b}}} \tag{35}$$

whose # operation forms

$$\tilde{z}_{\mathrm{b}}=(\tilde{z}_{\mathrm{b}}^{\#})^{\#}=\mathrm{diag}\{\nu^{*}\boldsymbol{H}\}\tilde{\boldsymbol{s}}_{\mathrm{p}}^{\#}+\tilde{\boldsymbol{w}}_{\mathrm{b}}^{\#}=$$
$$\mathrm{diag}\{\tilde{s}_{\mathrm{p}}\}\boldsymbol{F}\begin{bmatrix}\nu^{*}\boldsymbol{h}\\ \boldsymbol{0}_{(N-L-1)\times 1}\end{bmatrix}+\tilde{\boldsymbol{w}}_{\mathrm{b}}^{\#} \tag{36}$$

and Eq. (36) results in the LS estimation of $\nu^{*}\boldsymbol{h}$.

$$\nu^{*}\hat{\boldsymbol{h}}_{\text{TD-LS}}=$$
$$\boldsymbol{P}\boldsymbol{F}^{\mathrm{H}}\mathrm{diag}\{\tilde{s}_{\mathrm{p}}\}^{-1}\tilde{z}_{\mathrm{b}}=\nu^{*}\boldsymbol{h}+\boldsymbol{P}\boldsymbol{F}^{\mathrm{H}}\mathrm{diag}\{\tilde{s}_{\mathrm{p}}\}^{-1}\tilde{\boldsymbol{w}}_{\mathrm{b}}^{\#} \tag{37}$$

Then, we have the estimate of $\nu^{*}\boldsymbol{H}$ as

$$\nu^{*}\hat{\boldsymbol{H}}_{\text{TD-LS}}=\boldsymbol{F}\begin{bmatrix}\nu^{*}\hat{\boldsymbol{h}}_{\text{TD-LS}}\\ \boldsymbol{0}_{(N-L-1)\times 1}\end{bmatrix} \tag{38}$$

In terms of Eqs. (33), (34), (37), and (38), the estimate of κ can be defined as

$$\hat{\kappa}_{\text{TD-LS}}=\frac{\sum_{k=1}^{N}\nu^{*}\hat{\boldsymbol{H}}_{\text{TD-LS}}^{*}(k)}{\sum_{k=1}^{N}\mu\hat{\boldsymbol{H}}_{\text{TD-LS}}^{*}(k)}=\frac{\sum_{k=1}^{L+1}\nu^{*}\hat{\boldsymbol{h}}_{\text{TD-LS}}^{*}(k)}{\sum_{k=1}^{L+1}\mu\hat{\boldsymbol{H}}_{\text{TD-LS}}^{*}(k)} \tag{39}$$

From Eqs. (33), (34), (37) and (38), we obtain the MSEs of $\nu^{*}\boldsymbol{H}$ and $\mu\boldsymbol{H}$ as follows:

$$\frac{E\{(\mu\hat{\boldsymbol{H}}_{\text{TD-LS}}-\mu\boldsymbol{H})^{\mathrm{H}}(\mu\hat{\boldsymbol{H}}_{\text{TD-LS}}-\mu\boldsymbol{H})\}}{N}=$$
$$\frac{E\{(\nu^{*}\hat{\boldsymbol{H}}_{\text{TD-LS}}-\nu^{*}\boldsymbol{H})^{\mathrm{H}}(\nu^{*}\hat{\boldsymbol{H}}_{\text{TD-LS}}-\nu^{*}\boldsymbol{H})\}}{N}=\frac{(L+1)\beta}{N\gamma} \tag{40}$$

where γ is the SNR and it is defined as $E\{s(k)^{*}(s(k))\}/\sigma_n^{2}$ [19], and

$$\beta=\frac{E\{s(k)^{*}(s(k))\}}{E\{(1/s(k))^{*}(1/s(k))\}} \tag{41}$$

3 Simulation and Discussion

In the following, a typical OFDM system is used to evaluate the performance of the proposed channel estimator, the proposed compensation method and those existing channel estimations and compensation schemes. An ideal OFDM receiver with no IQ imbalance and a receiver with no compensation are used as performance references. It is noted that A stands for the channel estimator and B denotes the equalizer below.

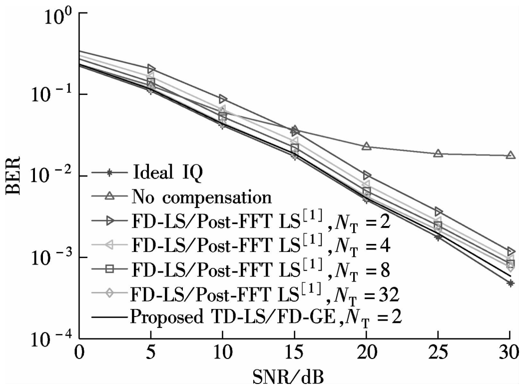
Simulation parameters are as follows: The OFDM symbol length $N=128$, cyclic prefix $L=16$, signal bandwidth BW = 2 MHz, digital modulation QPSK, and the carrier frequency $f_{\mathrm{c}}=2$ GHz. Channel $\boldsymbol{h}$ is selected to be the typical urban (TU) channel.

Fig. 2 compares the proposed schemes with the FD-LS channel estimator plus the Post-FFT LS equalizer (FD-LS/Post-FFT) for different values of IQ imbalance parameters where N_{T} denotes the number of training OFDM symbols (TOSs) per frame. From these figures, it is evident that the proposed joint scheme of the TD-LS estimator plus GE equalizer, with only two TOSs, can achieve the BER performance of ideal IQ at low and medium SNRs, whereas the joint scheme of the FD-LS estimator plus the LS equalizer in Ref. [1] costs approximately 32 TOSs to realize almost the same BER performance. Thus, the proposed scheme has a higher bandwidth efficiency compared to the FD-LS estimator plus the LS equalizer.

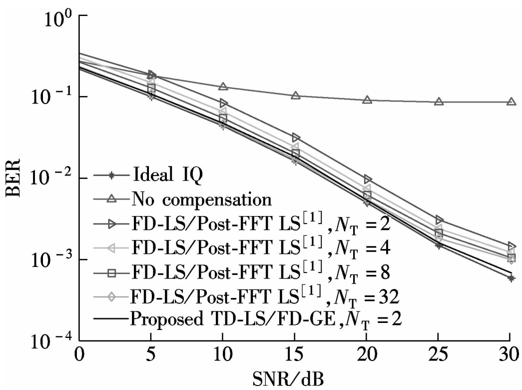
Tab. 1 makes a complexity comparison of four schemes including the proposed scheme of TD-LS/GE, and the existing schemes of FD-LS/Post-FFT, FD-LS/Pre-FFT Corr, and SPP/Pre-FFT Corr in Ref. [1], where N_{F} is the total number of non-training OFDM symbols. The TD-LS/GE scheme put forward has almost the same computational amount as the three existing schemes.

Tab. 1 Complexity comparison for five schemes of channel estimation plus compensation

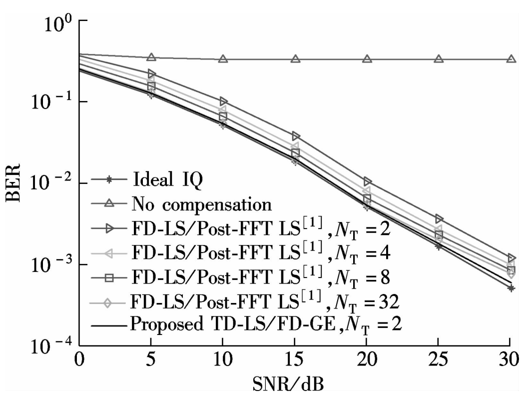
Channel estimator/equalizer	Channel estimation	Distortion correction or equalizer
Proposed TD-LS/FD-GE	$0.5N_{\mathrm{T}}N\log_2 N+0.5N_{\mathrm{T}}N\log_2(L+1)+0.5N_{\mathrm{T}}N$	$(2N+0.5N\log_2 N)N_{\mathrm{F}}$
FD-LS/Post-FFT-LS[1]	$0.5N_{\mathrm{T}}N\log_2 N+16NN_{\mathrm{T}}+0.5NN_{\mathrm{T}}\log_2(L+1)$	$(10N+0.5N\log_2 N)N_{\mathrm{F}}$
FD-LS/Pre-FFT-Corr[1]	$0.5NN_{\mathrm{T}}\log_2 N+16NN_{\mathrm{T}}$	$(2N+0.5N\log_2 N)N_{\mathrm{F}}$
SPP/Pre-FFT-Corr[1]	$0.5N_{\mathrm{T}}N\log_2 N+0.5N_{\mathrm{T}}N\log_2(L+1)+N$	$(2N+0.5N\log_2)N_{\mathrm{F}}$



(a)



(b)



(c)

Fig. 2 BER performance comparison of the proposed joint scheme with existing joint schemes of channel estimation plus compensation. (a) $\theta = 2°$ and $\alpha = 1$ dB; (b) $\theta = 10°$ and $\alpha = 2$ dB; (c) $\theta = 20°$ and $\alpha = 4$ dB

4 Conclusion

In this paper, a joint scheme of combining a TD-LS channel estimator and a FD-GE equalizer is investigated in OFDM systems with IQ imbalance at the receiver. Compared with the existing schemes such as FD-LS/Post-FFT LS, SPP/Pre-FFT Corr and FD-LS/Pre-FFT Corr in Ref. [1], this scheme shows much better BER performance. More importantly, it requires only two OFDM training symbols to achieve the BER performance of the ideal IQ at low and medium SNR regions, whereas three existing schemes, including FD-LS/Post-FFT LS, SPP/Pre-FFT Corr and FD-LS/Pre-FFT Corr, require around twenty OFDM training symbols to achieve the same BER. Concerning complexity, the proposed TD-LS/FD-GE has the same as the existing schemes.

References

[1] Tarighat A, Bagheri R, Sayed A H. Compensation schemes and performance analysis of IQ imbalances in OFDM receivers[J]. *IEEE Transactions on Signal Processing*, 2005, **53**(8): 3257 − 3268.

[2] Razavi B, Behzad R. *RF microelectronics*[M]. New Jersey: Prentice Hall, 1998.

[3] Abidi A A. Direct-conversion radio transceivers for digital communications[J]. *IEEE Journal of Solid-State Circuits*, 1995, **30**(12): 1399 − 1410.

[4] Tarighat A, Sayed A H. Joint compensation of transmitter and receiver impairments in OFDM systems[J]. *IEEE Transactions on Wireless Communications*, 2007, **6**(1): 240 − 247.

[5] Valkama M, Renfors M, Koivunen V. Advanced methods for I/Q imbalance compensation in communication receivers[J]. *IEEE Transactions on Signal Processing*, 2001, **49**(10): 2335 − 2344.

[6] Xing G, Shen M, Liu H. Frequency offset and I/Q imbalance compensation for direct-conversion receivers[J]. *IEEE Transactions on Wireless Communications*, 2005, **4**(2): 673 − 680.

[7] Tubbax J, Come B, Van der Perre L, et al. Compensation of IQ imbalance and phase noise in OFDM systems[J]. *IEEE Transactions on Wireless Communications*, 2005, **4**(3): 872 − 877.

[8] Inamori M, Bostamam A, Sanada Y, et al. IQ imbalance compensation scheme in the presence of frequency offset and dynamic DC offset for a direct conversion receiver[J]. *IEEE Transactions on Wireless Communications*, 2009, **8**(5): 2214 − 2220.

[9] Lin H, Yamashita K. Subcarrier allocation based compensation for carrier frequency offset and I/Q imbalances in OFDM systems[J]. *IEEE Transactions on Wireless Communications*, 2009, **8**(1): 18 − 23.

[10] Tandur D, Moonen M. Joint adaptive compensation of transmitter and receiver IQ imbalance under carrier frequency offset in OFDM-based systems[J]. *IEEE Transactions on Signal Processing*, 2007, **55**(11): 5246 − 5252.

[11] Feigin J, Brady D. Joint transmitter/receiver I/Q imbalance compensation for direct conversion OFDM in packet-switched multipath environments[J]. *IEEE Transactions on Signal Processing*, 2009, **57**(11): 4588 − 4593.

[12] Narasimhan B, Narayanan S, Minn H, et al. Reduced-complexity baseband compensation of joint Tx/Rx I/Q imbalance in mobile MIMO-OFDM[J]. *IEEE Transactions on Wireless Communications*, 2010, **9**(5): 1720 − 1728.

[13] Minn H, Munoz D. Pilot designs for channel estimation of MIMO OFDM systems with frequency-dependent I/Q imbalances[J]. *IEEE Transactions on Communications*, 2010, **58**(8): 2252 − 2264.

[14] Gregorio F, Cousseau J, Werner S, et al. Compensation

of IQ imbalance and transmitter nonlinearities in broadband MIMO-OFDM[C]//2011 *IEEE International Symposium on Circuits and Systems*. Rio de Janeiro, Brazil, 2011: 2393 – 2396.

[15] Beheshti M, Omidi M J, Doost-Hoseini A M. Joint compensation of transmitter and receiver IQ imbalance for MIMO-OFDM over doubly selective channels[J]. *Wireless Personal Communications*, 2013, **70**(2): 537 – 559.

[16] Liu Z J, Sun D C, Wang J L, et al. Impact and compensation of I/Q imbalance on channel reciprocity of time-division-duplexing multiple-input multiple-output systems [J]. *IET Communications*, 2013, **7**(7): 663 – 672.

[17] Manasseh E, Ohno S, Yamamoto T. Efficient training design for estimation of channel, CFO, and receiver I/Q imbalance in MIMO-OFDM systems[C]//2013 13*th International Symposium on Communications and Information Technologies*. Surat Thani, Thailand, 2013: 364 – 368.

[18] Deng J H, Hsieh H S, Feng K T. Low complexity I/Q imbalance and channel estimation techniques for MIMO OFDM systems[C]//2013 15*th IEEE International Conference on Advanced Communication Technology*. Pyeongchang, Korea, 2013: 238 – 243.

[19] Edfors O, Sandell M, Van de Beek J J, et al. OFDM channel estimation by singular value decomposition[J]. *IEEE Transactions on Communications*, 1998, **46**(7): 931 – 939.

接收 IQ 不平衡 OFDM 系统的高性能信道估计和补偿方案

束 锋[1,2,3] 童娟娟[1] 李 隽[1] 王 进[1] 顾 晨[1] 陆锦辉[1]

([1] 南京理工大学电子工程与光电技术学院,南京 210094)

([2] 东南大学移动通信国家重点实验室,南京 210096)

([3] 南京理工大学近程高速目标探测技术国防重点学科实验室,南京 210094)

摘要:针对接收 IQ 不平衡的 OFDM 系统,设计了横跨 2 个 OFDM 符号的特殊训练结构.提出了一种高效的时域最小二乘(TD-LS)信道估计和一种低复杂度的频域高斯消元补偿算法来消除 IQ 失真.前者与传统的频域 LS 算法相比,信道估计噪声的影响降低了 $N/(L+1)$ 倍,其中 N 为子载波总数,$L+1$ 为循环前缀长度;后者实现复杂度低,每个 OFDM 符号仅需要 $2N$ 次复数乘法.仿真结果表明:由于充分挖掘了信道参数的时域特性,提出的 TD-LS 信道估计算法与传统的 FD-LS 算法相比获得了可观的信噪比增益;提出的低复杂度的 GE 补偿算法能够获得与基于 LS 的频域补偿方案几乎相同的误码性能.

关键词:IQ 不平衡;均衡器;信道估计;时域;频域;最小二乘

中图分类号:TN929.5