

# Joint Compensation of IQ Imbalance and Phase Noise in OFDM Wireless Systems

Qiyue Zou, *Member, IEEE*, Alireza Tarighat, *Member, IEEE*, and Ali H. Sayed, *Fellow, IEEE*

**Abstract**—Physical impairments like IQ imbalance and phase noise can cause significant degradation in the performance of wireless communication systems. In this paper, the joint effects of IQ imbalance and phase noise on OFDM systems are analyzed, and a compensation scheme is proposed to improve the system performance in the presence of IQ imbalance and phase noise. The scheme consists of a joint estimation of channel and impairment parameters and a joint data symbol estimation algorithm. It is shown both by theory and computer simulations that the proposed scheme can effectively improve the signal-to-noise ratio at the receiver. As a result, the sensitivity of OFDM receivers to the physical impairments can be significantly lowered, simplifying the RF and analog circuitry design in terms of implementation cost, power consumption, and silicon fabrication yield.

**Index Terms**—Direct-conversion receiver, equalization, IQ imbalance, Orthogonal Frequency Division Multiplexing (OFDM), phase noise.

## I. INTRODUCTION

THE Orthogonal Frequency Division Multiplexing (OFDM) modulation technique has recently received considerable attention and has been chosen for several standards. The surging interest in OFDM has resulted in research activities to make the implementation of OFDM receivers more reliable and less costly in practice. There are mainly two types of OFDM receiver structures: one is the direct-conversion receiver and the other is the Heterodyne receiver [2]. Compared to the Heterodyne receivers, the direct-conversion receivers have the advantages of low cost, low power consumption and easy integration, but they suffer more severely from analog domain impairments. One such impairment is caused by the imperfectness in the process of the radio-frequency (RF) signal down-conversion to baseband. Its effects have been modeled as IQ imbalance and phase noise in the literature [2], [3]. IQ imbalance is the mismatch in amplitude and phase between the I and Q branches in the receiver chain, while phase noise is the random unknown phase difference between the phase of the carrier signal and the phase of the local oscillator. The effects of IQ imbalance

and phase noise on OFDM receivers have been investigated in previous works, such as [4]–[7]. Some algorithms have also been proposed for the compensation of IQ imbalance [8]–[10] or the compensation of phase noise [11]–[16], separately. In [17], the joint effects of IQ imbalance and phase noise on OFDM systems were studied, but the analysis and the proposed compensation scheme were based on the concatenation model of IQ imbalance and phase noise, where only the common error term of phase noise was considered. To our best knowledge, there is still no thorough work on analyzing the system performance degradation caused by the coexistence of IQ imbalance and phase noise, as well as the optimal estimation of channel response and data symbols in the presence of both impairments.

In this paper, we pursue an explicit formulation for the joint effects of IQ imbalance and phase noise, and propose a joint compensation scheme with performance analysis. The scheme consists of a joint channel estimation algorithm and a joint data symbol estimation algorithm. In the channel estimation algorithm, block-type pilot symbols are transmitted periodically, and the channel coefficients are jointly estimated with the IQ imbalance parameters and phase noise components.<sup>1</sup> Instead of estimating the channel coefficients and phase noise in the frequency domain, we estimate them in the time domain by using interpolation techniques to reduce the number of unknowns. The joint estimation technique achieves a more accurate channel estimate than other conventional methods that either ignore the impairments or simply model them as additive Gaussian noise. The mean-square errors of channel estimation are compared with their associated Cramer-Rao lower bounds, which shows that our scheme works well with performance close to the ideal case without the impairments. In the proposed data symbol estimation algorithm, it is shown that the joint compensation can be decomposed into the IQ imbalance compensation followed by the phase noise compensation. During the payload portion of OFDM packets, which contains both data tones and pilot tones, the data symbols and the phase noise components are jointly estimated at the receiver. The performance of the proposed algorithm is analyzed in terms of the improvements in the effective signal-to-noise ratio, and is compared with other compensation methods.

The paper is organized as follows. The next section describes the system model and formulates the joint effects of IQ imbalance and phase noise. The proposed joint channel es-

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Q. Zou was with the Electrical Engineering Department, University of California, Los Angeles, CA 90095, USA. He is now with the WiLinx Corporation, San Diego, CA 90025, USA (e-mail: eqyzou@ee.ucla.edu).

A. Tarighat is with the WiLinx Corporation, San Diego, CA 90025, USA (e-mail: tarighat@ee.ucla.edu).

A. H. Sayed is with the Electrical Engineering Department, University of California, Los Angeles, CA 90095, USA (e-mail: sayed@ee.ucla.edu).

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<sup>1</sup>All standardized OFDM systems today provide such full pilot symbols at the beginning of every packet. Therefore, the proposed scheme does not require any modification to the packet structure and can be applied to existing standards.

timization algorithm is presented in Section III, and the Cramer-Rao lower bounds for estimating the channel response are also derived. The joint data estimation algorithm is presented in Section IV, and its performance is analyzed in terms of the effective signal-to-noise ratio degradation. Simulation results and performance comparison of different algorithms are discussed in Section V.

Throughout this paper, we adopt the following notations.  $(\cdot)^T$  denotes the matrix transpose,  $(\cdot)^*$  represents the matrix conjugate transpose, and  $\text{conj}\{\cdot\}$  takes the complex conjugate of its argument elementwisely.  $\text{Re}\{\cdot\}$  and  $\text{Im}\{\cdot\}$  return the real and imaginary parts of its argument, respectively.  $\text{diag}\{\cdot\}$  represents the diagonal matrix whose diagonal entries are determined by its argument.  $\text{Tr}\{\cdot\}$  returns the trace of a matrix.  $\mathbf{E}\{\cdot\}$  is the expected value with respect to the underlying probability measure.  $\mathbf{I}_K$  is the identity matrix of size  $K \times K$ , and  $\mathbf{I}_\theta$  is the Fisher information matrix associated with the parameter vector  $\theta$ .

## II. SYSTEM MODEL

Fig. 1 shows the block diagrams of a direct-conversion receiver with and without IQ imbalance and phase noise. We first formulate the effects of IQ imbalance and phase noise on the received continuous-time baseband signals in Subsection II-A, and then discuss their effects on the received OFDM symbols in Subsection II-B.

### A. IQ Imbalance and Phase Noise

Let  $x(t)$  be the transmitted continuous-time baseband signal. The transmitted passband signal, namely, the radio-frequency signal  $x_p(t)$  is given by

$$\begin{aligned} x_p(t) &= \text{Re}\{\sqrt{2}x(t)e^{j2\pi f_c t}\} \\ &= \sqrt{2}\text{Re}\{x(t)\}\cos(2\pi f_c t) - \sqrt{2}\text{Im}\{x(t)\}\sin(2\pi f_c t), \end{aligned} \quad (1)$$

where  $f_c$  is the carrier frequency and the normalization factor  $\sqrt{2}$  ensures that  $x(t)$  and  $x_p(t)$  have the same average power. Let  $h_p(t)$  be the continuous-time impulse response function of the passband channel. Then the received passband signal  $y_p(t)$  is given by the convolutional integral of  $x_p(t)$  and  $h_p(t)$  plus additive white Gaussian noise  $w_p(t)$ , i.e.,

$$y_p(t) = \int_{-\infty}^{\infty} h_p(t - \tau)x_p(\tau)d\tau + w_p(t). \quad (2)$$

Let  $h(t) = h_p(t)e^{-j2\pi f_c t}$  represent the continuous-time impulse response function of the equivalent baseband channel. We use

$$y_0(t) = \int_{-\infty}^{\infty} h(t - \tau)x(\tau)d\tau \quad (3)$$

to represent the received baseband signal in the absence of the impairments and noise. Note that the passband signals  $x_p(t)$ ,  $y_p(t)$  and the passband channel response  $h_p(t)$  are all real functions, while the baseband signal  $x(t)$  and the baseband channel response  $h(t)$  are complex. It then follows from (1)-(3) that  $y_p(t)$  can be expressed as

$$y_p(t) = \text{Re}\{\sqrt{2}y_0(t)e^{j2\pi f_c t}\} + w_p(t).$$

In an ideal direct-conversion receiver, as shown in Fig. 1(a), the sinusoidal signals used for I- and Q-branch mixing have

the same amplitude and are orthogonal to each other. Also, their phase is perfectly aligned with the carrier signal. In this case, the received baseband signal after down-conversion is

$$y(t) = y_0(t) + w(t),$$

where  $w(t)$  is the additive Gaussian noise in the baseband. However, as shown in Fig. 1(b), the actual oscillator signals in the I and Q branches are slightly different from the ideal oscillator signals due to the imperfectness of the local oscillator and  $90^\circ$  phase shifter. The constants  $\alpha$  and  $\theta$  model the amplitude and phase imbalances between the I and Q branches, while the phase noise term  $\phi(t)$  models the phase difference between the carrier signal and the local oscillator. Based on this model, it can be shown that the received baseband signal  $y(t)$  is related to the transmitted baseband signal  $x(t)$  by (see [18] for a derivation):

$$\begin{aligned} y(t) &= \left[ \cos\left(\frac{\theta}{2}\right) - j\alpha \sin\left(\frac{\theta}{2}\right) \right] e^{j\phi(t)} y_0(t) \\ &\quad + \left[ \alpha \cos\left(\frac{\theta}{2}\right) + j \sin\left(\frac{\theta}{2}\right) \right] e^{-j\phi(t)} y_0^*(t) + w(t), \end{aligned} \quad (4)$$

where  $y_0(t)$  is given by (3). Letting

$$\mu = \cos\left(\frac{\theta}{2}\right) - j\alpha \sin\left(\frac{\theta}{2}\right), \quad \nu = \alpha \cos\left(\frac{\theta}{2}\right) + j \sin\left(\frac{\theta}{2}\right),$$

expression (4) can be represented as

$$y(t) = \mu e^{j\phi(t)} y_0(t) + \nu e^{-j\phi(t)} y_0^*(t) + w(t). \quad (5)$$

In the absence of IQ imbalance and phase noise, i.e., when  $\alpha = 0$ ,  $\theta = 0$ , and  $\phi(t) = 0$ , we have the traditional relation

$$y(t) = y_0(t) + w(t).$$

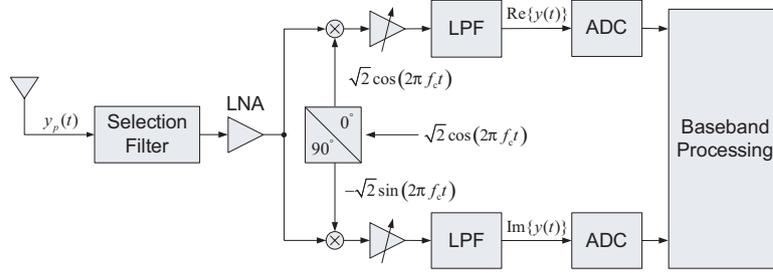
### B. OFDM Modulation and Demodulation

At the OFDM transmitter, the bits from information sources are first mapped into constellation symbols, and then converted into a block of  $N$  symbols  $X[k]$ ,  $k = 0, 1, \dots, N-1$ , by a serial-to-parallel converter. The  $N$  symbols are the frequency components to be transmitted using the  $N$  subcarriers of the OFDM modulator, and are converted to OFDM symbols  $x[n]$ ,  $n = 0, 1, \dots, N-1$ , by the unitary inverse Fast Fourier Transform (IFFT), i.e.,

$$x[n] = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X[k] e^{j\frac{2\pi nk}{N}}, \quad n = 0, 1, \dots, N-1.$$

A cyclic prefix of length  $P$  ( $P \leq N$ ) is added to the IFFT output in order to eliminate the inter-symbol interference caused by multipath propagation. The resulting  $N+P$  symbols are converted into a baseband signal  $x(t)$  for transmission. Let  $T_s$  be the sampling time of the system. At the demodulator, the received baseband signal  $y(t)$  is sampled at period  $T_s$ . After removing the cyclic prefix, a block of  $N$  symbols  $y[n]$ ,  $n = 0, 1, \dots, N-1$ , is obtained, whose elements are related to  $x[n]$ ,  $n = 0, 1, \dots, N-1$ , through (5) by

$$\begin{aligned} y[n] &= y(nT_s) \\ &= \mu e^{j\phi(nT_s)} y_0(nT_s) + \nu e^{-j\phi(nT_s)} y_0^*(nT_s) \\ &\quad + w(nT_s). \end{aligned} \quad (6)$$



(a) A direct-conversion receiver with perfect oscillator signals.

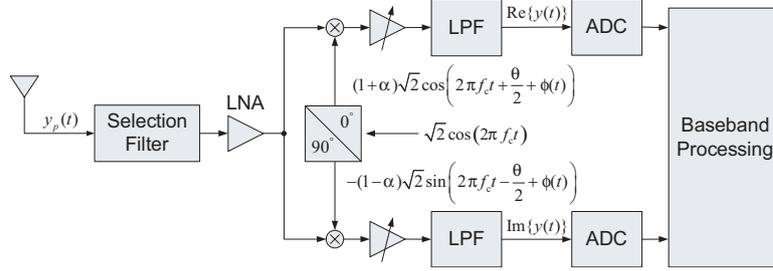
(b) A direct-conversion receiver with IQ imbalance and phase noise. The constants  $\alpha$  and  $\theta$  model the amplitude and phase imbalances between the I and Q branches, while  $\phi(t)$  models the phase noise.

Fig. 1. Block diagrams of a direct-conversion receiver with and without IQ imbalance and phase noise.

In the continuous-time domain,  $y_0(t)$  is equal to the convolutional integral of the channel impulse response and the channel input, as shown in expression (3). Let  $h[n]$  represent the equivalent discrete-time baseband channel impulse response. Assume that  $h[n]$  has length  $L$ , i.e.,  $h[n] = 0$  if  $n \notin \{0, 1, \dots, L-1\}$ . With the aid of the cyclic prefix and provided that  $L-1 \leq P$ , linear convolution becomes circular convolution in the discrete-time domain, i.e.,

$$y_0(nT_s) = h[n] \circledast x[n] = \sum_{r=0}^{N-1} h[(n-r)_N]x[r],$$

where  $\circledast$  denotes circular convolution and  $(n-r)_N$  stands for  $((n-r) \bmod N)$ . Expression (6) can then be written as

$$y[n] = \mu e^{j\phi(nT_s)} (h[n] \circledast x[n]) + \nu e^{-j\phi(nT_s)} (h[n] \circledast x[n])^* + w[n], \quad (7)$$

where  $w[n]$  is additive Gaussian noise. The unitary Fast Fourier Transform (FFT) is then performed on  $y[n]$ ,  $n = 0, 1, \dots, N-1$ , to obtain  $Y[k]$ ,  $k = 0, 1, \dots, N-1$ . Let

$$A[k] = \frac{1}{N} \sum_{n=0}^{N-1} e^{j\phi(nT_s)} e^{-j\frac{2\pi kn}{N}}, \quad k = 0, 1, \dots, N-1, \quad (8)$$

and

$$H[k] = \sum_{n=0}^{L-1} h[n] e^{-j\frac{2\pi kn}{N}}, \quad k = 0, 1, \dots, N-1.$$

Also, we denote

$$Z[k] = A[k] \circledast (H[k]X[k]) = \sum_{r=0}^{N-1} A[(k-r)_N]H[r]X[r]. \quad (9)$$

It then follows from (7) that the output symbols  $Y[k]$ ,  $k = 0, 1, \dots, N-1$ , after OFDM demodulation are related to the

data symbols  $X[k]$  by

$$\begin{aligned} Y[k] &= \mu Z[k] + \nu Z^*[(N-k)_N] + W[k] \\ &= \mu \sum_{r=0}^{N-1} A[(k-r)_N]H[r]X[r] \\ &\quad + \nu \sum_{r=0}^{N-1} A^*[(N-k-r)_N]H^*[r]X^*[r] + W[k] \end{aligned} \quad (10)$$

where  $W[k]$  is the additive noise in the  $k^{\text{th}}$  subcarrier and is given by the discrete Fourier transform of  $w[n]$ . Using matrix notation, (9) can be represented as

$$\mathbf{z} = \mathbf{A}\mathbf{H}\mathbf{x} \quad (11)$$

where

$$\begin{aligned} \mathbf{z} &= [Z[0] \ Z[1] \ \dots \ Z[N-1]]^T, \\ \mathbf{x} &= [X[0] \ X[1] \ \dots \ X[N-1]]^T, \\ \mathbf{A} &= \begin{bmatrix} A[0] & A[N-1] & \dots & A[1] \\ A[1] & A[0] & \dots & A[2] \\ \vdots & \vdots & \ddots & \vdots \\ A[N-1] & A[N-2] & \dots & A[0] \end{bmatrix}, \quad (12) \\ \mathbf{H} &= \begin{bmatrix} H[0] & 0 & \dots & 0 \\ 0 & H[1] & \dots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \dots & H[N-1] \end{bmatrix}. \end{aligned}$$

Then, expression (10) can be represented as

$$\mathbf{y} = \mu\mathbf{z} + \nu\tilde{\mathbf{z}} + \mathbf{w}, \quad (13)$$

where

$$\begin{aligned} \mathbf{y} &= [ Y[0] \ Y[1] \ \dots \ Y[N-1] ]^T, \\ \tilde{\mathbf{z}} &= [ Z^*[0] \ Z^*[N-1] \ \dots \ Z^*[1] ]^T, \\ \mathbf{w} &= [ W[0] \ W[1] \ \dots \ W[N-1] ]^T. \end{aligned}$$

Combining (11) and (13), we have

$$\mathbf{y} = \mu \mathbf{A} \mathbf{H} \mathbf{x} + \nu \tilde{\mathbf{A}} \cdot \text{conj}\{\mathbf{H}\} \cdot \text{conj}\{\mathbf{x}\} + \mathbf{w}, \quad (14)$$

where

$$\tilde{\mathbf{A}} = \begin{bmatrix} A^*[0] & A^*[N-1] & \dots & A^*[1] \\ A^*[N-1] & A^*[N-2] & \dots & A^*[0] \\ \vdots & \vdots & \ddots & \vdots \\ A^*[1] & A^*[0] & \dots & A^*[2] \end{bmatrix}.$$

Expression (10) and (14) formulate the effects of IQ imbalance and phase noise on the received symbols after OFDM demodulation. Based on this model, we shall develop a compensation scheme and analyze the system performance with and without compensation.

### III. CHANNEL ESTIMATION

#### A. Proposed Algorithm

One significant characteristic that distinguishes wireless communications from wireline communications is that wireless channels are time-varying. In OFDM systems, a training stage is required to estimate or track the channel response. In the presence of IQ imbalance and phase noise, the problem becomes more challenging because the received signals are altered not only by the channel but also by the physical impairments associated with the receiver. In the proposed channel estimation algorithm, block-type pilot symbols are transmitted, in which all subcarriers are used for the pilot symbols known to the receiver. For convenience of exposition, we assume that at each time, only *one* OFDM symbol is used as the block-type pilot symbol for channel estimation. Since the OFDM demodulation output  $\mathbf{y}$  is related to the training symbol  $\mathbf{x}$  through expression (14), the proposed algorithm is based on the following optimization problem:

$$\min_{\mu, \nu, \mathbf{A}, \mathbf{H}} \left\| \mathbf{y} - \mu \mathbf{A} \mathbf{H} \mathbf{x} - \nu \tilde{\mathbf{A}} \cdot \text{conj}\{\mathbf{H}\} \cdot \text{conj}\{\mathbf{x}\} \right\|^2. \quad (15)$$

We notice that there are  $N$  unknowns in  $\mathbf{H}$ ,  $N$  unknowns in  $\mathbf{A}$ , plus two additional unknowns  $\mu$  and  $\nu$ . Thus, the solution to this problem is not unique, since we have less observations (in  $\mathbf{y}$ ) than unknowns. To overcome this difficulty, we can reduce the number of unknowns by modeling the channel and the phase noise process with fewer parameters, as proposed in [16]. Since the length  $L$  of the discrete-time baseband channel impulse response is normally less than the OFDM symbol size  $N$ , we can relate  $H[k]$ ,  $k = 0, 1, \dots, N-1$ , to  $h[n]$ ,  $n = 0, 1, \dots, L-1$ , through

$$\mathbf{h} = \mathbf{F}_h \mathbf{h}', \quad (16)$$

where

$$\begin{aligned} \mathbf{h} &= [ H[0] \ H[1] \ \dots \ H[N-1] ]^T, \\ \mathbf{h}' &= [ h[0] \ h[1] \ \dots \ h[L-1] ]^T, \\ \mathbf{F}_h &= \begin{bmatrix} 1 & 1 & \dots & 1 \\ 1 & e^{-j\frac{2\pi}{N}} & \dots & e^{-j\frac{2\pi(N-L)}{N}} \\ \vdots & \vdots & \ddots & \vdots \\ 1 & e^{-j\frac{2\pi(N-1)}{N}} & \dots & e^{-j\frac{2\pi(N-1)(L-1)}{N}} \end{bmatrix}. \end{aligned}$$

Instead of estimating  $\mathbf{h}$ , we can estimate  $\mathbf{h}'$ . This reduces the number of unknown channel coefficients from  $N$  (in the frequency domain) to  $L$  (in the time domain). For the phase noise, instead of estimating  $A[k]$ ,  $k = 0, 1, \dots, N-1$ , we can estimate the phase noise components in the time domain, i.e.,  $e^{j\phi(nT_s)}$ ,  $n = 0, 1, \dots, N-1$ . In order to reduce the number of unknowns, we can estimate  $e^{j\phi(m(N-1)T_s/(M-1))}$  for  $m = 0, 1, \dots, M-1$  ( $M < N$ ), and then obtain the approximation of  $e^{j\phi(nT_s)}$ ,  $n = 0, 1, \dots, N-1$ , by interpolation. Let

$$\begin{aligned} \mathbf{c} &= [ e^{j\phi(0)} \ e^{j\phi(T_s)} \ \dots \ e^{j\phi((N-1)T_s)} ]^T, \\ \mathbf{c}' &= [ e^{j\phi(0)} \ e^{j\phi(\frac{(N-1)T_s}{M-1})} \ \dots \ e^{j\phi((N-1)T_s)} ]^T. \end{aligned}$$

Then,

$$\mathbf{c} \approx \mathbf{P} \mathbf{c}', \quad (17)$$

where  $\mathbf{P}$  is an interpolation matrix.<sup>2</sup> Using (8), we have

$$\mathbf{a} = \frac{1}{N} \mathbf{F}_a \mathbf{c} \approx \frac{1}{N} \mathbf{F}_a \mathbf{P} \mathbf{c}', \quad (18)$$

where

$$\begin{aligned} \mathbf{a} &= [ A[0] \ A[1] \ \dots \ A[N-1] ]^T, \\ \mathbf{F}_a &= \begin{bmatrix} 1 & 1 & \dots & 1 \\ 1 & e^{-j\frac{2\pi}{N}} & \dots & e^{-j\frac{2\pi(N-1)}{N}} \\ \vdots & \vdots & \ddots & \vdots \\ 1 & e^{-j\frac{2\pi(N-1)}{N}} & \dots & e^{-j\frac{2\pi(N-1)^2}{N}} \end{bmatrix}. \end{aligned}$$

Moreover, we realize that in (15)

- (1) there exists an ambiguity of a scaling factor in the estimates of  $\mu$ ,  $\mathbf{A}$  and  $\mathbf{H}$ ;
- (2) there exists an ambiguity of a scaling factor in the estimates of  $\nu$ ,  $\tilde{\mathbf{A}}$  and  $\text{conj}\{\mathbf{H}\}$ .

<sup>2</sup>If the power spectral density (PSD) of the phase noise is unknown,  $\mathbf{P}$  can be constructed from linear interpolation. That is, its element at the  $n^{\text{th}}$  row and  $m^{\text{th}}$  column is given by the equation at the top of the next page, where  $n = 1, 2, \dots, N$  and  $m = 1, 2, \dots, M$ . If the PSD of the phase noise is known, the optimal interpolation matrix  $\mathbf{P}_o$  can be obtained by minimizing the mean-square error of interpolating  $\mathbf{c}$  from  $\mathbf{c}'$ , i.e.,

$$\mathbf{P}_o = \arg \min_{\mathbf{P}} \mathbf{E} \|\mathbf{c} - \mathbf{P} \mathbf{c}'\|^2,$$

from which the optimal  $\mathbf{P}_o$  is given by

$$\mathbf{P}_o = \mathbf{R}_{\mathbf{c}\mathbf{c}'} \mathbf{R}_{\mathbf{c}'}^{-1}$$

where  $\mathbf{R}_{\mathbf{c}\mathbf{c}'} = \mathbf{E}\{\mathbf{c}\mathbf{c}'^*\}$  and  $\mathbf{R}_{\mathbf{c}'} = \mathbf{E}\{\mathbf{c}'\mathbf{c}'^*\}$ . Throughout this paper, we assume that the receiver has no information about the phase noise spectrum characteristics. The interpolator constructed from linear interpolation is convenient because it does not require any information about the phase noise spectrum.

$$\mathbf{P}(n, m) = \begin{cases} m - \frac{(n-1)(M-1)}{N-1}, & \text{if } \frac{(m-1)(N-1)}{M-1} \leq n-1 < \frac{m(N-1)}{M-1}, \\ \frac{(n-1)(M-1)}{N-1} - (m-2), & \text{if } \frac{(m-2)(N-1)}{M-1} \leq n-1 < \frac{(m-1)(N-1)}{M-1}, \\ 0, & \text{otherwise,} \end{cases}$$

To resolve the ambiguities, instead of estimating  $\mu$ ,  $\nu$ ,  $\mathbf{A}$  and  $\mathbf{H}$ , we estimate the following quantities:

$$\nu'' = \frac{\nu}{\mu}, \quad \mathbf{A}'' = \frac{1}{A[0]}\mathbf{A}, \quad \mathbf{H}'' = \mu A[0]\mathbf{H}, \quad (19)$$

where

$$\mathbf{A}'' = \begin{bmatrix} A''[0] & A''[N-1] & \dots & A''[1] \\ A''[1] & A''[0] & \dots & A''[2] \\ \vdots & \vdots & \ddots & \vdots \\ A''[N-1] & A''[N-2] & \dots & A''[0] \end{bmatrix} \quad (20)$$

with

$$A''[0] = 1 \text{ and } A''[k] = \frac{1}{A[0]}A[k] \text{ for } k = 1, 2, \dots, N-1,$$

and

$$\mathbf{H}'' = \begin{bmatrix} H''[0] & 0 & \dots & 0 \\ 0 & H''[1] & \dots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \dots & H''[N-1] \end{bmatrix} \quad (21)$$

with

$$H''[k] = \mu A[0]H[k] \text{ for } k = 0, 1, \dots, N-1.$$

With (19), expression (14) can be rewritten as

$$\mathbf{y} = \mathbf{A}''\mathbf{H}''\mathbf{x} + \nu''\widetilde{\mathbf{A}}'' \cdot \text{conj}\{\mathbf{H}''\} \cdot \text{conj}\{\mathbf{x}\} + \mathbf{w},$$

where

$$\widetilde{\mathbf{A}}'' = \begin{bmatrix} (A''[0])^* & (A''[N-1])^* & \dots & (A''[1])^* \\ (A''[N-1])^* & (A''[N-2])^* & \dots & (A''[0])^* \\ \vdots & \vdots & \ddots & \vdots \\ (A''[1])^* & (A''[0])^* & \dots & (A''[2])^* \end{bmatrix}. \quad (22)$$

Correspondingly, we define

$$\mathbf{c}'' = \frac{1}{A[0]}\mathbf{c}' \text{ and } \mathbf{h}'' = \mu A[0]\mathbf{h}'.$$

It follows from (18) and (16) that

$$\bar{\mathbf{a}} \approx \frac{1}{N}\mathbf{F}_a\mathbf{P}\mathbf{c}'' \text{ and } \bar{\mathbf{h}} = \mathbf{F}_h\mathbf{h}'' \quad (23)$$

where

$$\bar{\mathbf{a}} = [A''[0] \quad A''[1] \quad \dots \quad A''[N-1]]^T, \\ \bar{\mathbf{h}} = [H''[0] \quad H''[1] \quad \dots \quad H''[N-1]]^T.$$

Note that  $A''[0] = 1$ . Consequently, knowing  $\mathbf{x}$  and  $\mathbf{y}$ , we can estimate  $\mathbf{H}''$  by solving

$$\min_{\nu'', \mathbf{c}'', \mathbf{h}''} \left\| \mathbf{y} - \mathbf{A}''\mathbf{H}''\mathbf{x} - \nu''\widetilde{\mathbf{A}}'' \cdot \text{conj}\{\mathbf{H}''\} \cdot \text{conj}\{\mathbf{x}\} \right\|^2 \\ \text{subject to } A''[0] = 1. \quad (24)$$

This optimization problem (24) is nonlinear and nonconvex. An iterative method for finding a sub-optimal solution is

presented in Algorithm 1, in which we improve the estimates of  $\nu''$ ,  $\mathbf{c}''$  and  $\mathbf{h}''$  recursively by allowing small perturbations in them and then finding the optimal values for these perturbation terms. It can also be viewed as local linearization of a nonlinear system by using a first-order approximation. Since the amplitude of IQ imbalances and phase noise is usually small, the true values of  $\nu''$  and  $\mathbf{c}''$  are close to their nominal values  $\hat{\nu}''_0 = 0$  and  $\hat{\mathbf{c}}''_0 = [1 \ 1 \ \dots \ 1]^T$ . It is shown by computer simulations that the objective function decreases with  $i = 1, 2, \dots$ , and eventually converges to a local minimum. The obtained estimates of  $\nu''$  and  $\mathbf{H}''$  will be used in the data transmission stage for estimating data symbols.

### B. Cramer-Rao Lower Bound (CRLB) for Channel Estimation

To evaluate the proposed algorithm, we compare its performance with the Cramer-Rao lower bound (CRLB) that gives a lower bound on the covariance matrix of any unbiased estimator of unknown parameters [20]. Three scenarios are considered here and they can be either exactly or approximately modeled by

$$\mathbf{y} = \mathbf{s}_\theta + \mathbf{w},$$

where  $\mathbf{y}$  is the observed data vector of length  $N$ ,  $\mathbf{s}_\theta$  is the noise-free data vector that depends on the parameter vector  $\boldsymbol{\theta}$ , and  $\mathbf{w}$  is the vector of circularly symmetric Gaussian noise with covariance matrix  $\mathbf{E}\{\mathbf{w}\mathbf{w}^*\} = \sigma_w^2\mathbf{I}_N$ . The Fisher information matrix for this data model is given by [20]

$$\mathbf{I}_\theta = \frac{2}{\sigma_w^2} \sum_{k=0}^{N-1} \text{Re} \left\{ \frac{\partial S_\theta[k]}{\partial \boldsymbol{\theta}} \left( \frac{\partial S_\theta[k]}{\partial \boldsymbol{\theta}} \right)^* \right\}, \quad (26)$$

where  $\frac{\partial S_\theta[k]}{\partial \boldsymbol{\theta}} = \left[ \frac{\partial S_\theta[k]}{\partial \theta_1} \quad \frac{\partial S_\theta[k]}{\partial \theta_2} \quad \dots \quad \frac{\partial S_\theta[k]}{\partial \theta_{|\boldsymbol{\theta}|}} \right]^T$  ( $|\boldsymbol{\theta}|$  is the dimension of  $\boldsymbol{\theta}$ ). By the CRLB, any unbiased estimator  $\hat{\boldsymbol{\theta}}$  of  $\boldsymbol{\theta}$  has a covariance matrix that satisfies

$$\text{var}\{\hat{\boldsymbol{\theta}}\} = \mathbf{E}\{(\hat{\boldsymbol{\theta}} - \boldsymbol{\theta})(\hat{\boldsymbol{\theta}} - \boldsymbol{\theta})^*\} \geq \mathbf{I}_\theta^{-1}, \quad (27)$$

where  $\text{var}\{\hat{\boldsymbol{\theta}}\} \geq \mathbf{I}_\theta^{-1}$  is interpreted as meaning that the matrix  $\text{var}\{\hat{\boldsymbol{\theta}}\} - \mathbf{I}_\theta^{-1}$  is positive semidefinite. In the following derivation, we assume: 1) The pilot symbols  $X[k]$  are independent and identically distributed with zero mean. They also have the same power and let  $\sigma_P^2 = |X[k]|^2$ . 2) The pilot symbols, the phase noise, the channel coefficients and the additive noise are independent of each other. 3) The channel coefficients  $H[k]$  are circularly symmetric Gaussian distributed with mean zero and variance  $\sigma_H^2 = \mathbf{E}\{|H[k]|^2\}$ .

#### Scenario 1: No Impairment

In this scenario, there is no analog impairment in the system. We consider two cases: one estimates  $\mathbf{h}$  and the other estimates  $\mathbf{h}'$ . If  $\mathbf{h}$  is estimated, the system is modeled as

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{w},$$

**Algorithm 1** Joint Channel Estimation

0: Let  $\hat{\nu}'' = 0$  and  $\hat{\mathbf{c}}''_0 = [1 \ 1 \ \dots \ 1]^T$ . Find the initial  $\hat{\mathbf{h}}''_0$  by solving

$$\hat{\mathbf{h}}''_0 = \arg \min_{\mathbf{h}''} \|\mathbf{y} - \mathbf{H}'' \mathbf{x}\|^2.$$

The expression for  $\hat{\mathbf{h}}''_0$  is given by

$$\hat{\mathbf{h}}''_0 = (\mathbf{F}_h^* \mathbf{X}^* \mathbf{X} \mathbf{F}_h)^{-1} \mathbf{F}_h^* \mathbf{X}^* \mathbf{y},$$

where  $\mathbf{X} = \text{diag}\{\mathbf{x}\}$ .

1:  $i = 1$

2: **repeat**

3: Let  $\bar{\mathbf{a}}_{i-1} = \frac{1}{N} \mathbf{F}_a \mathbf{P} \hat{\mathbf{c}}''_{i-1}$  and  $\bar{\mathbf{h}}_{i-1} = \mathbf{F}_h \hat{\mathbf{h}}''_{i-1}$ .

4: Find  $\Delta \nu''_i$ ,  $\Delta \mathbf{c}''_i$  and  $\Delta \mathbf{h}''_i$  by solving the following optimization problem:

$$\begin{aligned} \min_{\Delta \nu'', \Delta \mathbf{c}'', \Delta \mathbf{h}''} & \left\| \mathbf{y} - (\mathbf{A}''_{i-1} \mathbf{H}''_{i-1} \mathbf{x} + \hat{\nu}''_{i-1} \tilde{\mathbf{A}}''_{i-1} \right. & (25) \\ & \cdot \text{conj}\{\mathbf{H}''_{i-1}\} \cdot \text{conj}\{\mathbf{x}\}) - [(\Delta \mathbf{A}'') \mathbf{H}''_{i-1} \mathbf{x} + \mathbf{A}''_{i-1} (\Delta \mathbf{H}'') \mathbf{x} \\ & - [(\Delta \nu'') \tilde{\mathbf{A}}''_{i-1} \cdot \text{conj}\{\mathbf{H}''_{i-1}\} \cdot \text{conj}\{\mathbf{x}\} + \hat{\nu}''_{i-1} (\Delta \tilde{\mathbf{A}}'') \\ & \cdot \text{conj}\{\mathbf{H}''_{i-1}\} \cdot \text{conj}\{\mathbf{x}\} + \hat{\nu}''_{i-1} \tilde{\mathbf{A}}''_{i-1} \cdot \text{conj}\{\Delta \mathbf{H}''\} \cdot \text{conj}\{\mathbf{x}\}] \left\|^2 \right. \\ & \text{subject to } \mathbf{g} \Delta \mathbf{c}'' = 0, \end{aligned}$$

where  $\mathbf{g}$  is the first row of  $\frac{1}{N} \mathbf{F}_a \mathbf{P}$ , and  $\mathbf{A}''_{i-1}$ ,  $\tilde{\mathbf{A}}''_{i-1}$ ,  $\mathbf{H}''_{i-1}$  are determined from  $\bar{\mathbf{a}}_{i-1}$  and  $\bar{\mathbf{h}}_{i-1}$  according to (20), (22) and (21). The constraint  $\mathbf{g} \Delta \mathbf{c}'' = 0$  guarantees that  $A''[0]$  is equal to 1. Problem (25) can be formulated as a standard least-squares problem of the following form (see [18] for more details):

$$\min_{\bar{\mathbf{x}}} \|\bar{\mathbf{y}} - \bar{\mathbf{A}} \bar{\mathbf{x}}\|^2.$$

Here,

$$\bar{\mathbf{x}} = \begin{bmatrix} \text{Re}\{\Delta \nu''\} & \text{Im}\{\Delta \nu''\} & \text{Re}\{\Delta \mathbf{c}''\}^T & \text{Im}\{\Delta \mathbf{c}''\}^T \\ \text{Re}\{\Delta \mathbf{h}''\}^T & \text{Im}\{\Delta \mathbf{h}''\}^T & & \end{bmatrix}^T,$$

where the vector  $\Delta \mathbf{c}'' = [\Delta c''[1] \ \Delta c''[2] \ \dots \ \Delta c''[M-1]]^T$  contains all elements of  $\Delta \mathbf{c}''$  except its first element,  $\Delta c''[0]$ . The matrix  $\bar{\mathbf{A}}$  and vector  $\bar{\mathbf{y}}$  are formed according to (25), and the constraint  $\mathbf{g} \Delta \mathbf{c}'' = 0$  eliminates  $\Delta c''[0]$  from the optimization parameters. The optimal solution is given by

$$\bar{\mathbf{x}}_o = (\bar{\mathbf{A}}^T \bar{\mathbf{A}})^{-1} \bar{\mathbf{A}}^T \bar{\mathbf{y}}.$$

5: Update the estimates of  $\nu''$ ,  $\mathbf{c}''$  and  $\mathbf{h}''$  according to

$$\hat{\nu}''_i = \hat{\nu}''_{i-1} + \Delta \nu''_i, \quad \hat{\mathbf{c}}''_i = \hat{\mathbf{c}}''_{i-1} + \Delta \mathbf{c}''_i, \quad \hat{\mathbf{h}}''_i = \hat{\mathbf{h}}''_{i-1} + \Delta \mathbf{h}''_i.$$

6:  $i = i + 1$

7: **until** there is no significant improvement in the objective function  $\|\mathbf{y} - \mathbf{A}''_i \mathbf{H}''_i \mathbf{x} - \hat{\nu}''_i \tilde{\mathbf{A}}''_i \cdot \text{conj}\{\mathbf{H}''_i\} \cdot \text{conj}\{\mathbf{x}\}\|^2$ .

where  $\mathbf{s}_\theta = \mathbf{H} \mathbf{x}$  and

$$\boldsymbol{\theta} = \begin{bmatrix} \text{Re}\{H[0]\} & \dots & \text{Re}\{H[N-1]\} & \text{Im}\{H[0]\} \\ \dots & \dots & \text{Im}\{H[N-1]\} & \end{bmatrix}^T.$$

By (26) and (27), we have

$$\mathbf{E}\{|\hat{H}[k] - H[k]|^2\} \geq \frac{\sigma_W^2}{N \sigma_P^2}. \quad (28)$$

If  $\mathbf{h}'$  is estimated instead of  $\mathbf{h}$ , the system model becomes

$$\mathbf{y} = \mathbf{X} \mathbf{F}_h \mathbf{h}' + \mathbf{w},$$

where  $\mathbf{X} = \text{diag}\{\mathbf{x}\}$ . In this case,  $\mathbf{s}_\theta = \mathbf{X} \mathbf{F}_h \mathbf{h}'$ . Using (26) and (27), we have

$$\mathbf{E}\{|\hat{h}[n] - h[n]|^2\} \geq \frac{\sigma_W^2}{N \sigma_P^2},$$

from which it follows immediately that

$$\mathbf{E}\{|\hat{H}[k] - H[k]|^2\} \geq \frac{L \sigma_W^2}{N \sigma_P^2}. \quad (29)$$

*Scenario 2: Without any Compensation when Both IQ Imbalance and Phase Noise are Present*

In the presence of the impairments, the system model (14) can be rewritten as

$$\begin{aligned} \mathbf{y} &= \mu \mathbf{A} \mathbf{H} \mathbf{x} + \nu \tilde{\mathbf{A}} \cdot \text{conj}\{\mathbf{H}\} \cdot \text{conj}\{\mathbf{x}\} + \mathbf{w} \\ &= \mu A[0] \mathbf{H} \mathbf{x} + \mu (\mathbf{A} - A[0] \mathbf{I}_N) \mathbf{H} \mathbf{x} \\ &\quad + \nu \tilde{\mathbf{A}} \cdot \text{conj}\{\mathbf{H}\} \cdot \text{conj}\{\mathbf{x}\} + \mathbf{w}. \end{aligned}$$

We treat  $\mathbf{H}'' = \mu A[0] \mathbf{H}$  as the ‘‘true’’ channel response to be estimated because of the scalar ambiguity. The term  $\mu (\mathbf{A} - A[0] \mathbf{I}_N) \mathbf{H} \mathbf{x} + \nu \tilde{\mathbf{A}} \cdot \text{conj}\{\mathbf{H}\} \cdot \text{conj}\{\mathbf{x}\} + \mathbf{w}$  can be *approximately* regarded as additive white Gaussian noise with covariance matrix  $\{[(1 - \sigma_{A,0}^2) |\mu|^2 + |\nu|^2] \sigma_H^2 \sigma_P^2 + \sigma_W^2\} \cdot \mathbf{I}_N$ , where  $\sigma_{A,0}^2 = \mathbf{E}\{|A[0]|^2\}$ . The CRLB can then be computed as

$$\begin{aligned} \mathbf{E}\{|\mu A[0] \hat{H}[k] - \mu A[0] H[k]|^2\} \\ \geq [(1 - \sigma_{A,0}^2) |\mu|^2 + |\nu|^2] \sigma_H^2 + \frac{\sigma_W^2}{\sigma_P^2}. \end{aligned} \quad (30)$$

Similarly, if  $\mathbf{h}'$  is estimated, then

$$\begin{aligned} \mathbf{E}\{|\mu A[0] \hat{H}[k] - \mu A[0] H[k]|^2\} \\ \geq \frac{L}{N} [(1 - \sigma_{A,0}^2) |\mu|^2 + |\nu|^2] \sigma_H^2 + \frac{L \sigma_W^2}{N \sigma_P^2}. \end{aligned} \quad (31)$$

*Scenario 3: With the Proposed Joint Estimation when Both IQ Imbalance and Phase Noise are Present*

In this case, the CRLB for estimating  $\mathbf{H}$  is computed based on the following model:

$$\begin{aligned} \mathbf{y} &= \mathbf{A}'' \mathbf{H}'' \mathbf{x} + \nu'' \tilde{\mathbf{A}}'' \cdot \text{conj}\{\mathbf{H}''\} \cdot \text{conj}\{\mathbf{x}\} + \mathbf{w} \\ &= \mathbf{A}''_{\text{appro}} \mathbf{H}'' \mathbf{x} + \nu'' \tilde{\mathbf{A}}''_{\text{appro}} \cdot \text{conj}\{\mathbf{H}''\} \cdot \text{conj}\{\mathbf{x}\} \\ &\quad + (\mathbf{A}'' - \mathbf{A}''_{\text{appro}}) \mathbf{H}'' \mathbf{x} + \nu'' (\tilde{\mathbf{A}}'' - \tilde{\mathbf{A}}''_{\text{appro}}) \\ &\quad \cdot \text{conj}\{\mathbf{H}''\} \cdot \text{conj}\{\mathbf{x}\} + \mathbf{w}, \end{aligned} \quad (32)$$

where  $\mathbf{A}''_{\text{appro}}$  is determined by the vector  $\bar{\mathbf{a}}_{\text{appro}} = \frac{1}{N} \mathbf{F}_a \mathbf{P} \mathbf{c}''$  according to the construction of  $\mathbf{A}''$ . Note that  $\mathbf{A}'' - \mathbf{A}''_{\text{appro}}$  represents the modeling error existing in the approximation given by (23). The parameter vector to be estimated is

$$\boldsymbol{\theta} = \begin{bmatrix} \text{Re}\{\nu''\} & \text{Im}\{\nu''\} & \text{Re}\{\mathbf{c}''\}^T & \text{Im}\{\mathbf{c}''\}^T \\ \text{Re}\{\mathbf{h}''\}^T & \text{Im}\{\mathbf{h}''\}^T & & \end{bmatrix}^T,$$

where the vector  $\mathbf{c}'' = [c''[1] \ c''[2] \ \dots \ c''[M-1]]^T$  contains all elements of  $\mathbf{c}''$  except its first element  $c''[0]$ . Note that  $c''[0]$  is determined by  $c''[m]$ ,  $m = 1, 2, \dots, M-1$ , because of the constraint  $A''[0] = 1$ . Hence,  $\mathbf{s}_\theta = \mathbf{A}''_{\text{appro}} \mathbf{H}'' \mathbf{x} + \nu'' \tilde{\mathbf{A}}''_{\text{appro}} \cdot \text{conj}\{\mathbf{H}''\} \cdot \text{conj}\{\mathbf{x}\}$ , and the noise term in expression (32) is

$$\begin{aligned} \mathbf{w}'' &= (\mathbf{A}'' - \mathbf{A}''_{\text{appro}}) \mathbf{H}'' \mathbf{x} \\ &\quad + \nu'' (\tilde{\mathbf{A}}'' - \tilde{\mathbf{A}}''_{\text{appro}}) \cdot \text{conj}\{\mathbf{H}''\} \cdot \text{conj}\{\mathbf{x}\} + \mathbf{w}. \end{aligned}$$

The covariance matrix of  $\mathbf{w}''$  is *approximately* equal to  $\sigma_{W''}^2 \mathbf{I}_N$ , where

$$\begin{aligned} \sigma_{W''}^2 &= (1 + |\nu''|^2) \cdot \mathbf{E}\{\|\bar{\mathbf{a}} - \bar{\mathbf{a}}_{\text{appro}}\|^2 |H''[k]|^2\} \cdot \sigma_P^2 + \sigma_W^2 \\ &= (|\mu|^2 + |\nu|^2) \cdot \mathbf{E}\{\|\mathbf{a} - \mathbf{a}_{\text{appro}}\|^2\} \cdot \sigma_H^2 \sigma_P^2 + \sigma_W^2 \\ &\approx (|\mu|^2 + |\nu|^2) \\ &\quad \cdot \left(1 - \frac{1}{N} \text{Tr}\{\mathbf{P}(\mathbf{P}^* \mathbf{P})^{-1} \mathbf{P}^* \mathbf{R}_c\}\right) \sigma_H^2 \sigma_P^2 + \sigma_W^2. \end{aligned}$$

Here,  $\mathbf{R}_c = \mathbf{E}\{\mathbf{c}\mathbf{c}^*\}$  and  $\mathbf{E}\{\|\mathbf{a} - \mathbf{a}_{\text{appro}}\|^2\} = 1 - \frac{1}{N} \text{Tr}\{\mathbf{P}(\mathbf{P}^* \mathbf{P})^{-1} \mathbf{P}^* \mathbf{R}_c\}$  is given by the minimum mean-square error of approximating  $\mathbf{c}$  by  $\mathbf{P}\mathbf{c}'$  in (17).

Consequently,  $\mathbf{I}_\theta$  and the associated CRLB for  $\mathbf{h}''$  can be computed (see [18] for more details). In the computation, the covariance matrix  $\mathbf{R}_c$  of the phase noise vector  $\mathbf{c}$  depends on the phase noise spectral characteristics. Given a specific phase noise model,  $\mathbf{R}_c$  can be computed analytically [16]. By using the relation  $\bar{\mathbf{h}} = \mathbf{F}_h \mathbf{h}''$ , we have

$$\mathbf{E}\{|\mu A[0] \hat{H}[k] - \mu A[0] H[k]|^2\} = \mathbf{E}\{\|\hat{\mathbf{h}}'' - \mathbf{h}''\|^2\}. \quad (33)$$

The lower bound for  $H[k]$  can then be derived from the lower bound for  $\mathbf{h}''$ . It is noted that the estimation performance depends on  $M$ , and the CRLB allows us to select an appropriate  $M$ . A trade-off exists here, because a large  $M$  gives better interpolation performance but at the cost of the degree of freedom, while a small  $M$  reduces the number of unknowns but causes larger interpolation error [16]. In Section V, we compare the mean-square errors of channel estimation with the derived CRLB, and show that the CRLB is a good theoretical measure for the estimation accuracy.

#### IV. DATA SYMBOL ESTIMATION

##### A. Proposed Algorithm

Assume that the receiver has acquired the channel response  $\mathbf{H}''$  and the IQ imbalance parameter  $\nu''$ . Given the system model

$$\mathbf{y} = \mathbf{z}'' + \nu'' \widetilde{\mathbf{z}}'' + \mathbf{w}$$

where  $\mathbf{z}'' = \mathbf{A}'' \mathbf{H}'' \mathbf{x}$ ,

$$\begin{aligned} \mathbf{z}'' &= [Z''[0] \quad Z''[1] \quad \dots \quad Z''[N-1]]^T, \\ \widetilde{\mathbf{z}}'' &= [(Z''[0])^* \quad (Z''[N-1])^* \quad \dots \quad (Z''[1])^*]^T, \end{aligned}$$

we are now interested in estimating the transmitted vector  $\mathbf{x}$ . By inspecting the model, it is noticed that the problem can be decomposed into two separate compensation problems: IQ imbalance compensation and phase noise compensation, as illustrated in Fig. 2. First,  $\mathbf{z}''$  is estimated from  $\mathbf{y}$  by using any IQ imbalance compensation method; then,  $\mathbf{x}$  is estimated from  $\hat{\mathbf{z}}''$  by using any phase noise compensation method. Here, we apply the post-FFT IQ compensation technique developed in [10] and the phase noise compensation technique developed in [16]. To compensate for phase noise, we also assume that comb-type OFDM symbols are transmitted in the payload portion of each packet. In each comb-type symbol, some subcarriers are used for pilot symbols, while the others are used for data symbols. The two-stage algorithm is summarized in Algorithm 2.

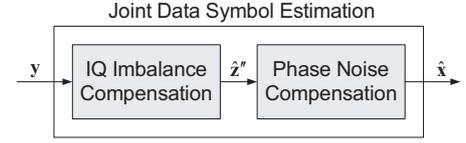


Fig. 2. Block diagram of the data symbol estimation algorithm. It can be decomposed into the IQ imbalance compensation block and the phase noise compensation block.

##### Algorithm 2 Data Symbol Estimation

1: Estimate  $\mathbf{z}''$  from  $\mathbf{y}$ . This can be done by

$$\hat{z}''[k] = \frac{Y[k] - \nu'' Y^*[(N-k)N]}{1 - |\nu''|^2}, \quad k = 0, 1, \dots, N-1.$$

2: Use the method given in [11] to estimate the common phase rotation  $A[0]$ . Assume that  $k_{\text{pilot},j}$ ,  $j = 1, 2, \dots, Q$ , are the subcarrier indices of the  $Q$  pilot tones. The common phase error coefficient  $A[0]$  is estimated by

$$\hat{A}[0] = \frac{\sum_{j=1}^Q (H''[k_{\text{pilot},j}])^* (X[k_{\text{pilot},j}])^* \hat{Z}''[k_{\text{pilot},j}]}{\sum_{j=1}^Q |H''[k_{\text{pilot},j}]|^2 |X[k_{\text{pilot},j}]|^2}.$$

3: Let  $\hat{\mathbf{c}}'_0 = [\hat{A}[0] \quad \hat{A}[0] \quad \dots \quad \hat{A}[0]]^T$ .

4:  $i = 1$

5: **repeat**

6: Let  $\hat{\mathbf{a}}_{i-1} = \frac{1}{N} \mathbf{F}_a \mathbf{P} \hat{\mathbf{c}}'_{i-1}$  and construct  $\mathbf{A}_{i-1}$  from  $\hat{\mathbf{a}}_{i-1}$  according to (12). Find the associated optimal  $\hat{\mathbf{x}}_{\text{data},i-1}$  by solving the following least-squares problem:

$$\begin{aligned} \hat{\mathbf{x}}_{\text{data},i-1} &= \arg \min_{\mathbf{x}_{\text{data}}} \|\hat{\mathbf{z}}'' - \mathbf{A}_{\text{pilot},i-1} \mathbf{H}''_{\text{pilot}} \mathbf{x}_{\text{pilot}} \\ &\quad - \mathbf{A}_{\text{data},i-1} \mathbf{H}''_{\text{data}} \mathbf{x}_{\text{data}}\|^2 \end{aligned} \quad (34)$$

where  $\mathbf{x}_{\text{pilot}}$  is the sub-vector of  $\mathbf{x}$  that consists of all the pilot symbols and  $\mathbf{A}_{\text{pilot},i-1}$  and  $\mathbf{H}''_{\text{pilot}}$  are the associated sub-matrices of  $\mathbf{A}_{i-1}$  and  $\mathbf{H}''$ . Moreover,  $\mathbf{x}_{\text{data}}$  is the sub-vector of  $\mathbf{x}$  that consists of all the data symbols and  $\mathbf{A}_{\text{data},i-1}$  and  $\mathbf{H}''_{\text{data}}$  are the associated sub-matrices of  $\mathbf{A}_{i-1}$  and  $\mathbf{H}''$ . The expression for  $\hat{\mathbf{x}}_{\text{data},i-1}$  is given by

$$\begin{aligned} \hat{\mathbf{x}}_{\text{data},i-1} &= (\mathbf{H}''_{\text{data}})^{-1} \left( \mathbf{A}_{\text{data},i-1}^* \mathbf{A}_{\text{data},i-1} \right)^{-1} \mathbf{A}_{\text{data},i-1}^* \\ &\quad \cdot \left( \hat{\mathbf{z}}'' - \mathbf{A}_{\text{pilot},i-1} \mathbf{H}''_{\text{pilot}} \mathbf{x}_{\text{pilot}} \right). \end{aligned}$$

7: Find the optimal  $\hat{\mathbf{c}}'_i$  by solving the following least-squares problem:

$$\begin{aligned} \hat{\mathbf{c}}'_i &= \arg \min_{\mathbf{c}'_i} \|\hat{\mathbf{z}}'' - \mathbf{A}_{\text{pilot}} \mathbf{H}''_{\text{pilot}} \mathbf{x}_{\text{pilot}} \\ &\quad - \mathbf{A}_{\text{data}} \mathbf{H}''_{\text{data}} \hat{\mathbf{x}}_{\text{data},i-1}\|^2 \end{aligned} \quad (35)$$

where  $\mathbf{A}_{\text{pilot}}$  and  $\mathbf{A}_{\text{data}}$  are determined from  $\mathbf{c}'$  according to (18) and (12). The expression for  $\hat{\mathbf{c}}'_i$  is given by

$$\hat{\mathbf{c}}'_i = N (\mathbf{P}^* \mathbf{F}_a^* \mathbf{T}^* \mathbf{T} \mathbf{F}_a \mathbf{P})^{-1} \mathbf{P}^* \mathbf{F}_a^* \mathbf{T}^* \hat{\mathbf{z}}'',$$

where  $\mathbf{T}$  is the circulant matrix formed by the elements of  $\mathbf{H}'' \hat{\mathbf{x}}_{i-1}$  with  $\hat{\mathbf{x}}_{i-1}$  formed by  $\mathbf{x}_{\text{pilot}}$  and  $\hat{\mathbf{x}}_{\text{data},i-1}$ .

8:  $i = i + 1$

9: **until** there is no significant improvement in the objective function  $\|\hat{\mathbf{z}}'' - \mathbf{A}_{\text{pilot},i} \mathbf{H}''_{\text{pilot}} \mathbf{x}_{\text{pilot}} - \mathbf{A}_{\text{data},i} \mathbf{H}''_{\text{data}} \hat{\mathbf{x}}_{\text{data},i-1}\|^2$ .

##### B. Performance Analysis

In this subsection, we analyze the effects of IQ imbalance and phase noise on OFDM systems in terms of the signal-to-noise ratio degradation. The expressions of the effective signal-to-noise ratio at the receiver are derived by assuming that 1) the data symbols  $X[k]$  are independent and identically distributed with mean zero and variance  $\sigma_X^2 = \mathbf{E}\{|X[k]|^2\}$ ; 2) the data symbols, the phase noise, the channel coefficients and the additive noise are independent of each other; 3) the channel

coefficients  $H[k]$  are independently identically distributed and circularly symmetric Gaussian with mean zero and variance  $\sigma_H^2 = \mathbf{E}\{|H[k]|^2\}$ .

#### Scenario 1: No Impairment

In this case, there is no impairment in the system. Since

$$Y[k] = H[k]X[k] + W[k],$$

the effective signal-to-noise ratio is then given by

$$\text{SNR}_0 = \frac{\mathbf{E}\{|H[k]X[k]|^2\}}{\mathbf{E}\{|W[k]|^2\}} = \frac{\sigma_H^2 \sigma_X^2}{\sigma_W^2}. \quad (36)$$

#### Scenario 2: No Compensation for Phase Noise and IQ Imbalance

In the presence of the IQ and phase noise impairments, the receiver does not perform any compensation. The system model (10) can be rewritten as

$$\begin{aligned} Y[k] &= H[k]X[k] + (\mu A[0] - 1)H[k]X[k] \\ &+ \mu \sum_{r=0, r \neq k}^{N-1} A[(k-r)_N]H[r]X[r] \\ &+ \nu \sum_{r=0}^{N-1} A^*[(N-k-r)_N]H^*[r]X^*[r] + W[k], \end{aligned}$$

where  $H[k]X[k]$  is the desired signal component and the other terms are regarded as additive noise. The effective signal-to-noise ratio is computed as

$$\begin{aligned} \text{SNR}_{\text{no}} &= \frac{\sigma_H^2 \sigma_X^2}{(1 - 2\text{Re}\{\mu A[0]\} + |\mu|^2 + |\nu|^2) \sigma_H^2 \sigma_X^2 + \sigma_W^2} \\ &= \frac{\text{SNR}_0}{(1 - 2\text{Re}\{\mu A[0]\} + |\mu|^2 + |\nu|^2) \text{SNR}_0 + 1}. \quad (37) \end{aligned}$$

#### Scenario 3: IQ and Common Phase Error (CPE) Compensation, i.e., $\mu$ , $\nu$ and $A[0]$ are Known

In this case, the IQ imbalance and the CPE term are compensated for at the receiver, as proposed in [17]. The system model (10) can now be rewritten as

$$\begin{aligned} Y[k] &= \mu A[0]H[k]X[k] \\ &+ \nu A^*[0]H^*[(N-k)_N]X^*[(N-k)_N] \\ &+ \mu \sum_{r=0, r \neq k}^{N-1} A[(k-r)_N] \cdot H[r]X[r] \\ &+ \nu \sum_{\substack{r=0, \\ r \neq (N-k)_N}}^{N-1} A^*[(N-k-r)_N]H^*[r]X^*[r] + W[k], \end{aligned}$$

where  $\mu A[0]H[k]X[k] + \nu A^*[0]H^*[(N-k)_N]X^*[(N-k)_N]$  is the desired signal component and the other terms are regarded as additive noise. The effective signal-to-noise ratio is given by

$$\begin{aligned} \text{SNR}_{\text{IQ+CPE}} &= \frac{(|\mu|^2 + |\nu|^2) \sigma_{A,0}^2 \sigma_H^2 \sigma_X^2}{(|\mu|^2 + |\nu|^2)(1 - \sigma_{A,0}^2) \sigma_H^2 \sigma_X^2 + \sigma_W^2} \\ &= \frac{(|\mu|^2 + |\nu|^2) \sigma_{A,0}^2 \text{SNR}_0}{(|\mu|^2 + |\nu|^2)(1 - \sigma_{A,0}^2) \text{SNR}_0 + 1}. \quad (38) \end{aligned}$$

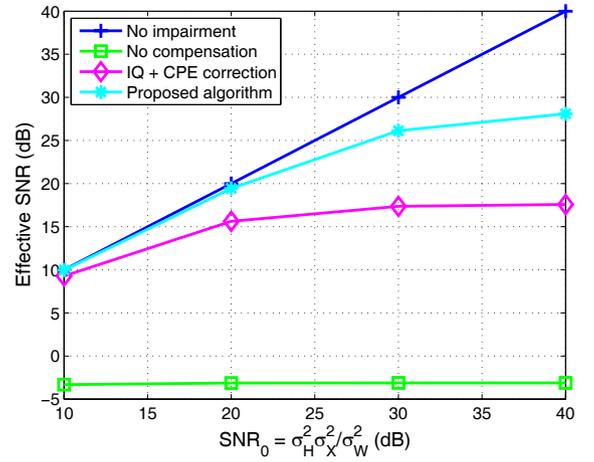


Fig. 3. Plots of the effective signal-to-noise ratio at the receiver by using (36)-(39) when  $\alpha = 0.1$ ,  $\theta = 10^\circ$  and  $\xi = 5$  kHz.

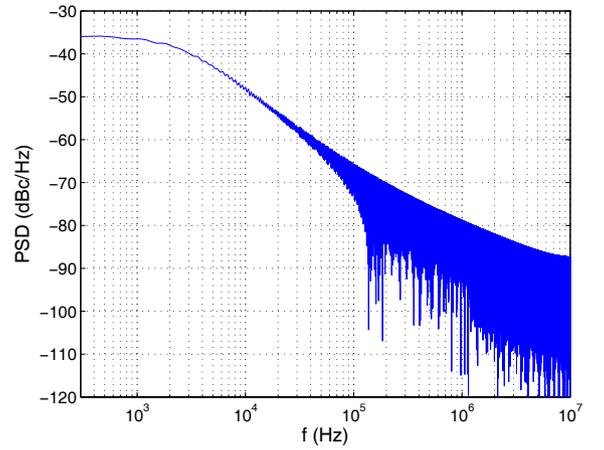


Fig. 4. Power spectral density (PSD) of the phase noise with  $\xi = 5$  kHz. The PSD is measured in dB with respect to the carrier power, namely, dBc.

#### Scenario 4: the Proposed Joint Compensation Scheme

With the proposed algorithm, we rewrite the system model (14) in the matrix form as

$$\begin{aligned} \mathbf{y} &= \mu \mathbf{A}_{\text{appro}} \mathbf{H} \mathbf{x} + \nu \tilde{\mathbf{A}}_{\text{appro}} \cdot \text{conj}\{\mathbf{H}\} \cdot \text{conj}\{\mathbf{x}\} \\ &+ \mu (\mathbf{A} - \mathbf{A}_{\text{appro}}) \mathbf{H} \mathbf{x} + \nu (\tilde{\mathbf{A}} - \tilde{\mathbf{A}}_{\text{appro}}) \\ &\cdot \text{conj}\{\mathbf{H}\} \cdot \text{conj}\{\mathbf{x}\} + \mathbf{w}, \end{aligned}$$

and hence the effective signal-to-noise ratio is given by (see the equation at the top of the next page).

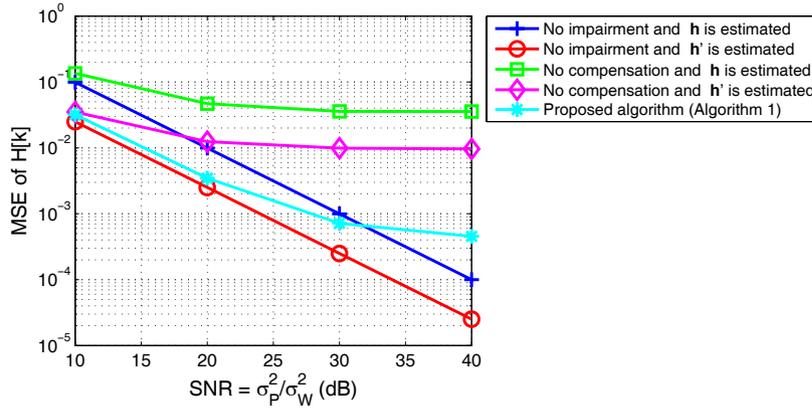
Fig. 3 plots the effective signal-to-noise ratio for different compensation scenarios by using expressions (36)-(39) when  $\alpha = 0.1$ ,  $\theta = 10^\circ$ , and the phase noise is generated by an oscillator according to the model given in [16] with linewidth  $\xi = 5$  kHz.

## V. COMPUTER SIMULATIONS

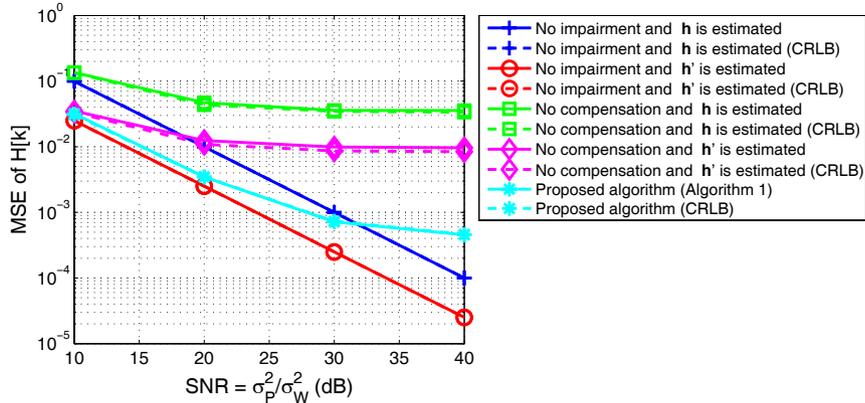
In the simulations, the system bandwidth is 20 MHz, i.e.,  $T_s = 0.05 \mu\text{s}$ , and the constellation used for symbol mapping is 16-QAM. The OFDM symbol size is  $N = 64$  and the prefix length is  $P = 16$ .<sup>3</sup> The channel length is 6, and each tap

<sup>3</sup>This is the same as in the IEEE 802.11a standard.

$$\begin{aligned}
\text{SNR}_{\text{prop}} &= \frac{\mathbf{E}\left\{\left\|\mu\mathbf{A}_{\text{appro}}\mathbf{H}\mathbf{x} + \nu\tilde{\mathbf{A}}_{\text{appro}} \cdot \text{conj}\{\mathbf{H}\} \cdot \text{conj}\{\mathbf{x}\}\right\|^2\right\}}{\mathbf{E}\left\{\left\|\mu(\mathbf{A} - \mathbf{A}_{\text{appro}})\mathbf{H}\mathbf{x} + \nu(\tilde{\mathbf{A}} - \tilde{\mathbf{A}}_{\text{appro}}) \cdot \text{conj}\{\mathbf{H}\} \cdot \text{conj}\{\mathbf{x}\} + \mathbf{w}\right\|^2\right\}} \\
&= \frac{N(|\mu|^2 + |\nu|^2) \cdot \mathbf{E}\{\|\mathbf{a}_{\text{appro}}\|^2\} \cdot \sigma_H^2 \sigma_X^2}{N(|\mu|^2 + |\nu|^2) \cdot \mathbf{E}\{\|\mathbf{a} - \mathbf{a}_{\text{appro}}\|^2\} \cdot \sigma_H^2 \sigma_X^2 + N\sigma_W^2} \\
&= \frac{(|\mu|^2 + |\nu|^2) \cdot \frac{1}{N} \text{Tr}\left\{\mathbf{P}(\mathbf{P}^*\mathbf{P})^{-1}\mathbf{P}^*\mathbf{R}_c\right\} \cdot \sigma_H^2 \sigma_X^2}{(|\mu|^2 + |\nu|^2) \cdot \left(1 - \frac{1}{N} \text{Tr}\left\{\mathbf{P}(\mathbf{P}^*\mathbf{P})^{-1}\mathbf{P}^*\mathbf{R}_c\right\}\right) \cdot \sigma_H^2 \sigma_X^2 + \sigma_W^2} \\
&\approx \frac{(|\mu|^2 + |\nu|^2) \cdot \frac{1}{N} \text{Tr}\left\{\mathbf{P}(\mathbf{P}^*\mathbf{P})^{-1}\mathbf{P}^*\mathbf{R}_c\right\} \cdot \text{SNR}_0}{(|\mu|^2 + |\nu|^2) \cdot \left(1 - \frac{1}{N} \text{Tr}\left\{\mathbf{P}(\mathbf{P}^*\mathbf{P})^{-1}\mathbf{P}^*\mathbf{R}_c\right\}\right) \cdot \text{SNR}_0 + 1} \quad (39)
\end{aligned}$$



(a) Mean-square error obtained by computer simulations.



(b) Mean-square error vs. CRLB computed by using the formulas derived in Subsection III-B.

Fig. 5. Plots of the MSE and CRLB for channel estimation when  $\alpha = 0.1$ ,  $\theta = 10^\circ$  and  $\xi = 5$  kHz. Five cases are simulated: i) There is no impairment and  $\mathbf{h}$  is estimated. ii) There is no impairment and  $\mathbf{h}'$  is estimated. iii) Both IQ imbalance and phase noise are present but the receiver assumes no impairment when estimating  $\mathbf{h}$ . iv) Both IQ imbalance and phase noise are present but the receiver assumes no impairment when estimating  $\mathbf{h}'$ . v) Both IQ imbalance and phase noise are present and the proposed channel estimation algorithm is applied.

is independently Rayleigh distributed with the power profile specified by 6 dB decay per tap. The average power of the channel response is normalized to 1, i.e.,  $\sigma_H^2 = 1$ . We simulate an OFDM receiver with the IQ imbalance specified by  $\alpha = 0.1$  and  $\theta = 10^\circ$ . The phase noise is generated according to the model given in [16] with linewidth  $\xi = 5$  kHz, and its spectrum is shown in Fig. 4.

We first examine the performance of different channel estimation algorithms for different scenarios. In the simulations, only one block-type pilot symbol is used for each time of

channel estimation. The assumed channel length in the time domain is  $L = 16$  and the length of the phase noise vector to be estimated is  $M = 8$ .<sup>4</sup> Fig. 5(a) plots the mean-square errors (MSE) of different channel estimation algorithms vs. the normalized signal-to-noise ratio at the receiver, i.e.,  $\text{SNR} = \sigma_P^2/\sigma_W^2$ . It is shown that estimating  $\mathbf{h}'$  rather than  $\mathbf{h}$  can improve the accuracy in terms of MSE by a factor of  $L/N = 16/64 = -6.02$  dB. The proposed joint channel

<sup>4</sup> $L$  is the assumed maximum channel length and equal to the cyclic prefix length  $P$ .

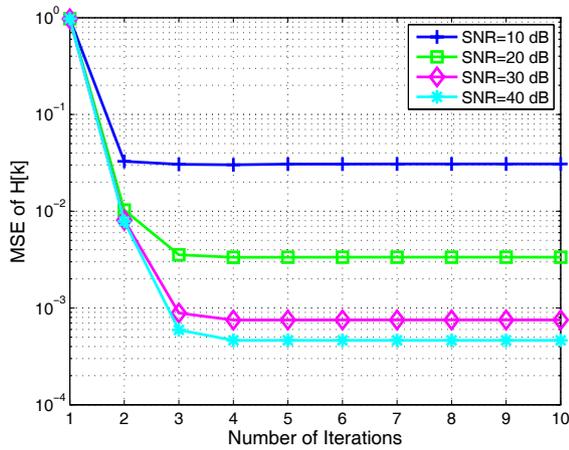


Fig. 6. Plots of the MSE of channel estimation vs. the number of iterations when  $\alpha = 0.1$ ,  $\theta = 10^\circ$  and  $\xi = 5$  kHz.

estimation algorithm performs better than the conventional methods that simply treat the impairments as additive noise. In Fig. 5(b), the CRLB is plotted in dotted lines by using the expressions (28)-(31) and (33). By comparing Fig. 5(a) and Fig. 5(b), it can be seen that the CRLB gives a good measure about the accuracy of different algorithms. Moreover, Fig. 6 demonstrates that the proposed channel estimation algorithm requires about 10 iterations to ensure convergence.

The proposed data symbol estimation algorithm is simulated in comparison with the ideal OFDM receiver with no impairment and the IQ+CPE (common phase error) correction scheme proposed in [17]. During the payload portion of OFDM packets, 16 out of the 64 subcarriers are used for pilot tones, i.e.,  $Q = 16$ . Fig. 7 shows the uncoded bit error rate (BER) performance of the system when the receiver has the *perfect* channel information, while Fig. 8 shows the uncoded BER performance when the receiver only has the *estimated* channel information. It is demonstrated by Fig. 7 that the proposed algorithm achieves better performance in phase noise compensation even if the receiver has perfect channel information. Compared to the IQ+CPE scheme, the proposed method achieves lower BERs, because it not only corrects the common phase rotation of the received constellation but also suppresses part of the inter-carrier interference caused by phase noise. In other words, the proposed algorithm can reduce the sensitivity of OFDM receivers to the analog impairments effectively. Fig. 8 shows that if the receivers have to estimate the channel response, the proposed channel estimation algorithm obtains better channel estimates and thus improves the system performance.

The proposed channel estimation algorithm needs to solve a linear least-squares problem, i.e., (25), iteratively, while the proposed data symbol estimation algorithm needs to solve two least-squares problems, i.e., (34) and (35), iteratively. The simulations suggest that about 10 iterations for the channel estimation and 20 iterations for the data symbol estimation are generally sufficient to guarantee convergence. Solving a general least-squares problem of size  $N$  has computational complexity  $O(N^3)$ . Thus the complexity of the proposed scheme is  $O(KN^3)$ , where  $K$  denotes the number of iter-

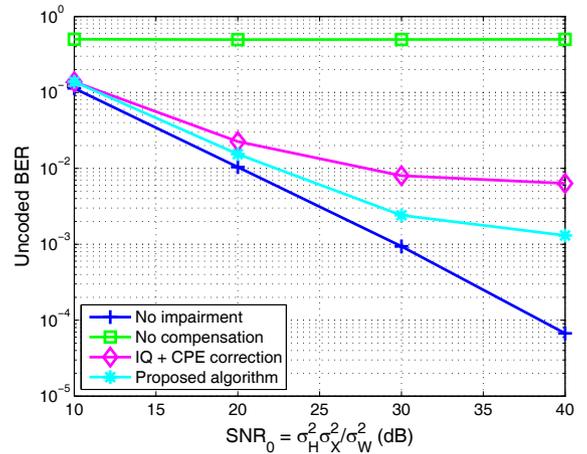


Fig. 7. Plots of uncoded BER vs.  $\text{SNR}_0$  when the receiver has the *perfect* channel information. Four scenarios are simulated: i) There is no impairment. ii) Both IQ imbalance and phase noise are present, but no compensation is applied. iii) Both IQ imbalance and phase noise are present, and the IQ+CPE correction scheme proposed in [17] is applied. iv) Both IQ imbalance and phase noise are present, and the proposed data symbol estimation algorithm is applied.

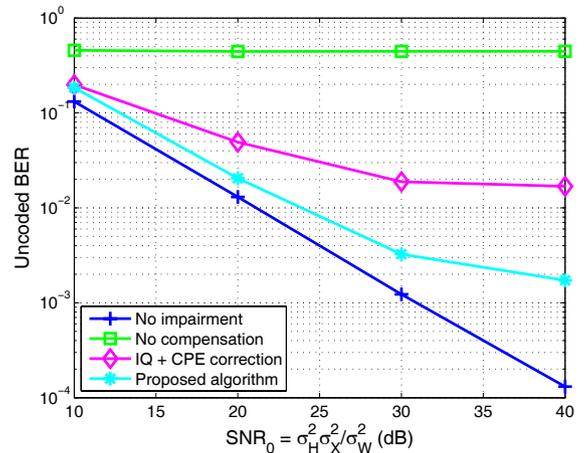


Fig. 8. Plots of uncoded BER vs.  $\text{SNR}_0$  when the receiver only has the *estimated* channel information. Four scenarios are simulated: i) There is no impairment. ii) Both IQ imbalance and phase noise are present, but no compensation is applied. iii) Both IQ imbalance and phase noise are present, and the IQ+CPE correction scheme proposed in [17] is applied. iv) Both IQ imbalance and phase noise are present, and the proposed data symbol estimation algorithm is applied.

ations. The aforementioned complexity can be reduced by the following efficient implementation. First, the channel estimation algorithm is only exploited occasionally, e.g., once per several packets, because the channel and IQ parameters are usually slowly time-varying. For data symbol estimation, the circularly symmetric structure of  $\mathbf{A}$  and the sparse structure of  $\mathbf{P}$  can be exploited to achieve complexity  $O(N \log_2 N)$ .

## VI. CONCLUSIONS

In this paper, the joint effects of IQ imbalance and phase noise on OFDM systems are studied. A compensation scheme is proposed that consists of two stages. One stage is the joint channel estimation, and the other is the joint data symbol estimation. The proposed channel estimation algorithm

performs close to the derived Cramer-Rao lower bound in the presence of the impairments. Also, the analysis and simulations show that the compensation scheme can effectively improve the system performance and reduce the sensitivity of OFDM receivers to the analog impairments. This work can be further extended to include other analog distortions, e.g., carrier frequency offset [22]. Since receivers with less analog impairments usually have the disadvantage of high implementation cost, our technique enables the use of low-cost receivers for OFDM communications.

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**Qiyue Zou** (S'06-M'09) received the B.Eng. and M.Eng. degrees in electrical and electronic engineering from Nanyang Technological University (NTU), Singapore, in 2001 and 2004, respectively. He received the M.A. degree in Mathematics and the Ph.D. degree in Electrical Engineering from University of California, Los Angeles (UCLA) in 2008. From 2001 to 2004, he was with the Centre for Signal Processing of NTU, working on statistical and array signal processing. He is currently with the WiLinx Corporation, San Diego, working on ultra-wideband communications. His research interests include signal processing, information theory, and wireless communications. He received the Young Author Best Paper Award in 2007 from the IEEE Signal Processing Society.



**Alireza Tarighat** (S'00-M'05) received the B.Sc. degree in electrical engineering from Sharif University of Technology, Tehran, Iran, in 1998. He received the M.Sc. and Ph.D. degrees in electrical engineering from the University of California, Los Angeles (UCLA) in 2001 and 2005, respectively.

During the summer of 2000, he was with Broadcom, El Segundo, CA, where he worked on IEEE 802.11 transceivers. From 2001 to 2002 he was with Innovics Wireless, Los Angeles, CA, working on system and ASIC development of advanced antenna diversity and rake processing for 3G WCDMA mobile terminals. Since 2005, he has been with WiLinx, San Diego, CA, working on system and silicon development of UWB wireless networks. His research interests are in communications theory and signal processing, including MIMO OFDM systems, multi-user MIMO wireless networks, algorithms for impairments compensation, and experimental and practical communications systems.

Mr. Tarighat was the recipient of the Gold Medal of the National Physics Olympiad, Iran, 1993, and the Honorable Mention Diploma of the 25th International Physics Olympiad, Beijing, China, 1994. He received the 2006 Outstanding PhD Dissertation Award in electrical engineering from UCLA.



**Ali H. Sayed** (F'01) received the Ph.D. degree in 1992 from Stanford University, Stanford, CA. He is Professor and Chairman of Electrical Engineering with the University of California, Los Angeles (UCLA) and Principal Investigator of the Adaptive Systems Laboratory. He has published widely, with more than 300 articles and four books, in the areas of statistical signal processing, estimation theory, adaptive filtering, signal processing for communications and wireless networking, and fast algorithms for large structured problems. He is coauthor of the

textbook *Linear Estimation* (Englewood Cliffs, NJ: Prentice Hall, 2000), of the research monograph *Indefinite Quadratic Estimation and Control* (Philadelphia, PA: SIAM, 1999), and coeditor of *Fast Algorithms for Matrices with Structure* (Philadelphia, PA: SIAM, 1999). He is also the author of the textbooks *Fundamentals of Adaptive Filtering* (Hoboken, NJ: Wiley, 2003), and *Adaptive Filters* (Hoboken, NJ: Wiley, 2008). He has contributed several encyclopedia and handbook articles. Dr. Sayed is a Fellow of IEEE for his contributions to adaptive filtering and estimation algorithms. He has served on the editorial boards of the IEEE SIGNAL PROCESSING MAGAZINE, the EUROPEAN SIGNAL PROCESSING JOURNAL, the INTERNATIONAL JOURNAL ON ADAPTIVE CONTROL AND SIGNAL PROCESSING, and the SIAM JOURNAL ON MATRIX ANALYSIS AND APPLICATIONS. He has also served as the Editor-in-Chief of the IEEE TRANSACTIONS ON SIGNAL PROCESSING (2003–2005), and the EURASIP JOURNAL ON ADVANCES IN SIGNAL PROCESSING (2006–2007). He is a member of the Signal Processing for Communications and the Signal Processing Theory and Methods Technical Committees of the IEEE Signal Processing Society. He has served on the Publications (2003–2005), Awards (2005), and Conference (2007–present) Boards of the IEEE Signal Processing Society. He also served on the Board of Governors of the IEEE Signal Processing Society (2007–2008) and is currently the Vice-President of Publications of the same Society.

His work has received several recognitions including the 1996 IEEE Donald G. Fink Award, 2002 Best Paper Award from the IEEE Signal Processing Society, 2003 Kuwait Prize in Basic Sciences, 2005 Terman Award, 2005 Young Author Best Paper Award from the IEEE Signal Processing Society, and two Best Student Paper Awards at international meetings (1999 and 2001). He has served as a 2005 Distinguished Lecturer of the IEEE Signal Processing Society and as General Chairman of ICASSP 2008.