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Cognitive Networks in the Presence of I/Q Imbalance and Imperfect CSI: Receiver Design and Performance Analysis

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ABSTRACT Future wireless communication systems, including fifth-generation (5G) networks and the Internet of Things (IoT), require a massive number of inexpensive transceivers. These transceivers come with various hardware impairments, such as phase noise and in-phase/quadrature phase (I/Q) imbalance. This piece of work studies the performance of underlay cognitive radio (CR) networks, considering the joint effect of I/Q imbalance and imperfect channel-state information (CSI) at the secondary user. In order to mitigate the effect of I/Q imbalance, an optimal maximum likelihood (ML) receiver design is proposed and analyzed. Specifically, a closed-form expression of the average pairwise error probability (APEP) and a tight upper bound of the average bit error rate (ABER) are derived. In addition, a widely linear equalization (WLE) receiver that has performance close to the optimal receiver with a computational complexity close to the traditional blind receiver is proposed. In particular, the exact PEP of this WLE receiver is obtained and its APEP is calculated numerically. Moreover, an exact expression is derived for Cramer–Rao lower bound (CRLB) of the secondary system receiver channel estimation error in the presence of I/Q imbalance at the secondary transmitter/receiver (STx/SRx) sides. Computer simulations prove the analytical results of the proposed receivers. The obtained results show that the optimal receiver has the best performance and the WLE receiver outperforms the traditional ML receiver in most cases. In addition, the analysis shows that the best estimator that reaches the CRLB is not affected by the I/Q imbalance at STx/SRx.

INDEX TERMS Channel estimation errors, cognitive radio, Cramer–Rao lower bound, error performance analysis, hardware impairments, I/Q imbalance.

I. INTRODUCTION

The accelerated developments in wireless communication technology are moving the world toward a fully connected network with new challenges, including the increased demands for the radio spectrum. At the same time, the traditional way of spectrum allocation policy has made the limited spectrum overcrowded. Moreover, the available spectrum has not been utilized sufficiently as reported by the Federal Communications Commission [1].

Cognitive radio (CR) was proposed as a novel solution to alleviate the spectrum scarcity by allowing the secondary user (SU) to share the spectrum of the primary user (PU) [2]. Intensive research has been done in the spectrum-sharing side of CR, which has revealed many CR models. These can be classified into three main models: underlay, overlay, and interweave networks [3].

CR communication systems adapt their transmission to the surrounding radio environment. Accordingly, the performance of the CR systems can be significantly affected by different types of practical imperfections, including noise signal uncertainty, imperfect channel state information (CSI), transceiver hardware impairments, and synchronization issues [4]. Hardware impairments, such as in-phase and quadrature-phase (I/Q) imbalance in the radio frequency

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front-end, high power amplifier imperfections, and low noise amplifier non-linearity, can dramatically degrade the system performance [5].

Although all hardware impairments can degrade the system performance, I/Q imbalance represents the most significant source of analog impairments in high-speed wireless communication systems [6]. Specifically, this degradation in performance results from the insufficient rejection of the image frequency band [7], [8]. Moreover, taking the effect of I/Q imbalance into account is not only important in the design of the transceiver, but also in choosing coding techniques as well as in resource management of radio communication systems [9]. Furthermore, hardware impairments have a negative impact on the system secrecy [10].

A. RELATED WORKS

The effect of I/Q imbalance on the energy detection based spectrum sensing for half-duplex CR was studied in [6] and [11]. In [6], it was shown that the I/Q imbalance can cause the SU to interfere with an OFDMA primary system and harshly destroy the performance of the CR system. The work in [11] concluded that the I/Q imbalance is negligible in the case of the single-channel receiver but has a dramatic impact on the wide-band multi-channel sensing receiver. The joint effect of I/Q imbalance and self-interference suppression on the energy detection based spectrum sensing of full-duplex CR was studied in [12]. In this study, it was proven that ignoring the effects of I/Q imbalance and partial self-interference suppression can lead to a dramatic degradation in the system performance in the case of single-channel while the energy detection capability can be entirely restrict in case of multi-channel.

The work in [13] analyzed the performance of cognitive amplify-and-forward (AF) multi-relay networks with active direct link in the presence of relay transceiver hardware impairments. It was shown that the hardware impairments have a high impact on the partial relay selection scheme and a worse impact on the opportunistic relay selection scheme. The authors in [14] studied the impact of transceiver hardware impairments on decode-and-forward (DF) CR networks. In the mentioned work, the authors showed that the effect of transceiver impairments on the outage performance in the high SNR region is more critical than that in low SNR.

The same authors examined the impact of the transceiver impairment on the outage probability and throughput of the DF/AF CR relay in [15]. It was shown that the hardware impairments can deteriorate the network performance, and the DF CR networks outperforms the AF CR networks in terms of both outage probability and throughput but with more system complexity. The study in [16] considered the soft information relaying (SIR) with transceiver hardware impairments in CR networks. It was proven that the SIR protocol outperforms hard DF technique, and the ceiling capacity exists even when the transmitter power approximates to the infinity, while it decreases with increasing levels of hardware impairments.

The joint impact of the hardware impairments and imperfect CSI on the CR networks was studied in [17] and [18]. The work in [17] examined this joint impact on cognitive spatial modulation multiple-input multiple-output systems. This work did not propose any receivers design to mitigate the joint effects of the hardware impairments and imperfect CSI. Furthermore, In [18], the joint effects of the hardware impairments and imperfect CSI on spectrum sharing multiple-relay networks was studied. It was shown that the effect of the hardware impairments limits the system performance and causes various ceiling effects including relay cooperation ceiling (RCC), direct link ceiling (DLC), and overall system ceiling (OSC). This work neither study the the effect of hardware impairments and imperfect CSI on the system bit error rate nor propose any receivers design to mitigate these effects.

It is worthy mentioning that all the aforementioned works of [6], [11]–[18] modeled the hardware impairments including the I/Q imbalance as additive proper Gaussian noise with different means and variances, which is not accurate at least for I/Q imbalance impairment. In this paper, the joint effects of I/Q imbalance and imperfect CSI are proven to be an improper Gaussian random variable (RV), consequently, all receivers design and performance analysis are performed based on this result.

B. CONTRIBUTIONS

Compared to the existing literature and motivated by the importance of the aforementioned reasons, the contributions of this paper can be summarized as follows:

1) The performance of the underlay CR secondary system is studied under the joint effect of I/Q imbalance at STx/SRx sides and imperfect CSI at the SRx side. Subsequently, it is shown that these effects can degrade the system performance and change the noise behavior from proper to improper Gaussian distribution.
2) An optimal maximum likelihood (ML) receiver design that can diminish the effect of I/Q imbalance, is presented and examined. Specifically, a closed-form expression for the average pairwise error probability (APEP) and a tight upper bound of the average bit error rate (ABER) are derived. The simulated results prove that the presented design outperforms all other existing receivers.
3) Moreover, a widely linear equalization (WLE) receiver, which achieves a performance close to the optimal receiver with less complexity, is proposed and analyzed. The exact PEP is derived and the APEP is calculated analytically. Interestingly, the obtained results showed that this receiver outperforms the traditional ML receiver if there is I/Q imbalance at both STx/SRx sides. In addition, it has the same performance with the optimal receiver if the I/Q imbalance is only at SRx side.
4) Besides, an exact expression is derived for CRLB of the secondary system channel estimation in the presence of I/Q imbalance at STx/SRx sides. This expression

can be used as a benchmark to predict and evaluate the performance of the estimators.

5) Finally, the computational complexities of the proposed receivers are calculated and compared with the traditional blind receiver. Our calculations show that the WLE receiver has a computational complexity close to the blind one.

C. PRELIMINARIES

Consider a zero mean complex random variable (RV) $Z = Z^I + jZ^Q$ with the real part Z^I, and the imaginary part Z^Q The variance of Z is defined as $\sigma_Z^2 = \mathbb{E}\{ZZ^*\}$ and the pseudo-variance is defined as $\hat{\sigma}_Z^2 = \mathbb{E}\{ZZ\}$

A complex RV is called proper or circular RV if its pseudo-variance is equal to zero, otherwise it is called non-circular or improper RV [19], [20]. From this definition, the pseudo-variance equals to zero if and only if the real and imaginary parts are circularly symmetric (uncorrelated and have the same variance). There are two different forms of an improper RV [21]: Identical correlated RV, when real and imaginary parts are correlated and have equal variances. Non-identical uncorrelated RV, when real and imaginary parts are uncorrelated but have different variances.

Organization: Section II provides the system and channel models. Next, in Section III, ML optimal and WLE receivers designs as well as the performance analysis for both receivers are presented. In Section IV, an exact closed-form CRLB expression is derived. Subsequently, the computational complexity analysis is studied in V. Afterward, the numerical analysis and results are discussed in Section VI. Finally, Section VII concludes this paper.

Notation: $(.)^{-1}$ indicates matrix inverse and $[\ .\ ]^T$ is the vector transpose. $(.)^I$ and $(.)^Q$ denote the I and Q components. $(.)^*$ is the complex conjugate. $\mathbb{E}\{\cdot\}$ is the expectation operator, and $\Re\{\cdot\}$ denotes the real part of a complex variable. $\mathcal{CN}(\mu, \sigma^2)$ represents the complex-valued Gaussian distribution with mean μ and variance σ^2. $Pr(.)$ is the probability of the event. $Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^{\infty} \exp\left(\frac{-u^2}{2}\right) du$.

II. SYSTEM AND CHANNEL MODELS

A. UNDERLAY COGNITIVE RADIO SYSTEM

In an underlay CR model, the STx can use the spectrum of PUs as long as the interference it generates to the most affected primary receiver (PRx) remains below a predefined threshold I_p. In this work, as it can be seen in Fig. 1, an underlay spectrum sharing system is considered with SUs pair of one STx and one SRx that coexist with another licensed primary transmitter (PTx). Hence, the STx energy (E) is constrained as

$$E = \min\left(\frac{I_p}{|f|^2}, E_m\right), \tag{1}$$

where E_m is the maximum available power at the STx, and f is the channel coefficient between the STx and PRx, $|f|^2$ has an exponential distribution with a mean equals to λ.

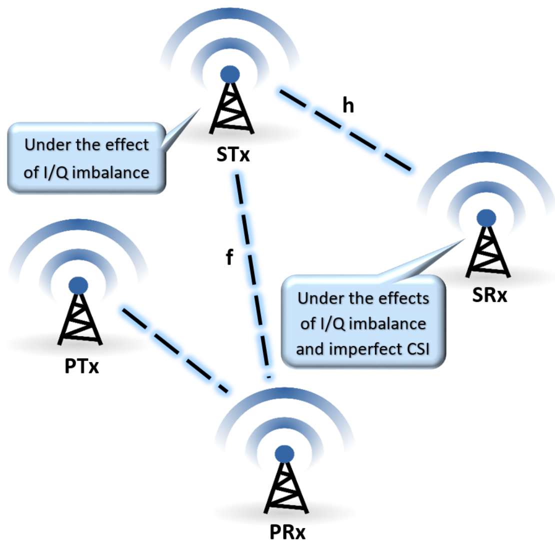


FIGURE 1. System model.

The received signals at the SRx with perfect matching (no I/Q imbalance) is given by

$$y_r = \sqrt{E} h x_i + n, \tag{2}$$

where, x_i represents the transmitted signal, $n \sim \mathcal{CN}(0, 1)$ is the additive white Gaussian noise (AWGN) with identical uncorrelated real and imaginary parts, i.e., $\sigma_{n_I}^2 = \sigma_{n_Q}^2 = 0.5$, and h is the fading channel between the STx and SRx.

B. TRANSCEIVER I/Q IMBALANCE MODEL

The direct-conversion transceiver is built upon the principle of directly up-converting the baseband signal to the radio frequency at the STx and directly down-converting the received signal at SRx. At the STx, the signal is first passed through a digital to analog converter (DAC), then passed through a local oscillator (LO) quadrature mixer. Next, I and Q parts are added together before transmission. At the SRx, the received signal is amplified by a low-noise amplifier, and then passed through a LO quadrature mixer.

After that, I and Q parts are filtered by low noise filters before they passed through an analog to digital converter (ADC) [22]–[24]. The hardware impairments of practical systems occur at LO, phase shifter and I/Q mixer. Consequently, due to the imperfection: 1) the phase difference between the I and Q parts of STx signal and/or SRx signal might not be exactly 90 degrees which is called phase imbalance, 2) small variations might be between the amplitude of the I and Q parts at the STx and/or at the SRx of the signal, which is called amplitude imbalance. Then, I/Q imbalance can dramatically affect the system performance by changing the transmitted signal at the STx and/or corrupting the received signal at the SRx.

Considering the effect of I/Q imbalance at the STx, the transmitted signal can be given as [25]

$$x_i^{IQI} = G_1 x_i + G_2 x_i^*, \tag{3}$$

where G_1 and G_2 are I/Q imbalance parameters at the STx which are defined by [26]

$$G_1 = \frac{1}{2}(1+\xi_t e^{j\beta_t}), \quad G_2 = \frac{1}{2}(1-\xi_t e^{j\beta_t}), \tag{4}$$

where β_t and ξ_t model the phase and amplitude imbalance, respectively. Considering the total effective of I/Q imbalance at the STx and SRx sides, the signal x_i^{IQI} is corrupted and the received signal can be given as

$$\begin{aligned} y &= K_1(\sqrt{E}hx_i^{IQI}+n) + K_2(\sqrt{E}hx_i^{IQI}+n)^* \\ &= K_1(\sqrt{E}h(G_1x_i+G_2x_i{}^*)+n) \\ &\quad +K_2(\sqrt{E}h(G_1x_i+G_2x_i{}^*)+n)^*, \end{aligned} \tag{5}$$

where K_1 and K_2 are I/Q imbalance parameters at the SRx which are defined by [26]

$$K_1 = \frac{1}{2}(1+\xi_r e^{-j\beta_r}), \quad K_2 = \frac{1}{2}(1-\xi_r e^{j\beta_r}), \tag{6}$$

where β_r and ξ_r model the phase and amplitude imbalance, respectively.

The terms $x_i{}^*$ and $(\sqrt{E}hx_i^{IQI}+n)^*$ in (3) and (5) are the self-interference introduced by the I/Q imbalance at the STx and SRx, respectively. It can be noted that for perfect I/Q balance, the amplitude imbalance parameters $\xi_t = \xi_r = 1$ and the phase imbalance parameters $\beta_t = \beta_r = 0$ Consequently, $G_1 = K_1 = 1$ and $G_2 = K_2 = 0$

C. IMPERFECT CSI MODEL

In the case of imperfect CSI, there is an estimation error at the receiver. The SRx uses one of the channel estimators such as ML or mean square error estimators to get a channel estimate, which can be characterized as follows [27]

$$h = \hat{h} + e, \tag{7}$$

where $\hat{h}$ is the channel estimation and $e \sim \mathcal{CN}(0, \sigma_e^2)$, is the channel estimation error. Here we assume that h and $\hat{h}$ are jointly ergodic and stationary Gaussian processes. Further, assuming orthogonality between the channel estimate and the estimation error. Note that the variance of e includes the information of the channel estimation quality [28]. In Section V, we calculate the CRLB of $\sigma_{e_I}^2$ and $\sigma_{e_Q}^2$ in the presence of I/Q imbalance at STx/SRx. There we prove that e is a proper RV, where e^I and e^Q are identical and uncorrelated even though the I/Q imbalance change the total noise behavior from proper to improper RV.

Considering the effect of imperfect CSI at the SRx, (5) can be rewritten as

$$\begin{aligned} y &= K_1\big(\sqrt{E}(\hat{h}+e)(G_1x_i+G_2x_i{}^*)+n\big) \\ &\quad + K_2\big(\sqrt{E}(\hat{h}+e)^*(G_1x_i+G_2x_i{}^*)^*+n^*\big) \\ &= \sqrt{E}\{\underbrace{K_1\hat{h}G_1+K_2\hat{h}^*G_2^*}_{\tilde{h}_1}\}x_i + \sqrt{E}\{\underbrace{K_1\hat{h}G_2+K_2\hat{h}^*G_1^*}_{\tilde{h}_2}\}x_i^* \\ &\quad + \sqrt{E}K_1e\underbrace{(G_1x_i+G_2x_i^*)}_{g_{a_i}} + \sqrt{E}K_2e^*\underbrace{(G_1^*x_i^*+G_2^*x_i)}_{g_{b_i}} \\ &\quad + K_1n+K_2n^*. \end{aligned} \tag{8}$$

The previous equation shows that the generated noise depends on the noise at the receiver, the channel estimation error, the transmitted symbol, the transmitted energy, and the I/Q imbalance parameters at the STx/SRx. Understanding the characteristics of this noise is the critical factor in designing and analyzing the appropriate receiver.

D. IMPROPER GAUSSIAN NOISE AT THE SRx

The remaining part of this section studies the characteristics of the resulting noise. The received signal can be separated into two parts: the signal part χ_i and the noise part $\tilde{n}$ as in the following

$$\begin{aligned} y &= \underbrace{\sqrt{E}(\tilde{h}_1x_i+\tilde{h}_2x_i^*)}_{\chi_i\ (signal)} \\ &\quad + \underbrace{(\sqrt{E}K_1g_{a_i})e+(\sqrt{E}K_2g_{b_i})e^*+K_1n+K_2n^*}_{\tilde{n}_i\ (noise)}. \end{aligned} \tag{9}$$

Real and imaginary components of the signal part (χ_i) can be given by

$$\begin{aligned} \chi_i^I &= (\tilde{h}_1^I+\tilde{h}_2^I)x_i^I + (\tilde{h}_2^Q-\tilde{h}_1^Q)x_i^Q. \\ \chi_i^Q &= (\tilde{h}_1^Q+\tilde{h}_2^Q)x_i^I + (\tilde{h}_1^I-\tilde{h}_2^I)x_i^Q. \end{aligned} \tag{10}$$

In the same way, real and imaginary parts of the noise ($\tilde{n}$) can be given by

$$\begin{aligned} \tilde{n}_i{}^I &= \sqrt{E}\big[e^I(\underbrace{K_1^Ig_{a_i}^I-K_1^Qg_{a_i}^Q}_{a_i}+\underbrace{K_2^Ig_{b_i}^I-K_2^Qg_{b_i}^Q}_{b_i}) \\ &\quad + e^Q(\underbrace{K_2^Ig_{b_i}^Q+K_2^Qg_{b_i}^I}_{c_i}-\underbrace{K_1^Ig_{a_i}^Q-K_1^Qg_{a_i}^I}_{-d_i})\big]+n^I \\ \tilde{n}_i{}^I &= \sqrt{E}\{(a_i+b_i)e^I+(c_i-d_i)e^Q\}+n^I. \\ \tilde{n}_i{}^Q &= \sqrt{E}\big[e^I(\underbrace{K_1^Ig_{a_i}^Q+K_1^Qg_{a_i}^I}_{d_i}+\underbrace{K_2^Ig_{b_i}^Q+K_2^Qg_{b_i}^I}_{c_i}) \\ &\quad + e^Q(\underbrace{K_1^Ig_{a_i}^I-K_1^Qg_{a_i}^Q}_{a_i}-\underbrace{K_2^Ig_{b_i}^I+K_2^Qg_{b_i}^Q}_{-b_i})\big]+n^IK_c+n^QK_d \\ \tilde{n}_i{}^Q &= \sqrt{E}\{(c_i+d_i)e^I+(a_i-b_i)e^Q\}+K_cn^I+K_dn^Q. \end{aligned} \tag{11}$$

Two important results can be observed from (11). The first one is that, $\tilde{n}^I$ and $\tilde{n}^Q$ are not identical RVs since $\sigma_{\tilde{n}_i^Q}^2$ and $\sigma_{\tilde{n}^I}^2$ are not equal. $\sigma_{\tilde{n}_i^Q}^2$ and $\sigma_{\tilde{n}^I}^2$ are given by

$$\begin{aligned} \sigma_{\tilde{n}_i^I}^2 &= E(a_i+b_i)^2\sigma_{e^I}^2 + E(c_i-d_i)^2\sigma_{e^Q}^2 + \frac{\sigma_n^2}{2}. \\ \sigma_{\tilde{n}_i^Q}^2 &= E(c_i+d_i)^2\sigma_{e^I}^2 + E(a_i-b_i)^2\sigma_{e^Q}^2 + (K_c^2+K_d^2)\frac{\sigma_n^2}{2}, \end{aligned} \tag{12}$$

where $K_c = K_1^Q + K_2^Q$, and $K_d = K_1^I - K_2^I$ The second note is that $\tilde{n}_i{}^I$ and $\tilde{n}_i{}^Q$ are correlated RVs, since $\mathbb{E}\{\tilde{n}_i{}^I\tilde{n}_i{}^Q\} \neq 0$ The correlation coefficient of $\tilde{n}_i{}^I$ and $\tilde{n}_i{}^Q$ can be calculated from

$$\varrho_i = \frac{\mathbb{E}\{\tilde{n}_i{}^I\tilde{n}_i{}^Q\}}{\sqrt{\sigma_{\tilde{n}_i^I}^2\sigma_{\tilde{n}_i^Q}^2}}, \tag{13}$$

where $\mathbb{E}\{\tilde{n}_i^I \tilde{n}_i^Q\}$ is given by

$$\mathbb{E}\{\tilde{n}_i^I \tilde{n}_i^Q\} = E\{(a_i+b_i)(c_i+d_i)\sigma_{e^I}^2 + (c_i-d_i)(a_i-b_i)\sigma_{e^Q}^2\} + K_c\frac{\sigma_n^2}{2}. \quad (14)$$

From (12) and (13), it is clear that, $\tilde{n}_i$ is improper Gaussian RV, since $\sigma^2_{\tilde{n}_i^I}$ and $\sigma^2_{\tilde{n}_i^Q}$ are not equal and, in general, $\tilde{n}_i^I$ and $\tilde{n}_i^Q$ are correlated. This change in noise behavior from proper to improper Gaussian noise demands new requirements in the receiver design.

III. RECEIVER DESIGNS AND PERFORMANCE ANALYSES

The previous section shows how the I/Q imbalance impairment can change the noise behavior from proper to improper RV. This requires designing new receivers to tackle these changes. In this section, two new receivers are proposed: optimal ML and WLE receiver, then the blind (traditional) receiver is discussed.

A. OPTIMAL ML RECEIVER

In this section, an optimal ML receiver is proposed for the presented CR wireless communication system, which has I/Q imbalance at STx/SRx and imperfect CSI at SRx. Considering the general signal model in (9), and the results in (12) and (13), and assuming that the I/Q imbalance parameters are known at the SRx, the joint probability density function (PDF) of the real part, y^I, and the imaginary part, y^Q, of the received signal can be written as [29], [30]

$$f_{y^I,y^Q}(y^I,y^Q|x_i) = \frac{1}{2\pi\sigma_{\tilde{n}_i^I}\sigma_{\tilde{n}_i^Q}\sqrt{1-\varrho_i^2}} \exp\left(\frac{-1}{2(1-\varrho_i^2)}\left[\frac{(y^I-\sqrt{E}\chi_i^I)^2}{\sigma^2_{\tilde{n}_i^I}} + \frac{(y^Q-\sqrt{E}\chi_i^Q)^2}{\sigma^2_{\tilde{n}_i^Q}} - \frac{2\varrho_i(y^I-\sqrt{E}\chi_i^I)(y^Q-\sqrt{E}\chi_i^Q)}{\sigma_{\tilde{n}_i^I}\sigma_{\tilde{n}_i^Q}}\right]\right). \quad (15)$$

The primary task of the ML receiver is to decide which x_i was transmitted among M hypotheses. Assuming that the channel inputs are equally likely, the optimal receiver is designed based on maximizing the following statement

$$\hat{x}_i = arg \max_{i=1,\ldots,M} \{f_{y^I,y^Q}(y^I,y^Q|x_i)\}. \quad (16)$$

Maximizing the previous statement is equivalent to

$$\hat{x}_i = arg \min_{i=1,\ldots,M}\left\{\frac{(y^I-\sqrt{E}\chi_i^I)^2}{\sigma^2_{\tilde{n}_i^I}} + \frac{(y^Q-\sqrt{E}\chi_i^Q)^2}{\sigma^2_{\tilde{n}_i^Q}} - \frac{2\varrho_i(y^I-\sqrt{E}\chi_i^I)(y^Q-\sqrt{E}\chi_i^Q)}{\sigma_{\tilde{n}_i^I}\sigma_{\tilde{n}_i^Q}}\right\}. \quad (17)$$

Conditional PEP_{opt}: From (17), the probability of detecting $\hat{x}_i$ at the SRx given that the STx transmitted x_i, is given by

$$\mathrm{PEP_{opt}} = Pr\left\{\frac{(y^I-\sqrt{E}\chi_i^I)^2}{\sigma^2_{\tilde{n}_i^I}} + \frac{(y^Q-\sqrt{E}\chi_i^Q)^2}{\sigma^2_{\tilde{n}_i^Q}} - \frac{2\varrho_i(y^I-\sqrt{E}\chi_i^I)(y^Q-\sqrt{E}\chi_i^Q)}{\sigma_{\tilde{n}_i^I}\sigma_{\tilde{n}_i^Q}} > \frac{(y^I-\sqrt{E}\hat{\chi}_i^I)^2}{\sigma^2_{\tilde{n}_i^I}} + \frac{(y^Q-\sqrt{E}\hat{\chi}_i^Q)^2}{\sigma^2_{\tilde{n}_i^Q}} - \frac{2\varrho_i(y^I-\sqrt{E}\hat{\chi}_i^I)(y^Q-\sqrt{E}\hat{\chi}_i^Q)}{\sigma_{\tilde{n}_i^I}\sigma_{\tilde{n}_i^Q}}\right\}. \quad (18)$$

After some simplifications, (18) can be written as

$$\mathrm{PEP_{opt}} = Pr\left\{\mathcal{N}_i > \frac{E(\chi_i^I-\hat{\chi}_i^I)^2}{\sigma^2_{\tilde{n}_i^Q}} + \frac{E(\chi_i^Q-\hat{\chi}_i^Q)^2}{\sigma^2_{\tilde{n}_i^Q}} - \frac{2\varrho_i E(\chi_i^I-\hat{\chi}_i^I)(\chi_i^Q-\hat{\chi}_i^Q)}{\sigma_{\tilde{n}_i^I}\sigma_{\tilde{n}_i^Q}}\right\}, \quad (19)$$

where $\mathcal{N}_i$ is given by

$$\mathcal{N}_i = 2\varrho_i\sqrt{E}\left[\frac{(\chi_i^I-\hat{\chi}_i^I)\tilde{n}_i^Q+(\chi_i^Q-\hat{\chi}_i^Q)\tilde{n}_i^I}{\sigma_{\tilde{n}_i^I}\sigma_{\tilde{n}_i^Q}}\right] - 2\sqrt{E}\left[\frac{(\chi_i^I-\hat{\chi}_i^I)\tilde{n}_i^I}{\sigma^2_{\tilde{n}_i^I}} + \frac{(\chi_i^Q-\hat{\chi}_i^Q)\tilde{n}_i^Q}{\sigma^2_{\tilde{n}_i^Q}}\right]. \quad (20)$$

From (20), it is clear that $\mathcal{N}_i$ is a Gaussian RV with zero-mean and its variance $\sigma^2_{\mathcal{N}_i}$ can be expressed as

$$\sigma^2_{\mathcal{N}_i} = 4E(1-\varrho_i^2) \times\left[\frac{(\chi_i^I-\hat{\chi}_i^I)^2}{\sigma^2_{\tilde{n}_i^I}} + \frac{(\chi_i^Q-\hat{\chi}_i^Q)^2}{\sigma^2_{\tilde{n}_i^Q}} - \frac{(\chi_i^I-\hat{\chi}_i^I)+(\chi_i^Q-\hat{\chi}_i^Q)}{\sigma_{\tilde{n}_i^I}\sigma_{\tilde{n}_i^Q}}\right]. \quad (21)$$

Hence, utilizing (1), (19) and (21), $\mathrm{PEP_{opt}}$ can also be written by using the well-known Q-function formula as in (22) at the top of the next page.

Average PEP_{opt}: Without loss of generality, limited feedback from the PRx is assumed, based on this, mean-value power allocation (MVPA) method can be exploited [31]–[33]. Relying on MVPA, the interference channel gain between the STx-PRx is usually assumed to be known at PRx. Therefore, PRx computes the mean value of this channel gain, instead of the instantaneous CSI. Then, PRx feeds the calculated mean back to the STx. As a result, sending one value rather than instantaneous CSI feedback for each symbol or block of symbols can significantly minimize the system complexity and decrease the feedback burden. Relying on MVPA, E in (1) is constrained as $E = \min(\frac{I_p}{\sigma_f^2}, E_m)$, where $\sigma_f^2 = \mathbb{E}\{|f|^2\}$

Although our receiver design is general for any fading type, the APEP will be calculated considering the Rayleigh fading channel. From (22), γ_{opt} can be written as $\gamma_{opt} = D_1^2/\sigma^2_{\tilde{n}_i^I} + D_2^2/\sigma^2_{\tilde{n}_i^Q} - 2\varrho_i D_1 D_2/(\sigma_{\tilde{n}_i^I}\sigma_{\tilde{n}_i^Q})$, where $D_1 = (\chi_i^I - \hat{\chi}_i^I)/\sigma_{\tilde{n}_i^I}$ and $D_2 = (\chi_i^Q - \hat{\chi}_i^Q)/\sigma_{\tilde{n}_i^Q}$. Hence, γ_{opt} has the PDF of a quadratic form of Gaussian RVs. From [29], [34] and given

$$\mathrm{PEP_{opt}} = Q\left(\sqrt{\frac{\min\left(\frac{I_p}{|f|^2}, E_m\right)}{4\left(1-\varrho_i^2\right)}\left[\frac{(\chi_i^I-\hat{\chi}_i^I)^2}{\sigma^2_{\tilde{n}_i^I}}+\frac{(\chi_i^Q-\hat{\chi}_i^Q)^2}{\sigma^2_{\tilde{n}_i^Q}}-\frac{2\varrho_i(\chi_i^I-\hat{\chi}_i^I)(\chi_i^Q-\hat{\chi}_i^Q)}{\sigma_{\tilde{n}_i^I}\sigma_{\tilde{n}_i^Q}}\right]}\right) = Q\left(\sqrt{\frac{\min\left(\frac{I_p}{|f|^2}, E_m\right)\gamma_{opt}}{4\left(1-\varrho_i^2\right)}}\right). \tag{22}$$

that, the variance of D_1 is equal to υ_1 and the variance of D_2 is equal to υ_2, the moment generating function (MGF) of γ_{opt} can be given by

$$M_{\gamma_{opt}}(t) = \frac{1}{\sqrt{1-2t\lambda_1}} \times \frac{1}{\sqrt{1-2t\lambda_2}}, \tag{23}$$

where, λ_1 and λ_2 are equal to

$$\lambda_{1,2} = \frac{\upsilon_1+\upsilon_2 \pm \sqrt{(\upsilon_1+\upsilon_2)^2 - 4\left(1-\varrho_i^2\right)\upsilon_1\upsilon_2}}{2}. \tag{24}$$

Now, from [35], and relying on MVPA a closed-form expression of the $\mathrm{APEP_{opt}}$ can be calculated by using (22)-(24) as follows

$$\mathrm{APEP_{opt}} = \frac{1}{\pi}\int_0^{\frac{\pi}{2}} M_{\gamma_{opt}}\left(-\frac{\min\left(\frac{I_p}{\sigma_f^2}, E_m\right)}{8\left(1-\varrho_i^2\right)\sin^2\theta}\right)d\theta. \tag{25}$$

The previous integration can be calculated by using a simple numerical integration technique. Furthermore, (25) can be simply upper bounded by

$$\mathrm{APEP_{opt}} \leq \frac{1}{2} M_{\gamma_{opt}}\left(-\frac{\min\left(\frac{I_p}{\sigma_f^2}, E_m\right)}{8\left(1-\varrho_i^2\right)}\right). \tag{26}$$

Average BER_{opt}: From (22), and using the well-known union bound technique [36], the ABER can be calculated as

$$\mathrm{ABER}_{opt} \leq \sum\nolimits_{t=1}^{M} \sum\nolimits_{\hat{t}=t+1}^{M} \frac{2N(\chi_i,\hat{\chi}_i)\,\mathrm{APEP}_{opt}}{M}, \tag{27}$$

where $N(\chi_i, \hat{\chi}_i)$ is the number of bit errors associated with the corresponding pairwise error event.

B. WIDELY LINEAR EQUALIZATION (WLE) RECEIVER

Here, WLE receiver is presented at SRx with imperfect CSI when the I/Q imbalance exists at STx/SRx. The goal of this filter is to entirely eliminate the I/Q imbalance. The filter parameters are calculated by adding the scaled received signal with its scaled conjugate and then matching the results with the received signal as in the case of perfect I/Q imbalance and perfect CSI. Here, the scaling parameters are calculated assuming that the I/Q imbalance parameters are known at the SRx. After that, the traditional ML detection is applied at the SRx to choose the correct one among M hypothesis.

Lemma 1: The output of the WLE receiver can be given as

$$Y = \sqrt{E}x_i + Z_i, \tag{28}$$

where $Z_i \sim \mathcal{CN}(0, \sigma^2_{Z_i})$ is an improper RV and the variances of its real and imaginary parts are equal to $\sigma^2_{Z_i^I}$ and $\sigma^2_{Z_i^Q}$, respectively, and the correlation factor between Z_i^I and Z_i^Q is ρ_{Z_i}, where ρ_{Z_i} is calculated as in (49).

Proof: See the Appendix A.

The traditional ML detection for the proposed WLE receiver, which ignores the improper characteristics of the noise, is relying on minimizing the following statement

$$\hat{x}_i = \arg\min_i \left\{|Y - \sqrt{E}x_i|^2\right\}. \tag{29}$$

Conditional PEP_{wle} : Considering (29) and assuming that x_i has been sent, the probability of receiving $\hat{x}_i$ is given as

$$\begin{aligned}\mathrm{PEP_{wle}} &= Pr\left\{|Y-\sqrt{E}x_i|^2 > |Y-\sqrt{E}\hat{x}_i|^2\right\}\\ &= Pr\left\{|Z_i|^2 > E|(x_i-\hat{x}_i)|^2\right.\\ &\quad \left.+|Z_i|^2 + 2\sqrt{E}\Re\left[(x_i-\hat{x}_i)Z_i^*\right]\right\}\\ &= Pr\left\{0 > E|(x_i-\hat{x}_i)|^2 + 2\sqrt{E}\Re\left[(x_i-\hat{x}_i)Z_i^*\right]\right\}\\ &= Pr\left\{0 > |E(x_i-\hat{x}_i)|^2 + \zeta_i\right\},\end{aligned} \tag{30}$$

where $\zeta_i = 2\sqrt{E}\Re\left[(x_i-\hat{x}_i)Z_i^*\right]$ Conditioned on $\hat{x}_i$ and x_i, ζ_i is a Gaussian RV with zero mean and variance of

$$\begin{aligned}\sigma^2_{\zeta_i} &= 4E\mathbb{E}\left\{\left[(x_i^I-\hat{x}_i^I)Z_i^I + (x_i^Q-\hat{x}_i^Q)Z_i^Q\right]^2\right\}\\ &= 4E(x_i^I-\hat{x}_i^I)^2\mathbb{E}\{Z_i^{I^2}\} + 4E(x_i^Q-\hat{x}_i^Q)^2\mathbb{E}\{Z_i^{Q^2}\}\\ &\quad +8E(x_i^I-\hat{x}_i^I)(x_i^Q-\hat{x}_i^Q)\mathbb{E}\{Z_i^I Z_i^Q\}\\ &= 4E(x_i^I-\hat{x}_i^I)^2\sigma^2_{Z_i^I} + 4E(x_i^Q-\hat{x}_i^Q)^2\sigma^2_{Z_i^Q}\\ &\quad +8E(x_i^I-\hat{x}_i^I)(x_i^Q-\hat{x}_i^Q)\sigma_{Z_i^I}\sigma_{Z_i^Q}\rho_{Z_i}.\end{aligned} \tag{31}$$

Considering the mean and variance of ζ_i, the closed-form of $\mathrm{PEP_{wle}}$ can be given using the Q-function as in (32) at the top of the next page.

Average PEP_{wle}: It is clear from (32) that the PDF of γ_{wle} is very complicated and considerably difficult, if not possible, to derive. Thus, the $\mathrm{APEP_{wle}}$ is found numerically by averaging the $\mathrm{PEP_{wle}}$ over a large number of channel realizations for each SNR value. Finally, the ABER of the WLE receiver ($\mathrm{ABER_{wle}}$) can be calculated directly using the formula in (27).

C. BLIND ML RECEIVER

The blind receiver can be defined as the one that utilizes the traditional ML receiver to detect CR signals as if there is no I/Q imbalance at STx or SRx, even in the case where it exists at one or both sides. Based on this scenario, and starting

$$\mathrm{PEP_{wle}} = Q\left(\sqrt{\frac{\min\left(\frac{I_p}{|f|^2}, E_m\right)|x_i - \hat{x}_i|^4}{4(x_i^I - \hat{x}_i^I)^2\sigma^2_{Z_i^I} + 4(x_i^Q - \hat{x}_i^Q)^2\sigma^2_{Z_i^Q} + 8(x_i^I - \hat{x}_i^I)(x_i^Q - \hat{x}_i^Q)\sigma_{Z_i^I}\sigma_{Z_i^Q}\rho_{Z_i}}}\right) = Q\left(\sqrt{\min\left(\frac{I_p}{|f|^2}, E_m\right)\gamma_{wle}}\right). \quad (32)$$

$$\mathrm{PEP_{bli}}=Q\left(\sqrt{\frac{\min\left(\frac{I_p}{|f|^2}, E_m\right)\left(|\chi_i - hx_i|^2 - |\chi_i - h\hat{x}_i|^2\right)^2}{4[(hx_i)^I - (h\hat{x}_i)^I]^2\sigma^2_{\tilde{n}_i^I} + 4[(hx_i)^Q - (h\hat{x}_i)^Q]^2\sigma^2_{\tilde{n}_i^Q} + 8[(hx_i)^I - (h\hat{x}_i)^I][(hx_i)^Q - (h\hat{x}_i)^Q]\sigma_{\tilde{n}_i^I}\sigma_{\tilde{n}_i^Q}\varrho}}\right)=Q(\sqrt{\gamma_{bli}}) \quad (37)$$

from (9), the traditional ML receiver for the blind CR receiver depends on minimizing the following statement

$$\hat{x}_i = arg \min_{i=1,\dots,M} \left\{|\mathbf{y} - \sqrt{E}hx_i|^2\right\}. \quad (33)$$

Conditional PEP_{bli} : The noise $\tilde{n}$ in (9) is an improper RV but it will be treated as if it is proper RV by this blind receiver. The probability of detecting $\hat{x}_i$ at the SRx given that the STx transmitted x_i, is given by

$$\begin{aligned}\mathrm{PEP_{bli}} &= Pr\left\{\left|y - \sqrt{E}hx_i\right|^2 > \left|y - \sqrt{E}h\hat{x}_i\right|^2\right\}\\ &= Pr\left\{\left|\sqrt{E}\chi_i - \sqrt{E}hx_i\right|^2 > \left|\sqrt{E}\chi_i - \sqrt{E}h\hat{x}_i\right|^2\right\}.\end{aligned} \quad (34)$$

After some algebraic simplifications, the conditional error probability can be written as

$$\mathrm{PEP_{bli}} = Pr\left\{0 > E|(\chi_i - h\hat{x}_i)|^2 - E|(\chi_i - hx_i)|^2 + \vartheta_i\right\}. \quad (35)$$

Therefore, after following the same procedure as in (31), $\vartheta_i = 2\sqrt{E}\Re\left[(hx_i - h\hat{x}_i)\tilde{n}_i^*\right]$ is a Gaussian RV with zero mean and variance of

$$\begin{aligned}\sigma^2_{\vartheta_i} &= 4E[(hx_i)^I - (h\hat{x}_i)^I]^2\sigma^2_{\tilde{n}_i^I} + 4E[(hx_i)^Q - (h\hat{x}_i)^Q]^2\sigma^2_{\tilde{n}_i^Q}\\ &\quad + 8E[(hx_i)^I - (h\hat{x}_i)^I][(hx_i)^Q - (h\hat{x}_i)^Q]\sigma_{\tilde{n}_i^I}\sigma_{\tilde{n}_i^Q}\varrho_i .\end{aligned} \quad (36)$$

Based on that, $\mathrm{PEP_{bli}}$ can be calculated as in (37) at the top of this page.

Average PEP_{bli} : It is clear from (37) that the PDF of γ_{bli} is more complicated than the PDF of γ_{wle} in (32), and it is very difficult to find its average. Based on this, the $\mathrm{APEP_{bli}}$ is found through simulation by averaging the $\mathrm{PEP_{bli}}$ over a large number of channel realizations for each SNR value. Finally, the ABER of the blind receiver ($\mathrm{ABER_{bli}}$) can be calculated directly by using (27).

IV. CRAMER-RAO LOWER BOUND (CRLB)

In this section, an exact expression is derived for CRLB of the channel estimation in the presence of I/Q imbalance at the STx/SRx. Since CRLB expresses a lower bound on the variance of unbiased estimators, it can be used as a benchmark to predict and evaluate the estimators performances in the presence of I/Q imbalance.

Lemma 2: Assuming N_P is the number of training pilots that used at the SRx to estimate the channel, CRLB matrix can be given as

$$CRLB = \begin{bmatrix} CRLB_{(h^I)} & 0 \\ 0 & CRLB_{(h^Q)} \end{bmatrix}, \quad (38)$$

where

$$CRLB_{(h^I)} = CRLB_{(h^Q)} = \frac{\sigma_n^2}{2N_pE} . \quad (39)$$

Proof: See the Appendix B.

It is worth mentioning that $CRLB_{(h^I)}$ and $CRLB_{(h^Q)}$ are independent even though the I/Q imbalance changes the Gaussian noise behavior from a proper to an improper RV, as proven in the Appendix. This can be concluded clearly from the diagonal CRLB matrix. Consequently, the quality of estimating h^I and h^Q does not degrade when the other one is unknown. In addition, the I/Q imbalance at the STx or SRx does not affect the channel estimation error.

V. COMPLEXITY ANALYSIS

The computational time complexity can be calculated by finding the number of real additions and real multiplications [37], [38]. It is widely known that each complex multiplication requires four real multiplications and two real additions, while the square of the absolute value of the complex number requires two real multiplications and one real addition.

The detection process of the proposed receivers requires a few calculations for each received symbol, and several calculations will be repeated M times depending on the modulation scheme. Moreover, some calculations will be found one time at the preparation phase only, and other calculations will be found for each coherence time (i.e., time duration over which the channel impulse response is not varying).

It is easy to show that the blind receiver given in (33) needs four real multiplications and two real summations to calculate the term $\sqrt{E}hx_i$ This term is calculated M times for each coherence time. After that, the square of the absolute value is calculated M times, and this needs two real multiplications and two real summations.

On the other hand, the WLE receiver given in (29) needs eight real multiplications and four real summations at the preprocessing phase to compute K_1G_1, $K_2G_2^*$, K_1G_2, and $K_2G_1^*$

TABLE 1. Number of multiplications and summations required for each frame.

	Optimal		WLE		Blind	
	Mul	Sum	Mul	Sum	Mul	Sum
Per received symbol	$6MI$	$6MI$	$8I+2MI$	$6I+3MI$	$2MI$	$3MI$
Per coherence time	$4M+16$	$6M+12$	24	15	$4M$	$2M$
At preprocessing time	$27M+16$	$26M+8$	16	8	–	–

Then, it calculates w_1 and w_2 in (44), and Y in (45). The variables w_1 and w_2 are required to be calculated once for each coherence time, which takes twenty-four real multiplications and fifteen real summations. Moreover, Y is required to be calculated once for each received symbol, and this takes eight real multiplications and six real summations. Finally, the square of the absolute value takes two real multiplications and three real summations and it is calculated M times for each received symbols.

Following the same logic, the optimal receiver detects $\hat{x}_i$ in (17) by calculating $\tilde{h}_1$, and $\tilde{h}_2$ in (9) once for each coherence time, and this takes sixteen real multiplications and twelve real summations. After that $\sigma^2_{\tilde{n}_i I}$, and $\sigma^2_{\tilde{n}_i Q}$ in (12), and ϱ_i in (13) are calculated M times at the preprocessing phase, and this takes twenty-seven real multiplications and twenty-six real summations. At the end, $\hat{x}_i$ can be detected after doing six real multiplications and six real summations M times.

Assuming that I is the number of symbols per coherence time, the total number of real multiplications and summations for all receivers can be seen in Table. 1.

VI. NUMERICAL ANALYSIS AND RESULTS

Considering the aforementioned CR receivers designs, comprehensive computer simulations are carried out to validate the analytical results and assess the performance of the STx/SRx that has I/Q imbalance with imperfect CSI at the SRx. More specifically, a SISO system scenario is assumed, and 4-QAM modulation is used. The computer simulations are performed under Rayleigh fading channel conditions and the receiver is affected by proper Gaussian noise $n \sim \mathcal{CN}(0, 1)$ Moreover, for a fair comparison, the transmitted signal energy is normalized by $(|x_i G_1 + x_i^* G_2|^2)$ All the comparisons are made against a system with perfect I/Q balance. In addition, we assume the predefined threshold $I_p = 30$ dB and f is the channel coefficient between the STx and PRx where $f \sim \mathcal{CN}(0, 1)$ Finally, the ABER for all figures is plotted versus the signal-to-noise ratio (SNR) E/σ_n^2 using (27).

Fig. 2 shows the simulation and numerical results where the conditional PEP averaging of a large number of channel realizations for the optimal, WLE, and blind receivers. In addition, the closed form APEP of the optimal receiver relying on MVPA [equation (25)] is presented in this figure. It is clear that the simulation results match with the analytical analysis for all receivers which validates our results.

Fig. 3 shows the effect of the I/Q imbalance at STx only, with perfect CSI at SRx. Fig. 3a shows the results of fixing the phase mismatch at 10° while varying the amplitude imbalance among 2, 4 and 6 dB. It shows that the performance of all receivers decreases when the level of the amplitude mismatching increases. In addition, the optimal receiver outperforms the WLE and blind receivers at all levels of SNR while the WLE receiver outperforms the blind receiver at higher values of SNR. Finally, it is clear that in the higher SNR region, the ABER performance saturates due to the power constraint limitation, leading to an error floor and zero power gain.

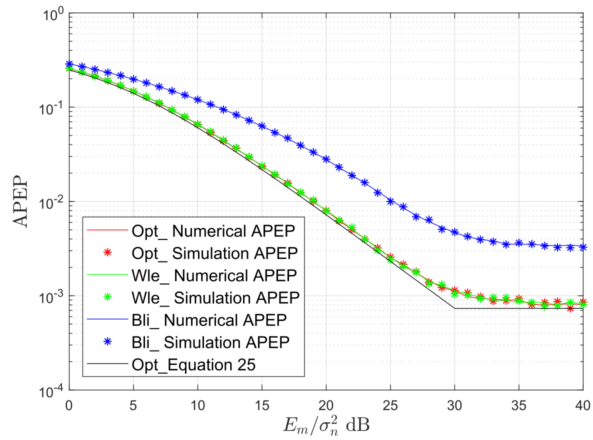


FIGURE 2. APEP at 3 dB amplitude mismatch and 5° phase imbalance at both STx and SRx with perfect CSI at SRx.

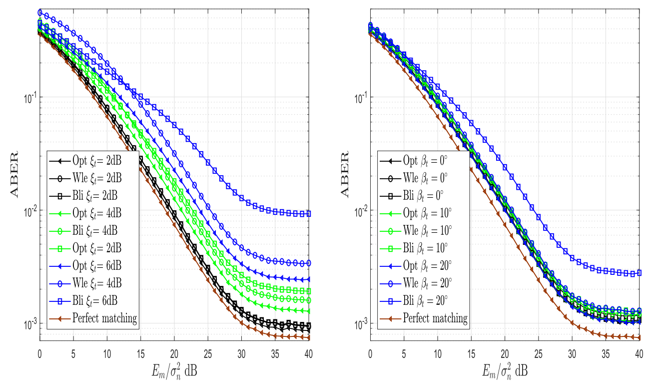


FIGURE 3. I/Q imbalance at STx side only with perfect CSI. (a) Amplitude mismatch. (b) Phase mismatch.

In Fig. 3b, the ABER is plotted when fixing the amplitude mismatch to 3 dB while changing the phase imbalance among 0°, 10° and 20°. It shows that the optimal receiver yields the best performance and it can completely resist the phase mismatching effect at the STx. In addition, the WLE and blind receivers have approximately the same performance at low and intermediate levels of phase mismatching while the WLE receiver exceeds the blind receiver at high level of phase mismatching. Even more, in the high SNR region, the ABER performance saturates, leading to an error floor and zero power gain due energy constraint limitation.

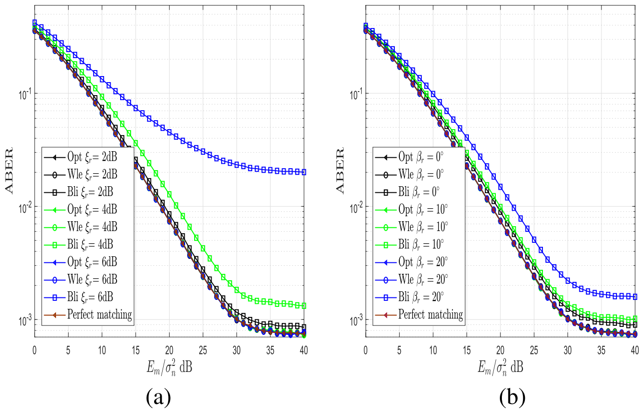


FIGURE 4. I/Q imbalance at SRx side only with perfect CSI. (a) Amplitude mismatch. (b) Phase mismatch.

Fig. 4 studies the effect the I/Q imbalance at SRx only, with perfect CSI at STx. Fig. 4a shows the results of fixing the phase mismatch at 10° while varying the amplitude imbalance among 2, 4 and 6 dB, and Fig. 3b illustrates the ABER when fixing the amplitude mismatch to 3 dB while changing the phase imbalance among 0°, 10° and 20°. It can be seen from these figures that the optimal and the WLE receivers have the same performance exactly and they outperform the blind receiver. In addition, the optimal and WLE receivers can resist the amplitude and phase mismatching effect completely. Finally, the blind receiver has an extremely poor performance at high values of the amplitude mismatching. For example, noting the 6 dB amplitude mismatching, 14 dB of gain decreasing between the blind and the other receivers can be seen when ABER equals 0.03.

Fig. 5 illustrates the performance of the receivers in the presence of low, average and high levels of I/Q imbalance at STx/SRx with perfect CSI. This figure emphasizes the conclusions of the previous discussion where the optimal receiver has the best performance and the WLE receiver outperforms the blind receiver at high SNR. Moreover, the optimal and WLE receivers have approximately the same performance at low level I/Q imbalance.

Fig. 6 investigates the effect of channel estimation errors on the ABER, where it is assumed that the channel estimation error has real and imaginary variances equal to the CRLB values that were calculated in Section IV. This figure shows, as expected, how channel estimation errors can degrade the system performance of all the receivers even if it is assumed that the estimator has the best estimated real and imaginary channel values.

Fig. 7 discusses the effect of the number of pilots on the channel estimation error for both perfect and imperfect I/Q

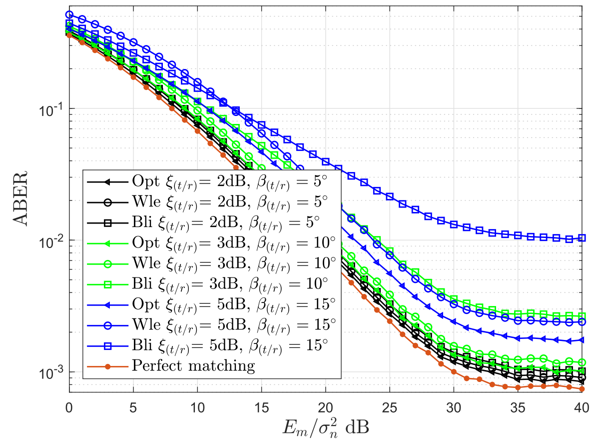


FIGURE 5. ABER in the presence of low, average and high levels of I/Q imbalance at STx/SRx with perfect CSI.

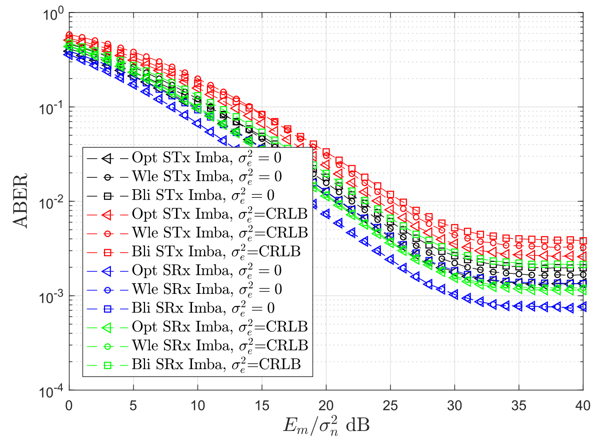


FIGURE 6. ABER at 4 dB amplitude mismatch and 10° phase imbalance at STx or SRx alone with perfect CSI or channel estimation errors that has CRLB variance at SRx.

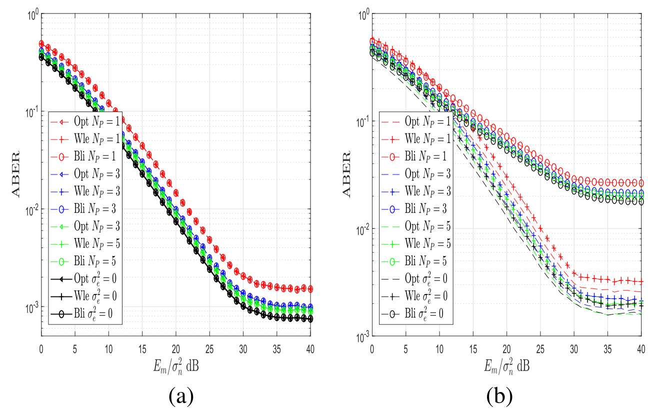
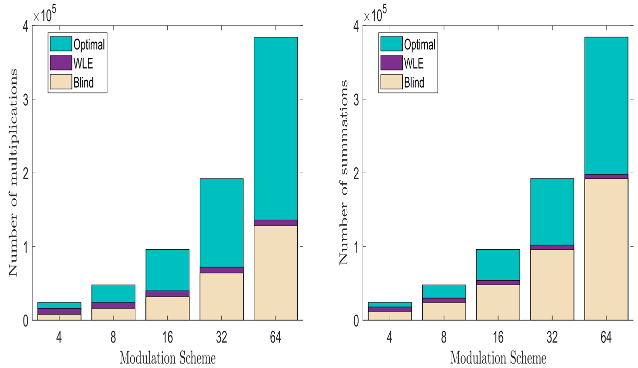


FIGURE 7. Studying the effect of number of pilots on the channel estimation error. The imperfect I/Q matching equals 4 dB amplitude mismatch and 10° phase imbalance at STx/SRx. Both figures 7a and 7b assume that σ_e^2 = CRLB. (a) Perfect I/Q matching. (b) Imperfect I/Q matching.

matching with, receptively. In Fig. 7a, the power gain between the perfect CSI and imperfect CSI for the optimal receiver when using one pilot equals 3 dB at ABER = 0.02, the same



(a) (b)

FIGURE 8. Complexity analysis of the proposed receivers assuming that the frame size is 1000 symbols. (a) Number of multiplications. (b) Number of summations.

power gain can be noticed for each receiver in Fig.7b. This clearly means that the best estimator that reaches the CRLB does not affect by the I/Q imbalance at STx/SRx considering that we normalize the transmitted power to get a fair comparison as it is shown in Appendix B.

Fig. 8 illustrates the computational complexities of the proposed receivers for 4-QAM, 8-QAM, 16-QAM, 32-QAM, and 64-QAM modulation schemes. This figure assumes that the frame time is equal to the coherence time and the number of symbols per frame is 1000. As expected, the blind receiver requires the minimum numbers of real multiplications and summations while the optimal receiver requires the largest values. Another interesting result from this figure is that the WLE receiver has approximately the same complexity as the blind one even though there is a noticeable difference in their performances.

VII. CONCLUSION

This paper presented an analytical framework to study the performance of a CR secondary system that has I/Q imbalance and imperfect CSI. It was shown that the I/Q imbalance affects the system performance and this effect is maximized in case of imperfect CSI. Two receivers were designed: optimal and widely linear receivers, where their performance was compared with the traditional ML receiver using different scenarios. It was proven that the optimal receiver has the best performance, the widely linear one out-performing the traditional one in all scenarios. Moreover, this work calculated the exact PEP of these receivers and APEP for the optimal one. Exact Fisher information matrix (FIM) and CRLB matrices were calculated, proving that CRLB elements are not correlated even if the I/Q imbalance changes the Gaussian noise behavior from proper to improper. The current work has only considered SISO quasi-static scenario, future work can be extended to include multi-in multi-out (MIMO) and time-selective scenarios.

APPENDIX A
PROOF OF LEMMA 1

From (9), and after simple algebraic operations, the received signal y can be written as

$$\begin{aligned} y = &(\sqrt{E}K_1hG_1 + \sqrt{E}K_2h^*G_2^*)x_i + (\sqrt{E}K_1hG_2 \\ &+\sqrt{E}K_2h^*G_1^*)x_i^* + (\sqrt{E}K_1G_1x_i + \sqrt{E}K_1G_2x_i^*)e \\ &+(\sqrt{E}K_2G_2^*x_i + \sqrt{E}K_2G_1^*x_i^*)e^* + K_1n + K_2n^*. \end{aligned} \tag{40}$$

As mentioned, the proposed filter multiplies the received signal y and its conjugate y^* by the scaling factors w_1 and w_2 respectively. The result of this scaling can be given as

$$\begin{aligned} Y &= w_1 \times y + w_2 \times y^* \\ &= \{w_1(K_1hG_1 + K_2h^*G_2^*) + w_2(K_1^*h^*G_2^* + K_2^*hG_1)\}\sqrt{E}x_i \\ &\quad +\{w_1(K_1hG_2 + K_2h^*G_1^*) + w_2(K_1^*h^*G_1^* + K_2^*hG_2)\}\sqrt{E}x_i^* \\ &\quad +\{w_1(K_1G_1x_i + K_1G_2x_i^*) + w_2(K_2^*G_2x_i^* + K_2^*G_1x_i)\}\sqrt{E}e \\ &\quad +\{w_1(K_2G_2^*x_i + K_2G_1^*x_i^*) + w_2(K_1^*G_1^*x_i^* + K_1^*G_2^*x_i)\}\sqrt{E}e^* \\ &\quad + w_1K_1n + w_1K_2n^* + w_2K_1^*n^* + w_2K_2^*n. \end{aligned} \tag{41}$$

In order to cancel the I/Q imbalance and get the transmitted symbol $\sqrt{E}x_i$, the proposed filter matches the resulting signal Y with the transmitted signal $\sqrt{E}x_i$. Consequently, the values of w_1 and w_2 should validate the following:

$$w_1(K_1hG_1 + K_2h^*G_2^*) + w_2(K_1^*h^*G_2^* + K_2^*hG_1) = 1. \tag{42}$$
$$w_1(K_1hG_2 + K_2h^*G_1^*) + w_2(K_1^*h^*G_1^* + K_2^*hG_2) = 0. \tag{43}$$

Solving (42) and (43) to find the values of w_1 and w_2 ends up with

$$w_1 = \frac{\alpha^*}{(\alpha\alpha^* - \beta\beta^*)}, \quad w_2 = \frac{-\beta}{(\alpha\alpha^* - \beta\beta^*)}, \tag{44}$$

where $\alpha = K_1hG_1 + K_2h^*G_2^*$, and $\beta = K_1hG_2 + K_2h^*G_1^*$ Substituting w_1 and w_2 values in (41) ends up with

$$Y = \sqrt{E}x_i + \underbrace{\Omega_1 n + \Omega_2 n^* + \sqrt{E}\Omega_3 e + \sqrt{E}\Omega_4 e^*}_{Z}. \tag{45}$$

where, $\Omega_1 = w_1K_1 + w_2K_2^*$, $\Omega_2 = w_1K_2 + w_2K_1^*$, $\Omega_3 = w_1(K_1G_1x_i + K_1G_2x_i^*) + w_2(K_2^*G_1x_i + K_2^*G_2x_i^*)$, and $\Omega_4 = w_2(K_1^*G_2^*x_i + K_1^*G_1^*x_i^*) + w_1(K_2G_2^*x_i + K_2G_1^*x_i^*)$ Let us rewrite (45) as

$$Y = \sqrt{E}x_i + Z_i^I + jZ_i^Q. \tag{46}$$

Considering that, $\Omega_1 = \Omega_1^I + j\Omega_1^Q$, $\Omega_2 = \Omega_2^I + j\Omega_2^Q$, $\Omega_3 = \sqrt{E}(\Omega_3^I + j\Omega_3^Q)$, and $\sqrt{E}(\Omega_4 = \Omega_4^I + j\Omega_4^Q)$. Z_i^I and Z_i^Q can be written after some mathematical manipulations as

$$\begin{aligned} Z_i^I &= (\Omega_1^I + \Omega_2^I)n^I + (\Omega_2^Q - \Omega_1^Q)n^Q + (\Omega_3^I + \Omega_4^I)\sqrt{E}e^I \\ &\quad + (\Omega_4^Q - \Omega_3^Q)\sqrt{E}e^Q. \\ Z_i^Q &= (\Omega_1^Q + \Omega_2^Q)n^I + (\Omega_1^I - \Omega_2^I)n^Q + (\Omega_3^Q + \Omega_4^Q)\sqrt{E}e^I \\ &\quad + (\Omega_3^I - \Omega_4^I)\sqrt{E}e^Q. \end{aligned} \tag{47}$$

The variances of Z_i^I and Z_i^Q are $\sigma^2_{Z_i^I}$ and $\sigma^2_{Z_i^Q}$ respectively which can be calculated as (note that $\sigma^2_{n^I} = \sigma^2_{n^Q} = \sigma_n^2/2$)

$$\begin{aligned}\sigma^2_{Z_i^I} &= \{(\Omega_1^I + \Omega_2^I)^2 + (\Omega_2^Q - \Omega_1^Q)^2\}\frac{\sigma_n^2}{2} + (\Omega_3^I + \Omega_4^I)^2 E\sigma^2_{e^I} \\ &\quad + (\Omega_4^Q - \Omega_3^Q)^2 E\sigma^2_{e^Q}. \\ \sigma^2_{Z_i^Q} &= \{(\Omega_1^Q + \Omega_2^Q)^2 + (\Omega_1^I - \Omega_2^I)^2\}\frac{\sigma_n^2}{2} + (\Omega_3^Q + \Omega_4^Q)^2 E\sigma^2_{e^I} \\ &\quad + (\Omega_3^I - \Omega_4^I)^2 E\sigma^2_{e^Q}. \end{aligned} \tag{48}$$

It can be seen that, Z_i^I and Z_i^Q are correlated where the correlation factor is given by

$$\rho_{Z_i} = \frac{\mathbb{E}\{Z_i^I Z_i^Q\}}{\sqrt{\sigma^2_{Z_i^I}\sigma^2_{Z_i^Q}}}, \tag{49}$$

where $\mathbb{E}\{Z_i^I Z_i^Q\}$ is given by

$$\begin{aligned}&\mathbb{E}\{Z_i^I Z_i^Q\} \\ &= \{(\Omega_1^I + \Omega_2^I)(\Omega_1^Q + \Omega_2^Q) + (\Omega_2^Q - \Omega_1^Q)(\Omega_1^I - \Omega_2^I)\}\frac{\sigma_n^2}{2} \\ &\quad + E\sigma^2_{e^I}(\Omega_3^I + \Omega_4^I)(\Omega_3^Q + \Omega_4^Q) + E\sigma^2_{e^Q}(\Omega_4^Q - \Omega_3^Q)(\Omega_3^I - \Omega_4^I). \end{aligned} \tag{50}$$

Finally, the received signal Y can be written as

$$Y = \sqrt{E}x_i + Z_i, \tag{51}$$

which concludes the proof.

APPENDIX B
PROOF OF LEMMA 2

To find the CRLB, the received signal in (8) can be written as

$$\begin{aligned} y &= \sqrt{E}h\big(\underbrace{K_1(G_1x_i + G_2{x_i}^*)}_{Q_1}\big) \\ &\quad + \sqrt{E}h^*\big(\underbrace{x_iK_2(G_1x_i + G_2{x_i}^*)^*}_{Q_2}\big) + \underbrace{K_1n + K_2n^*}_{u}. \end{aligned} \tag{52}$$

To simplify the derivation, let us rewrite y as

$$y = \sqrt{E}Ah^I + \sqrt{E}Bh^Q + j(\sqrt{E}Ch^I + \sqrt{E}Dh^Q) + u^I + ju^Q, \tag{53}$$

where, $A = Q_1^I + Q_2^I$, $B = Q_2^Q - Q_1^Q$, $C = Q_2^Q + Q_1^Q$, and $D = Q_1^I - Q_2^I$. It is worthy to note that u is an improper Gaussian noise with $u^I \sim (0, \frac{\sigma_n^2}{2})$, and $u^Q \sim (0, \frac{\sigma_n^2}{2}\xi_r^2)$ with correlation factor $\Psi = -\sin(\beta_r)$

From the previous equation, the joint likelihood function can be written as

$$\begin{aligned} &P_{\bar{y}}(\bar{y}; \theta) \\ &= \Big(\frac{1}{2\pi\sigma_u^I\sigma_u^Q\sqrt{1-\Psi^2}}\Big)^{N_P}\exp\Big(-\frac{1}{2(1-\Psi^2)}\sum_{N=0}^{N_p-1} \\ &\quad \Big[\frac{(y^I - \sqrt{E}Ah^I + \sqrt{E}Bh^Q)^2}{\sigma^2_{u^I}} + \frac{(y^Q - \sqrt{E}Ch^I - \sqrt{E}Dh^Q)^2}{\sigma^2_{u^Q}} \\ &\quad - \frac{2\Psi(y^I - \sqrt{E}Ah^I + \sqrt{E}Bh^Q)(y^Q\sqrt{E}Ch^I - \sqrt{E}Dh^Q)}{\sigma_u^I\sigma_u^Q}\Big]\Big), \end{aligned} \tag{54}$$

where N_P is the number of samples, and $\theta = [h^I \;\; h^Q]^T$ Based on that, the log likelihood function can be written as

$$\begin{aligned} &ln(P_{\bar{y}}(\bar{y}; \theta)) \\ &= -N_P\, ln\big(2\pi\sigma_u^I\sigma_u^Q\sqrt{1-\Psi^2}\big) - \frac{1}{2(1-\Psi^2)}\sum_{N=0}^{N_p-1} \\ &\quad \Big[\frac{(y^I - \sqrt{E}Ah^I - \sqrt{E}Bh^Q)^2}{\sigma^2_{u^I}} + \frac{(y^Q - \sqrt{E}Ch^I - \sqrt{E}Dh^Q)^2}{\sigma^2_{u^Q}} \\ &\quad - \frac{2\Psi\,(y^I - \sqrt{E}Ah^I - \sqrt{E}Bh^Q)(y^Q - \sqrt{E}Ch^I - \sqrt{E}Dh^Q)}{\sigma_u^I\sigma_u^Q}\Big]. \end{aligned} \tag{55}$$

In order to find CRLB, first, we need to obtain the FIM as follows

$$\begin{aligned} &I_{(\theta)}(\theta) \\ &= \begin{bmatrix} I_{(h^I)} = -\mathbb{E}\left\{\frac{\partial^2 ln(P_{\bar{y}}(\bar{y};\theta))}{\partial(h^I)^2}\right\} & I_{(h^I,h^Q)} = -\mathbb{E}\left\{\frac{\partial^2 ln(P_{\bar{y}}(\bar{y};\theta))}{\partial(h^I)(h^Q)}\right\} \\ I_{(h^I,h^Q)} = -\mathbb{E}\left\{\frac{\partial^2 ln(P_{\bar{y}}(\bar{y};\theta))}{\partial(h^I)(h^Q)}\right\} & I_{(h^Q)} = -\mathbb{E}\left\{\frac{\partial^2 ln(P_{\bar{y}}(\bar{y};\theta))}{\partial(h^Q)^2}\right\} \end{bmatrix}. \end{aligned} \tag{56}$$

To find FIM, we find the first, and then the second derivatives as follows

$$\begin{aligned} \frac{\partial ln(P_{\bar{y}}(\bar{y}; \theta))}{\partial(h^I)} &= \sum_{N=0}^{N_p-1}\frac{\sqrt{E}}{(1-\Psi^2)} \\ &\quad \times\Big[\frac{A(y^I - \sqrt{E}Ah^I - \sqrt{E}Bh^Q)}{\sigma^2_{u^I}} \\ &\quad + \frac{C(y^Q - \sqrt{E}Ch^I - \sqrt{E}Dh^Q)}{\sigma^2_{u^Q}} \\ &\quad - \frac{\Psi A(y^Q - \sqrt{E}Ch^I - \sqrt{E}Dh^Q)}{\sigma_u^I\sigma_u^Q} \\ &\quad - \frac{\Psi C(y^I - \sqrt{E}Ah^I - \sqrt{E}Bh^Q)}{\sigma_u^I\sigma_u^Q}\Big]. \end{aligned} \tag{57}$$

$$\frac{\partial^2 ln(P_{\bar{y}}(\bar{y}; \theta))}{\partial(h^I)^2} = \sum_{N=0}^{N_p-1}\frac{E}{(1-\Psi^2)}\Big[-\frac{C^2}{\sigma^2_{u^Q}} - \frac{A^2}{\sigma^2_{u^I}} + \frac{2\Psi AC}{\sigma_u^I\sigma_u^Q}\Big]. \tag{58}$$

From (56) and (58), $I_{(h^I)}$ can be calculated as

$$I_{(h^I)} = \frac{N_pE}{(1-\Psi^2)}\Big[\frac{A^2}{\sigma^2_{u^I}} + \frac{C^2}{\sigma^2_{u^Q}} - \frac{2\Psi AC}{\sigma_u^I\sigma_u^Q}\Big]. \tag{59}$$

Similar to the derivation of $I_{(h^I)}$, $I_{(h^Q)}$ can be derived as

$$I_{(h^Q)} = \frac{N_pE}{(1-\Psi^2)}\Big[\frac{B^2}{\sigma^2_{u^I}} + \frac{D^2}{\sigma^2_{u^Q}} - \frac{2\Psi BD}{\sigma_u^I\sigma_u^Q}\Big]. \tag{60}$$

In the same way, $I_{(h^I,h^Q)} = I_{(h^Q,h^I)}$ can be obtained by

$$I_{(h^I,h^Q)} = \frac{N_p E}{(1-\Psi^2)}\Big[\frac{AB}{\sigma_{u^I}^2} + \frac{CD}{\sigma_{u^Q}^2} - \frac{\Psi AD}{\sigma_u^I \sigma_u^Q} - \frac{\Psi BC}{\sigma_u^I \sigma_u^Q}\Big]. \quad (61)$$

Noting that $K_1^I + K_2^I = 1$, $K_1^Q - K_2^Q = 0$, $G_1^I + G_2^I = 1$, $G_1^Q + G_2^Q = 0$, and $1 - \Psi^2 = \cos(\beta_r)^2$, and after some mathematical simplifications, it can be proven that

$$I_{(h^I)} = I_{(h^Q)} = \frac{2N_p E(x_i^{I^2} + x_i^{Q^2}\xi_t^2 - 2x_i^I x_i^Q \xi_t \sin(\beta_t))}{\sigma_n^2}$$
$$I_{(h^I,h^Q)} = 0 \quad (62)$$

As mentioned in Section VI, for a fair comparison, the transmitted signal energy is normalized by $(|x_i G_1 + x_i^* G_2|^2)$ Observing that, $(|x_i G_1 + x_i^* G_2|^2) = (x_i^{I^2} + x_i^{Q^2}\xi_t^2 - 2x_i^I x_i^Q \xi_t \sin(\beta_t))$, (62) can be rewritten as

$$I_{(h^I)} = I_{(h^Q)} = \frac{2N_p E}{\sigma_n^2}$$
$$I_{(h^I,h^Q)} = 0 \quad (63)$$

From (56) and (63), CRLB matrix can be obtained by finding $\big(I_{(\theta)}(\theta)\big)^{-1}$ as

$$CRLB = \begin{bmatrix} CRLB_{(h^I)} & 0 \\ 0 & CRLB_{(h^Q)} \end{bmatrix}, \quad (64)$$

which concludes the proof.

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