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Impact of I/Q Imbalance on Amplify-and-Forward Relaying: Optimal Detector Design and Error Performance

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***Abstract*—Future wireless communication systems face several transceiver hardware imperfections that may significantly degrade their performance. In-phase (I) and quadrature-phase (Q) imbalance (IQI), which causes self-interference effects on the desired signal, is an important and practical example to these impairments. In this study, a channel state information (CSI)-assisted dual-hop amplify-and-forward (AF) relaying system in the presence of IQI is analyzed. The error performance of the relevant AF cooperative protocol is firstly studied by considering the traditional maximum likelihood detection (MLD) algorithm as a benchmark. Then, two compensation methods, weighting and zero-forcing, are proposed to mitigate IQI effects. Finally, an optimal MLD solution is introduced by adapting the traditional MLD technique in compliance with the asymmetric characteristics of the IQI. System performance is evaluated in terms of average symbol error probability (ASEP) through computer simulations. The ASEP is calculated analytically for the optimal MLD method as well under the assumption of point-to-point (P2P) communication, which has been envisioned as an allied technology of fifth generation (5G) wireless systems, between the source and the relay nodes. A power allocation (PA) algorithm is provided for this specific case. Extensive computer simulations and analytical results prove that the proposed optimal MLD method provides the best results.**

***Index Terms*—AF dual-hop relaying, error performance analysis, I/Q imbalance, optimal ML detection, power allocation.**

I. Introduction

The concept of relaying has found its way into recent standards, such as The Institute of Electrical and Electronics Engineers (IEEE) 802.11s, IEEE 802.16j and the 3$^{\text{rd}}$ generation partnership project long term evolution (3GPP LTE) [1]. Indeed, relaying transmission is one of the promising technologies for future wireless communication networks as well, since it improves the system reliability, extends the network coverage, mitigates channel impairments and ensures high quality of service (QoS) [2], [3]. Relay-assisted communications in wireless networks are more attractive especially when the direct link between the base station and the original mobile terminal is blocked due to a deep fade, heavy shadowing, an intermediate wall or when the destination is out of the source's reach [4]. Thus, the analysis of dual-hop relaying systems appears to be an attractive research area.

The main idea behind the dual-hop relaying technique is that the data signal is transmitted from the source to the destination by separating the channel between them into two links via a relay terminal [5]. Dual-hop transmission systems, based on the nature and complexity of the relays, can be mainly categorized into two popular protocols, namely, regenerative and nonregenerative systems [6]. In regenerative systems, also called decode-and-forward (DF) systems, the relay completely decodes the signal that went through the first hop and retransmits the decoded version into the second hop. In contrasting fashion, nonregenerative systems, also referred to as amplify-and-forward (AF) systems, use less complex relays that just amplify and forward the incoming signal without performing any kind of decoding [7].

The performance of AF and DF relaying systems has been well studied in [8]–[12]. In [8], outage probability for noise limited as well as interference affected systems were obtained, and average bit error rate and outage capacity of AF and DF systems were compared. Low-complexity cooperative diversity protocols that tackle fading in wireless networks, were developed, and it was proven that outage probability can be reduced via various cooperative designs in [9]. In [10], closed form expressions were derived for lower bounds of dual-hop AF relaying performance over generalized gamma distribution that accurately approximates several channel models to identify multi-path, shadowing or composite fading. In [11], bit error and outage probabilities as well as channel capacity of direct-sequence code-division multiple access systems with AF relaying were presented for different fading scenarios. Considering a finite number of co-channel interferers for AF dual-hop systems, the error and outage probabilities were investigated in [12].

For further classification, relays in AF systems can be divided into two sub-categories: blind relays and channel state information (CSI)-assisted relays. The systems that utilize blind relays, which employ fixed gain relaying, do not have the instantaneous CSI at the relay, and result in a signal with variable power at the relay output. However, the systems with such kind of blind relays are not expected to perform as well as the systems equipped with CSI-assisted relays. On the other hand, AF wireless communication systems with CSI-assisted relays use instantaneous CSI of the first hop to fix the power

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of the retransmitted signal [13]. Ergodic capacity performance of CSI-assisted AF cooperative networks with partial relay selection under out-dated CSI is analyzed in [14]. A unified moments-based framework, which is applicable to both one-way and two-way relaying over arbitrary Nakagami-m fading channels, for the general performance analysis of CSI-assisted AF relaying systems was presented in [15].

Since lots of devices are being equipped with wireless capabilities, and the price pressure on wireless products is high, low-cost and flexible solutions are required. The concept of direct-conversion (DC), which is also useful for AF relaying, is promising to fulfill these requirements, since it needs neither external intermediate frequency (IF) filters nor image rejection filters [16]. The majority of the related works in the area of relaying assumes that the transceiver hardware is perfect [9], [17]–[19]. However, in practice, relaying systems, including the ones utilizing DC architecture, are exposed to hardware impairments, e.g., power amplifier nonlinearities, phase noise, in-phase (I) and quadrature-phase (Q) imbalance (IQI) [16]. The impact of the first two imperfections on relaying systems were studied in [20] and [21], respectively. IQI, which refers to the phase and/or amplitude mismatch between the I and Q arms at the transmitter (Tx) and/or receiver (Rx) sides, results in an additional image signal and leads to significant performance loss especially in high-rate wireless communications systems [22]. Moreover, it causes many undesired effects, such as crosstalk, frequency interference and performance degradation [23]. It means that IQI is a critical issue for the next-generation wireless technologies.

Most of the studies done within the context of IQI have focused on the performance analysis and baseband compensation for single-hop systems [24], [25]. Outage performance of half-duplex AF relaying in an orthogonal frequency-division multiplexing (OFDM) system with IQI was analyzed and a low-complexity compensation scheme was presented in [26]. An OFDM dual-hop opportunistic AF relaying system with IQI was studied and a closed-form expression for the outage probability was derived in [27]. The power allocation (PA) problem in an orthogonal frequency-division multiple access system, when the served user equipment (UE) suffers from different levels of IQI was studied in terms of capacity, and a novel low-complexity solution was presented in [28]. The effects of IQI at both source and destination nodes on two-way AF relaying were presented in [29], where lower and upper bounds on the average symbol error probability (ASEP) were provided. The impact of IQI on one-way AF relaying systems was studied in [30], where new compensation algorithms were proposed, and an analytical expression for the ASEP was derived over Rayleigh fading channels, considering dual-hop AF relaying in the presence of IQI at the destination node. However, analytical results were limited to Rayleigh fading channels in [29] and [30], and the authors did not study PA policies to attenuate the effects of IQI. In addition, neither an exact analytical expression for the probability of error was derived, nor the optimal solution of AF relaying in the presence of IQI was not analyzed. Moreover, the asymptotic limits of dual-hop relaying, which are required to offer effective design policies, were not investigated.

In this context, the main contributions of this paper are summarized as follows:

- CSI-assisted AF dual-hop relaying wireless communication systems in the presence of IQI effects at both source and destination nodes are analyzed by using three different detector designs. The destructive effects of only Tx- and only Rx-side IQI on the system performance have also been discussed as special cases.
- Among the utilized detector designs, the optimal MLD has brought an optimum solution for the IQI problem of the future AF relaying systems. The effects of the imperfect CSI case at the destination node is also considered with optimal MLD to obtain more realistic results.
- MLD with compensation framework is designed by applying two different compensation techniques referred to as weighting and zero forcing. The weighting method is proposed as a new alternative to the zero-forcing method which is utilized for compensation procedure of AF relaying in literature.
- The proposed system designs are compared in terms of ASEP and computational complexity. The obtained results prove the effectiveness of the proposed designs on mitigating the effects of IQI. The impact of non-identical variance characteristics of the Rayleigh fading channel on the system performance is investigated as well.
- A closed-form expression of the ASEP is derived for the optimal MLD scheme under the assumption of point-to-point (P2P) connection, which is envisioned as one of the potential transmission scenarios of future 5G networks, between the source and the relay nodes. A power allocation algorithm is proposed for the optimal scheme under this assumption to further improve the QoS.

Organization: The system model of CSI-assisted AF dual-hop relaying under the effect of both Tx- and Rx-side IQI is presented in Section II, and error probability analyses are given in Section III for each Rx design. The symbol error probability (SEP) analyses for the special cases of only source/destination IQI are included in Section III as well. Average and asymptotic error probability for P2P connected source and relay case are presented in Section IV, and PA parameters are optimized for this special case in Section V. Complexity analysis is presented in Section VI, while numerical results are provided in Section VII. Finally, the study is concluded in Section VIII.

Notations: Bold lower and upper case letters denote vectors and matrices, respectively. $(\cdot)^T$ and $(\cdot)^{-1}$ indicate matrix transpose and matrix inverse operations. $\mathcal{N}(\mu, \sigma^2)$ represents real Gaussian distribution with mean μ and variance σ^2. $(\cdot)^I$ and $(\cdot)^Q$ denote the I and Q components. $(\cdot)^*$ is the complex conjugate, while $\mathbb{E}\{\cdot\}$ denoting the expectation. $\Re\{z\}$ stands for the real part of the complex variable z, and $\Pr\{\cdot\}$ denotes the probability of an event.

II. SYSTEM MODEL

A dual-hop AF relaying network model, in which the source and the destination nodes communicate with each other via a relay node, is considered in this work as illustrated in Fig. 1. All nodes are equipped with a single antenna. It is

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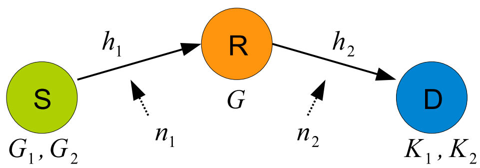

Fig. 1. System model of AF dual-hop relaying in the presence of IQI.

assumed that there is no direct link between the source and the destination, i.e., the source communicates with the destination only through the relay terminal, due to deep fading, blockage or heavy shadowing. The data transmission is realized in two phases.

In first phase, information is conveyed from the source to the relay. The relay amplifies the received signal and broadcasts it to the destination in the second phase. In this paper, the radio-frequency (RF) front-ends of the source and the destination are assumed to be affected by the IQI impairments, while the relay transmission is perfect. In addition, h_1 and h_2 are mutually-independent and non-identical Rayleigh fading channel coefficients assumed to be remaining constant during these two phases. The transceiver is affected by additive white Gaussian noise, where the noise terms, n_1 and n_2, have zero mean and equal variance of σ_n^2 ($\sigma_{n_1}^2 = \sigma_{n_2}^2 = \sigma_n^2$).

In ideal case, I and Q parts of the local oscillator (LO) signals should have exactly the same amplitude and 90° phase difference [23]. However, in any practical system, including the cooperative networks that use DC architecture, perfect matching between the I and Q branches of the quadrature mixer is not possible due to limited accuracy of the analog components, such as capacitors and resistors [31]. On the Tx side, this impairment can occur at the I/Q up-conversion step beside filters and digital-to-analog converters. On the Rx side, it can emerge at the down-conversion step as well as amplification and sampling stages. For simplicity, all mismatches are referred to the I/Q up- and down-conversion steps, which is typical in the literature in this context [22], [24], [31]. Considering both Tx and Rx IQI effects, the corresponding complex LO signals for Tx and Rx sides can be written as follows, respectively [31]:

$$\begin{aligned} z_T(t) &= \cos(\omega_L t) + j\xi_T \sin(\omega_L t + \phi_T) \\ &= G_1 e^{j\omega_L t} + G_2 e^{-j\omega_L t}, \end{aligned} \tag{1}$$

$$\begin{aligned} z_R(t) &= \cos(\omega_L t) - j\xi_R \sin(\omega_L t + \phi_R) \\ &= K_1 e^{-j\omega_L t} + K_2 e^{j\omega_L t}, \end{aligned} \tag{2}$$

where $\omega_L = 2\pi f_L$, while f_L is the LO frequency; $\{\xi_T, \phi_T\}$ and $\{\xi_R, \phi_R\}$ denote the total effective amplitude and phase imbalances of the Tx and Rx sides, respectively. IQI parameters of Tx side (G_1, G_2) and Rx side (K_1, K_2) are given by using (1) and (2) in the following form

$$G_1 = \frac{1}{2}(1 + \xi_T e^{j\phi_T}), \quad G_2 = \frac{1}{2}(1 - \xi_T e^{-j\phi_T}), \tag{3}$$

$$K_1 = \frac{1}{2}(1 + \xi_R e^{-j\phi_R}), \quad K_2 = \frac{1}{2}(1 - \xi_R e^{j\phi_R}). \tag{4}$$

Note that for perfect I/Q matching, IQI parameters reduce to $\xi_T = \xi_R = 1$ and $\phi_T = \phi_R = 0°$. Therefore, in this case, it is clear that $G_1 = K_1 = 1$, $G_2 = K_2 = 0$.

An asymmetrical IQI structure, where the I branch is assumed ideal and the errors are modeled in the Q branch [16], [24], is considered in this study. In this case, the baseband representation of the up-converted signal in the presence of Tx IQI at the source terminal is given as

$$x_k^{IQ} = G_1 x_k + G_2^* x_k^*. \tag{5}$$

Here, it is assumed that M-QAM modulation scheme is utilized (M is the modulation order of complex constellations), x_k is the modulated baseband Tx signal under perfect I/Q matching, where $k \in \{1, ..., M\}$ [22].

In phase 1, information is transmitted from the source node to the relay terminal. Considering Tx IQI effect at the source, the received baseband signal at the relay node is given as

$$y_r = \sqrt{E_s} h_1 (G_1 x_k + G_2^* x_k^*) + n_1, \tag{6}$$

where E_s is the average energy per transmitted symbol at the source node. Here, h_1 represents the gain coefficient of the channel between the source and the relay. In phase 2, the ideal CSI-assisted relay amplifies the received signal and then broadcasts it to the destination node. The received signal at the destination can be written as

$$y_d = \sqrt{E_r} h_2 G(\sqrt{E_s} h_1 G_1 x_k + \sqrt{E_s} h_1 G_2^* x_k^* + n_1) + n_2, \tag{7}$$

where E_r is the transmitting energy of the relay, h_2 represents the gain coefficient of the channel between the relay and the destination, and G is the relay gain that given as

$$G = \frac{1}{\sqrt{E_s |h_1|^2 (|G_1|^2 + |G_2|^2)}}. \tag{8}$$

Considering the Rx IQI effect on the received signal as well, the resulting signal at the destination node is given as

$$\begin{aligned} y_d^{IQ} =& K_1 y_d + K_2 y_d^* \\ =& \sqrt{E_s E_r} G (G_1 K_1 h_1 h_2 + G_2 K_2 h_1^* h_2^*) x_k \\ &+ \sqrt{E_s E_r} G (G_2^* K_1 h_1 h_2 + G_1^* K_2 h_1^* h_2^*) x_k^* \\ &+ \sqrt{E_r} G (K_1 h_2 n_1 + K_2 h_2^* n_1^*) + K_1 n_2 + K_2 n_2^* \\ =& \sqrt{E_r} (\Lambda x_k + \Omega x_k^*) + \tilde{n} \\ =& \sqrt{E_r} \Upsilon_k + \tilde{n}, \end{aligned} \tag{9}$$

where $\Lambda = \sqrt{E_s} G (G_1 K_1 h_1 h_2 + G_2 K_2 h_1^* h_2^*)$, $\Omega = \sqrt{E_s} G (G_2^* K_1 h_1 h_2 + G_1^* K_2 h_1^* h_2^*)$, $\Upsilon_k = \Lambda x_k + \Omega x_k^*$ and $\tilde{n} = \sqrt{E_r} G (K_1 h_2 n_1 + K_2 h_2^* n_1^*) + K_1 n_2 + K_2 n_2^*$. As seen from (9), IQI causes a self-interference effect, Ωx_k^*, beside distorting the signal. In order to assess the destructive effect of this self-interference, assuming that symbols have unit-energy, average signal-to-interference ratio (SIR) can be calculated as follows by using (9)

$$\begin{aligned} \text{SIR} &= \frac{\mathbb{E}\{|\sqrt{E_s E_r} G (G_1 K_1 h_1 h_2 + G_2 K_2 h_1^* h_2^*)|^2\}}{\mathbb{E}\{|\sqrt{E_s E_r} G (G_2^* K_1 h_1 h_2 + G_1^* K_2 h_1^* h_2^*)|^2\}} \\ &= \frac{|G_1|^2 |K_1|^2 + |G_2|^2 |K_2|^2}{|G_2|^2 |K_1|^2 + |G_1|^2 |K_2|^2}. \end{aligned} \tag{10}$$

In the case of perfect I/Q matching, SIR has the ideal value of SIR $= \infty$ in accordance with (10) as seen in Fig. 2. In contrasting fashion, even very small values of imbalance in

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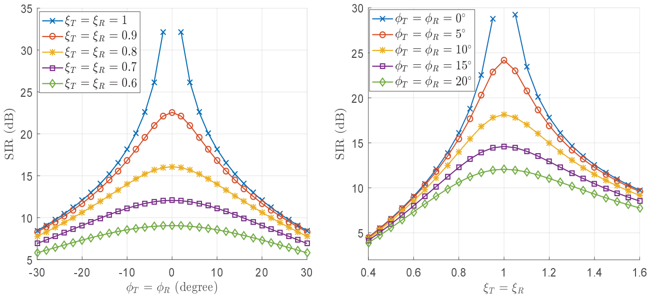

Fig. 2. SIR values given for IQI affected AF relaying system with fixed values of amplitude imbalance (left) and fixed values of phase imbalance (right).

gain, such as $\xi_T = \xi_R = 0.05$, or phase, such as $\phi_T = \phi_R = 2°$, cause a remarkable degradation in the SIR value. Thus, IQI needs to be considered as an important drawback of AF dual-hop relaying communication systems.

As a special case, and only in the presence of Tx IQI, the resulting signal is obtained by considering the ideal values of the Rx IQI parameters, i.e., $K_1 = 1$ and $K_2 = 0$, as follows

$$\begin{aligned} y_{d,T}^{IQ} =& \sqrt{E_s E_r} G(G_1 h_1 h_2) x_k + \sqrt{E_s E_r} G(G_2^* h_1 h_2) x_k^* \\ &+ \sqrt{E_r} G h_2 n_1 + n_2 \\ =& \sqrt{E_r} \Upsilon_{k,T} + \tilde{n}_T, \end{aligned} \tag{11}$$

where $\Upsilon_{k,T} = \sqrt{E_s} G(G_1 h_1 h_2 x_k + G_2^* h_1 h_2 x_k^*)$ and $\tilde{n}_T = \sqrt{E_r} G h_2 n_1 + n_2$. Similarly, the resulting signal only in the presence of Rx IQI is determined by substituting the ideal values of the Tx IQI parameters in (9), and given as

$$\begin{aligned} y_{d,R}^{IQ} =& \sqrt{E_s E_r} G_r (K_1 h_1 h_2) x_k + \sqrt{E_s E_r} G_r (K_2 h_1^* h_2^*) x_k^* \\ &+ \sqrt{E_r} G_r (K_1 h_2 n_1 + K_2 h_2^* n_1^*) + K_1 n_2 + K_2 n_2^* \\ =& \sqrt{E_r} \Upsilon_{k,R} + \tilde{n}_R. \end{aligned} \tag{12}$$

Here, $\Upsilon_{k,R} = \sqrt{E_s} G_r (K_1 h_1 h_2 x_k + K_2 h_1^* h_2^* x_k^*)$ and $\tilde{n}_R = \sqrt{E_r} G_r (K_1 h_2 n_1 + K_2 h_2^* n_1^*) + K_1 n_2 + K_2 n_2^*$, while G_r being the customized gain that is obtained by considering the ideal values of Tx IQI parameters in (8) as well, and is equal to

$$G_r = \frac{1}{\sqrt{E_s |h_1|^2}}. \tag{13}$$

Note that, in the case of ideal I/Q matching at both Tx and Rx sides, the resulting signal will be obtained as follows

$$y_{d,P} = \sqrt{E_r} h_2 G_r (\sqrt{E_s} h_1 x_k + n_1) + n_2 = \sqrt{E_r} \Upsilon_{k,P} + \tilde{n}_P, \tag{14}$$

where $\Upsilon_{k,P} = \sqrt{E_s} h_1 h_2 G_r x_k$ and $\tilde{n}_P = \sqrt{E_r} h_2 G_r n_1 + n_2$.

III. Detector Designs and Error Analyses

The proposed detector designs are defined in this section. It is assumed that the IQI parameters are known at the Rx-side during the formulation of all designs.

A. Traditional MLD

An Rx scheme is introduced by using the traditional MLD method, as similar to [32] and [33], for the CSI-assisted AF relaying transmission. Considering that y_d^{IQ}, which has two zero-mean identical additive white Gaussian noise constituents, is a Gaussian random variable (RV), and using the probability density function (PDF)[1] of Gaussian RVs, the decision rule for the traditional MLD technique can be written by using (9) as

$$\begin{aligned} \hat{x}_k &= \arg\max_k \{ f_{y_d^{IQ}} (y_d^{IQ} | x_k, G_1, G_2, K_1, K_2) \} \\ &= \arg\min_k \left\{ |y_d^{IQ} - \sqrt{E_r} \Upsilon_k|^2 \right\}. \end{aligned} \tag{15}$$

Assuming the symbol x_k has been transmitted from the source; however, $\hat{x}_k$ is erroneously detected at the Rx side, the conditional SEP can be calculated as follows

$$\mathrm{Ps}_t = \Pr \left\{ |y_d^{IQ} - \sqrt{E_r} \Upsilon_k|^2 > |y_d^{IQ} - \sqrt{E_r} \hat{\Upsilon}_k|^2 \right\}, \tag{16}$$

where $\hat{\Upsilon}_k = \Lambda \hat{x}_k + \Omega \hat{x}_k^*$. Following this with several mathematical operations, (16) can be given in the form of

$$\begin{aligned} \mathrm{Ps}_t =& \Pr \left\{ |\tilde{n}|^2 > |\sqrt{E_r} (\Upsilon_k - \hat{\Upsilon}_k) + \tilde{n}|^2 \right\} \\ =& \Pr \left\{ -2\Re\{ \sqrt{E_r} (\Upsilon_k - \hat{\Upsilon}_k) \tilde{n}^* \} > E_r |\Upsilon_k - \hat{\Upsilon}_k|^2 \right\}. \end{aligned} \tag{17}$$

Given that $N_t = -2\Re\{ \sqrt{E_r} (\Upsilon_k - \hat{\Upsilon}_k) \tilde{n}^* \}$, the variance of N_t is calculated as follows

$$\begin{aligned} \sigma_{N_t}^2 =& 4E_r \{ (\Upsilon_k^I - \hat{\Upsilon}_k^I)^2 \sigma_{\tilde{n}^I}^2 + (\Upsilon_k^Q - \hat{\Upsilon}_k^Q)^2 \sigma_{\tilde{n}^Q}^2 \\ &+ 2\rho \sigma_{\tilde{n}^I} \sigma_{\tilde{n}^Q} (\Upsilon_k^I - \hat{\Upsilon}_k^I)(\Upsilon_k^Q - \hat{\Upsilon}_k^Q) \}, \end{aligned} \tag{18}$$

where $\sigma_{\tilde{n}^I}^2$ and $\sigma_{\tilde{n}^Q}^2$ are the variances of the real and imaginary parts of $\tilde{n}$, respectively, and given by

$$\begin{aligned} \sigma_{\tilde{n}^I}^2 &= \frac{E_r |h_2|^2 \sigma_{n_1}^2}{2E_s |h_1|^2 (|G_1|^2 + |G_2|^2)} + \frac{\sigma_{n_2}^2}{2}, \\ \sigma_{\tilde{n}^Q}^2 &= \xi_R^2 \sigma_{\tilde{n}^I}^2. \end{aligned} \tag{19}$$

In addition to these, ρ is the correlation coefficient between $\tilde{n}^I$ and $\tilde{n}^Q$, and it is equal to

$$\rho = \frac{\mathbb{E}\{\tilde{n}^I \tilde{n}^Q\}}{\sqrt{\sigma_{\tilde{n}^I}^2 \sigma_{\tilde{n}^Q}^2}} = \frac{(K_1^Q + K_2^Q) \sigma_{\tilde{n}^I}^2}{\xi_R \sigma_{\tilde{n}^I}^2} = -\sin\phi_R. \tag{20}$$

Conditional SEP is derived by using (17), (18), and the Q-function[2] as presented in (21). However, it is relatively difficult, if not impossible, to derive the ASEP expression by using (21). Hence, the ASEP is calculated numerically during the computer simulations.

It is worth pointing out that, although the Rx is aware of the amplitude and phase mismatches, this design can not provide an optimal solution, which is explained as follows. Even if the noise part in (9), $\tilde{n}$, has improper characteristics, which will be explained in Subsection C, it has not been taken into account while formulating this MLD rule.

The conditional SEP in the presence of Tx IQI only case can be determined by changing Υ_k and $\hat{\Upsilon}_k$ values in (21) to $\Upsilon_{k,T}$ and $\hat{\Upsilon}_{k,T}$, respectively ($\hat{\Upsilon}_{k,T} = \sqrt{E_s} G(G_1 h_1 h_2 \hat{x}_k +$

[1] PDF of a Gaussian RV x, which has mean μ and variance σ^2, is given as $f(x|\mu, \sigma^2) = \frac{1}{\sqrt{2\pi\sigma^2}} e^{-(x-\mu)^2/2\sigma^2}$.

[2] $Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-t^2/2} dt$.

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$$\mathrm{Ps}_t = Q\left(\sqrt{\frac{E_r|\Upsilon_k-\hat{\Upsilon}_k|^4}{4\{(\Upsilon_k^I-\hat{\Upsilon}_k^I)^2\sigma^2_{\tilde{n}_I}+(\Upsilon_k^Q-\hat{\Upsilon}_k^Q)^2\sigma^2_{\tilde{n}_Q}+2\rho\sigma_{\tilde{n}_I}\sigma_{\tilde{n}_Q}(\Upsilon_k^I-\hat{\Upsilon}_k^I)(\Upsilon_k^Q-\hat{\Upsilon}_k^Q)\}}}\right). \quad (21)$$

$G_2^* h_1 h_2 \hat{x}_k^*)$. Considering $\rho = 0$ and $\sigma^2_{\tilde{n}^I} = \sigma^2_{\tilde{n}^Q}$, (21) can be redefined for this special case after some simplifications by

$$\mathrm{Ps}_{t_{\mathrm{Tx}}} = Q\left(\sqrt{\frac{E_r|\Upsilon_{k,T}-\hat{\Upsilon}_{k,T}|^2}{4\sigma^2_{\tilde{n}^I}}}\right). \quad (22)$$

Similarly, the conditional SEP of the special case of only Rx IQI existence for traditional MLD method can easily be obtained by replacing Υ_k, $\hat{\Upsilon}_k$ and $\sigma^2_{\tilde{n}^I}$ by $\Upsilon_{k,R}$, $\hat{\Upsilon}_{k,R}$ and $\sigma^2_{\tilde{n}_R^I}$ (the variance of the real part of $\tilde{n}_R$) in (21), respectively ($\hat{\Upsilon}_{k,R} = \sqrt{E_s} G_r (K_1 h_1 h_2 \hat{x}_k + K_2 h_1^* h_2^* \hat{x}_k^*)$).

B. MLD with Compensation

A linear Rx is designed by applying two different compensation techniques referred to as weighting and zero forcing, in this section. The main purposes of compensation are cancelling the self-interference signal and detecting the symbol more accurately.

Although zero-forcing is a known linear method, there are only several studies which utilize it as a compensation method for AF relaying in the presence of I/Q imbalance [29], [30], [34]. On the other hand, to the best of authors' knowledge, there is no study that uses weighting as a compensation technique for IQI-affected AF relaying systems in literature. Hence, the weighting method is proposed as a new alternative to the zero-forcing for the compensation algorithms. These two structures are introduced in the following subsections separately.

1) Weighting: In order to cancel the interference effect caused by IQI in (9), two weights, w_1 and w_2, are utilized in this method. One of these weights is multiplied by y_d^{IQ} and the other one is multiplied by its conjugate, $(y_d^{IQ})^*$. As a result, a new received signal is obtained as $y = w_1 y_d^{IQ} + w_2 (y_d^{IQ})^*$, and it can be given in an expanded form as follows

$$\begin{aligned} y =& \sqrt{E_r}(\Lambda w_1 + \Omega^* w_2) x_k + \sqrt{E_r}(\Omega w_1 + \Lambda^* w_2) x_k^* \\ &+ \sqrt{E_r} G(\gamma h_2 n_1 + \beta h_2^* n_1^*) + \gamma n_2 + \beta n_2^*, \end{aligned} \quad (23)$$

where $\gamma = w_1 K_1 + w_2 K_2^*$ and $\beta = w_1 K_2 + w_2 K_1^*$. Then, in order to cancel the IQI effect, w_1 and w_2 should be defined on the condition that

$$\sqrt{E_r}(\Lambda w_1 + \Omega^* w_2) = 1 \text{ and } \sqrt{E_r}(\Omega w_1 + \Lambda^* w_2) = 0. \quad (24)$$

According to the condition in (24), w_1 and w_2 are given by

$$\begin{aligned} w_1 &= \frac{\Lambda^* \sqrt{|G_1|^2+|G_2|^2}}{\sqrt{E_r |h_1|^2}|h_2|^2(|G_2|^2-|G_1|^2)(|K_2|^2-|K_1|^2)}, \\ w_2 &= \frac{-\Omega \sqrt{|G_1|^2+|G_2|^2}}{\sqrt{E_r |h_1|^2}|h_2|^2(|G_2|^2-|G_1|^2)(|K_2|^2-|K_1|^2)}. \end{aligned} \quad (25)$$

Substituting (25) into (23), the self-interference signal is eliminated. Now, y can be written in the form of $y = x_k + N$. Note that y is a Gaussian RV and N is equal to

$$\begin{aligned} N =& \frac{1}{\sqrt{E_s}(|G_2|^2-|G_1|^2)}\left(\frac{G_2^*}{h_1^*}n_1^* - \frac{G_1^*}{h_1}n_1\right) \\ &+ \frac{\sqrt{|G_1|^2+|G_2|^2}}{\sqrt{E_r}(|G_2|^2-|G_1|^2)|h_1|}\left(\frac{G_2^* h_1}{h_2^*}n_2^* - \frac{G_1^* h_1^*}{h_2}n_2\right). \end{aligned} \quad (26)$$

Afterwards, a decision rule can be constituted by using the Gaussian PDF and traditional MLD method as

$$\begin{aligned} \hat{x}_k &= \arg\max_k \{f_y(y|x_k, G_1, G_2, K_1, K_2)\} \\ &= \arg\min_k \left\{|y - x_k|^2\right\}. \end{aligned} \quad (27)$$

Conditional SEP can be calculated for the scenario that symbol x_k has been sent from the source; however, symbol $\hat{x}_k$ is detected erroneously at the Rx side by using (27) as

$$\begin{aligned} \mathrm{Ps}_c &= \Pr\left\{|y - x_k|^2 > |y - \hat{x}_k|^2\right\} \\ &= \Pr\left\{-2\Re\{(x_k - \hat{x}_k)N^*\} > |x_k - \hat{x}_k|^2\right\}. \end{aligned} \quad (28)$$

Given that $N_c = -2\Re\{(x_k - \hat{x}_k)N^*\}$, the variance of N_c is calculated as follows

$$\begin{aligned} \sigma^2_{N_c} =& 4\{(x_k^I - \hat{x}_k^I)^2 \sigma^2_{N^I} + (x_k^Q - \hat{x}_k^Q)^2 \sigma^2_{N^Q} \\ &+ 2\rho_c \sigma_{N^I}\sigma_{N^Q}(x_k^I - \hat{x}_k^I)(x_k^Q - \hat{x}_k^Q)\}, \end{aligned} \quad (29)$$

where ρ_c is the correlation coefficient between the real and imaginary parts of N, and it is calculated by

$$\rho_c = \frac{\mathbb{E}\{N^I N^Q\}}{\sqrt{\sigma^2_{N^I}\sigma^2_{N^Q}}} = \frac{2(G_1^I G_2^Q + G_1^Q G_2^I)}{\sqrt{(|G_1|^2+|G_2|^2)^2 - 4(G_1^Q G_2^Q - G_1^I G_2^I)^2}}. \quad (30)$$

Additionally, $\sigma^2_{N^I}$ and $\sigma^2_{N^Q}$ are the variances of the real and imaginary components of N that are obtained as follows

$$\begin{aligned} \sigma^2_{N^I} &= \Theta[|G_1|^2 + |G_2|^2 + 2(G_1^Q G_2^Q - G_1^I G_2^I)] \\ \sigma^2_{N^Q} &= \Theta[|G_1|^2 + |G_2|^2 - 2(G_1^Q G_2^Q - G_1^I G_2^I)], \end{aligned} \quad (31)$$

where $\Theta = \{\frac{\sigma^2_{n_1}}{E_s|h_1|^2} + \frac{(|G_1|^2+|G_2|^2)\sigma^2_{n_2}}{E_r|h_2|^2}\}/2(|G_2|^2 - |G_1|^2)^2$.

In light of this information, (28) can be expressed as given in (32), by utilizing the Q-function. Since defining a closed form of ASEP, considering (32), is not possible, it can be calculated numerically as well for this method.

The conditional SEP for the only Tx IQI case is equal to (32) as well, since there is no term related to Rx IQI in this equation. This means that Rx IQI has no effect on the SEP performance of the weighting compensation method thanks to the conditions of (24). Meanwhile, the conditional SEP of the only Rx IQI case can be simply defined by substituting $G_1 = 1$ and $G_2 = 0$ in (30) and (31). In this manner, $\sigma^2_{N^I}$ and $\sigma^2_{N^Q}$ are obtained as equal to $\sigma^2_{N_R} = \frac{1}{2}\{\frac{\sigma^2_{n_1}}{E_s|h_1|^2} + \frac{\sigma^2_{n_2}}{E_r|h_2|^2})\}$, while ρ_c is acquired as zero. Substituting these values in (32) ends up with

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$$\mathrm{Ps}_c = Q\left(\sqrt{\frac{|x_k-\hat{x}_k|^4}{4\{(x_k^I-\hat{x}_k^I)^2\sigma^2_{N^I}+(x_k^Q-\hat{x}_k^Q)^2\sigma^2_{N^Q}+2\rho_c\sigma_{N^I}\sigma_{N^Q}(x_k^I-\hat{x}_k^I)(x_k^Q-\hat{x}_k^Q)\}}}\right). \quad (32)$$

$$\mathrm{Ps}_{c_{\mathrm{Tx}}} = Q\left(\sqrt{\frac{E_sE_r|h_1|^2|h_2|^2|x_k-\hat{x}_k|^2}{2(E_r|h_2|^2\sigma^2_{n_1}+E_s|h_1|^2\sigma^2_{n_2})}}\right). \quad (33)$$

This equation is equal to the instantaneous SEP of the perfect case. This means that weighting technique is totally able to cancel the Rx IQI effect.

2) Zero-Forcing: The main idea of this compensation technique stems from the fact that the received signal, y_d^{IQ}, which is affected by IQI, and its conjugate, $(y_d^{IQ})^*$, can be written in a matrix form to detect the transmitted signal, and the self-interference signal can be cancelled by applying several matrix operations. In accordance with this, first, the resulting signal at the destination node is rewritten in a matrix form as

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{M}\mathbf{n_1} + \mathbf{K}\mathbf{n_2}, \quad (34)$$

where $\mathbf{y}=\begin{bmatrix} y_d^{IQ} & (y_d^{IQ})^* \end{bmatrix}^T$, $\mathbf{H}=\sqrt{E_r}\begin{bmatrix}\Lambda & \Omega \\ \Omega^* & \Lambda^*\end{bmatrix}$, $\mathbf{x}=\begin{bmatrix}x_k \\ x_k^*\end{bmatrix}$, $\mathbf{M}=\sqrt{E_r}G\begin{bmatrix}K_1h_1 & K_2h_2^* \\ K_2^*h_2 & K_1^*h_2^*\end{bmatrix}$, $\mathbf{n_z}=\begin{bmatrix}n_z & n_z^*\end{bmatrix}^T$ $(z=\{1,2\})$ and $\mathbf{K}=\begin{bmatrix}K_1 & K_2 \\ K_2^* & K_1^*\end{bmatrix}$. Then, $\mathbf{y}$ is multiplied by $\mathbf{H}^{-1}$ and this ends up with

$$\mathbf{H}^{-1}\mathbf{y} = \mathbf{x} + \tilde{\mathbf{n}}. \quad (35)$$

Here, $\tilde{\mathbf{n}} = \mathbf{H}^{-1}\mathbf{M}\mathbf{n}_1 + \mathbf{H}^{-1}\mathbf{K}\mathbf{n}_2 = \begin{bmatrix}N & N^*\end{bmatrix}^T$ is defined as the effective noise, and $\mathbf{H}^{-1}$ is obtained as follows

$$\mathbf{H}^{-1}=\frac{|G_1|^2+|G_2|^2}{\sqrt{E_r}|h_2|^2(|G_1|^2-|G_2|^2)(|K_1|^2-|K_2|^2)}\begin{bmatrix}\Lambda & \Omega \\ \Omega^* & \Lambda^*\end{bmatrix}. \quad (36)$$

Denoting $\mathbf{H}^{-1}\mathbf{y}=\begin{bmatrix}y & y^*\end{bmatrix}^T$ in (35), it can be written as

$$\begin{bmatrix}y \\ y^*\end{bmatrix} = \begin{bmatrix}x_k \\ x_k^*\end{bmatrix} + \begin{bmatrix}N \\ N^*\end{bmatrix}. \quad (37)$$

Hence, y can be expressed as $y = x_k + N$, where N is obtained exactly the same as (26) after a few matrix operations. Therefore, the traditional MLD rule given in (27) is also valid for zero-forcing compensation technique and the conditional SEP can be calculated as given in (32) as well. Cancelling the self-interence effect on the received signal removes the IQI effect on the symbol for both weighting and zero-forcing, so providing the same performance in terms of SEP is an expected result. The SEP definitions for only Tx and only Rx cases as well as the related issues are also applicable for the zero-forcing method. However, the way of avoiding the effect of IQI is entirely different between these methods, and zero-forcing is more advantageous than weighting in terms of complexity, which will be discussed later on.

C. Optimal MLD

Thus far, MLD rule is built-up ignoring the improper characteristics of the noise part, $\tilde{n}$, in (9), and utilizing the PDF of a circularly-symmetric complex Gaussian noise, i.e., the real and imaginary parts are independent and have equal variances. On the other hand, improper characteristics come into play for a complex Gaussian RV on three conditions: i) real and imaginary parts are correlated, ii) real and imaginary parts do not have the same variance (non-identical), or iii) real and imaginary parts are both correlated and non-identical. It is proven in (19) that $\tilde{n}$ has non-identical components, i.e., $\sigma^2_{\tilde{n}^I} \neq \sigma^2_{\tilde{n}^Q}$. Despite the fact that this proof is enough to understand that $\tilde{n}$ is an improper Gaussian noise (IGN) sample, it is also known that the correlation coefficient ρ is not equal to zero ($\rho = -\sin\phi_R$), which means the real and imaginary parts of $\tilde{n}$ are correlated as well. Therefore, modeling the effects of IQI as an IGN allows us to analyze the asymmetric characteristics of the IQI accurately [35]. As it is proved that utilizing the improperness of the Gaussian noise using MLD method leads to a decrease in the error probability [23], [36], the optimal MLD scheme is designed by considering this phenomenon.

The joint multi-variate PDF of the received signal at the destination node, which is under the effect of both Tx and Rx IQI, can be expressed as follows

$$\begin{aligned}&f_{(y_d^{IQ})^I,(y_d^{IQ})^Q}((y_d^{IQ})^I,(y_d^{IQ})^Q|x_k,G_1,G_2,K_1,K_2)\\ &=\frac{1}{2\pi\sigma_{\tilde{n}^I}\sigma_{\tilde{n}^Q}\sqrt{1-\rho^2}}\times\exp\left(-\frac{1}{2(1-\rho^2)}\right.\\ &\times\left[\frac{((y_d^{IQ})^I-\sqrt{E_r}\Upsilon_k^I)^2}{\sigma^2_{\tilde{n}^I}}+\frac{((y_d^{IQ})^Q-\sqrt{E_r}\Upsilon_k^Q)^2}{\sigma^2_{\tilde{n}^Q}}\right.\\ &\left.\left.-\frac{2\rho((y_d^{IQ})^I-\sqrt{E_r}\Upsilon_k^I)((y_d^{IQ})^Q-\sqrt{E_r}\Upsilon_k^Q)}{\sigma_{\tilde{n}^I}\sigma_{\tilde{n}^Q}}\right]\right). \quad (38)\end{aligned}$$

Hence, the optimal MLD decision rule is given as

$$\begin{aligned}\hat{x}_k &=\arg\max_k\{f_{(y_d^{IQ})^I,(y_d^{IQ})^Q}((y_d^{IQ})^I,(y_d^{IQ})^Q)\}\\ &=\arg\min_k\left\{\frac{((y_d^{IQ})^I-\sqrt{E_r}\Upsilon_k^I)^2}{\sigma^2_{\tilde{n}^I}}+\frac{((y_d^{IQ})^Q-\sqrt{E_r}\Upsilon_k^Q)^2}{\sigma^2_{\tilde{n}^Q}}\right.\\ &\left.-\frac{2\rho((y_d^{IQ})^I-\sqrt{E_r}\Upsilon_k^I)((y_d^{IQ})^Q-\sqrt{E_r}\Upsilon_k^Q)}{\sigma_{\tilde{n}^I}\sigma_{\tilde{n}^Q}}\right\}. \quad (39)\end{aligned}$$

The conditional SEP is calculated by using (38) and (39) as follows

$$\begin{aligned}\mathrm{Ps}_o=\Pr&\left\{\frac{((y_d^{IQ})^I-\sqrt{E_r}\Upsilon_k^I)^2}{\sigma^2_{\tilde{n}^I}}+\frac{((y_d^{IQ})^Q-\sqrt{E_r}\Upsilon_k^Q)^2}{\sigma^2_{\tilde{n}^Q}}\right.\\ &-\frac{2\rho((y_d^{IQ})^I-\sqrt{E_r}\Upsilon_k^I)((y_d^{IQ})^Q-\sqrt{E_r}\Upsilon_k^Q)}{\sigma_{\tilde{n}^I}\sigma_{\tilde{n}^Q}}>\\ &\frac{((y_d^{IQ})^I-\sqrt{E_r}\hat{\Upsilon}_k^I)^2}{\sigma^2_{\tilde{n}^I}}+\frac{((y_d^{IQ})^Q-\sqrt{E_r}\hat{\Upsilon}_k^Q)^2}{\sigma^2_{\tilde{n}^Q}}\\ &\left.-\frac{2\rho((y_d^{IQ})^I-\sqrt{E_r}\hat{\Upsilon}_k^I)((y_d^{IQ})^Q-\sqrt{E_r}\hat{\Upsilon}_k^Q)}{\sigma_{\tilde{n}^I}\sigma_{\tilde{n}^Q}}\right\}. \quad (40)\end{aligned}$$

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$$\mathrm{Ps}_o = Q\left(\sqrt{\frac{E_r}{4\left(1-\rho^2\right)}\left(\frac{(\Upsilon_k^I-\hat{\Upsilon}_k^I)^2}{\sigma_{\tilde{n}^I}^2}+\frac{(\Upsilon_k^Q-\hat{\Upsilon}_k^Q)^2}{\sigma_{\tilde{n}^Q}^2}-\frac{2\rho(\Upsilon_k^I-\hat{\Upsilon}_k^I)(\Upsilon_k^Q-\hat{\Upsilon}_k^Q)}{\sigma_{\tilde{n}^I}\sigma_{\tilde{n}^Q}}\right)}\right). \tag{43}$$

After simple mathematical operations, (40) is obtained in the form of

$$\begin{aligned}\mathrm{Ps}_o = \Pr\Bigg\{&\frac{2\rho\sqrt{E_r}\{(\Upsilon_k^I-\hat{\Upsilon}_k^I)\tilde{n}^Q+(\Upsilon_k^Q-\hat{\Upsilon}_k^Q)\tilde{n}^I\}}{\sigma_{\tilde{n}^I}\sigma_{\tilde{n}^Q}}\\ &-\frac{2\sqrt{E_r}(\Upsilon_k^I-\hat{\Upsilon}_k^I)\tilde{n}^I}{\sigma_{\tilde{n}^I}^2}-\frac{2\sqrt{E_r}(\Upsilon_k^Q-\hat{\Upsilon}_k^Q)\tilde{n}^Q}{\sigma_{\tilde{n}^Q}^2}\\ &-\frac{E_r(\Upsilon_k^I-\hat{\Upsilon}_k^I)^2}{\sigma_{\tilde{n}^I}^2}-\frac{E_r(\Upsilon_k^Q-\hat{\Upsilon}_k^Q)^2}{\sigma_{\tilde{n}^Q}^2}\\ &+\frac{2\rho E_r(\Upsilon_k^I-\hat{\Upsilon}_k^I)(\Upsilon_k^Q-\hat{\Upsilon}_k^Q)}{\sigma_{\tilde{n}^I}\sigma_{\tilde{n}^Q}}>0\Bigg\}\\ =\ &\Pr\{D>0\}.\end{aligned} \tag{41}$$

In this expression, D, conditioned on Υ_k, is a Gaussian RV with the variance of

$$\sigma_D^2 = 4E_r\left(1-\rho^2\right)\times\left(\frac{(\Upsilon_k^I-\hat{\Upsilon}_k^I)^2}{\sigma_{\tilde{n}^I}^2}+\frac{(\Upsilon_k^Q-\hat{\Upsilon}_k^Q)^2}{\sigma_{\tilde{n}^Q}^2}-\frac{2\rho(\Upsilon_k^I-\hat{\Upsilon}_k^I)(\Upsilon_k^Q-\hat{\Upsilon}_k^Q)}{\sigma_{\tilde{n}^I}\sigma_{\tilde{n}^Q}}\right). \tag{42}$$

Using (41) and (42), the conditional SEP of the optimal MLD can be given as in (43).

If there is only Tx IQI, (43) needs to be reorganized to define the conditional SEP of this case by considering that $\rho = 0$ while $\sigma_{\tilde{n}^I}^2 = \sigma_{\tilde{n}^Q}^2$; beside Υ_k^I and Υ_k^Q turn into $\Upsilon_{k,T}^I$ and $\Upsilon_{k,T}^Q$, respectively. After substituting these values in (43), the SEP is obtained as exactly the same as (22), which is the conditional SEP of the traditional MLD method in the presence of only Tx IQI. Hence, traditional MLD and optimal MLD are anticipated to provide the same performance results for only Tx IQI case. Additionally, this means that the noise part of the resulting signal given in (11), is not an IGN. Since $\mathbb{E}\{\tilde{n_T}^I\tilde{n_T}^Q\}=0$, this is an accurate conclusion.

In the presence of only the Rx IQI effect, $\sigma_{\tilde{n}^I}^2$ is obtained as equal to $\sigma_{N_R}^2$, and as a consequence of this, $\sigma_{\tilde{n}^Q}^2$ is calculated from $\sigma_{\tilde{n}^Q}^2 = \xi_R^2\sigma_{N_R}^2$. In order to obtain the conditional SEP for this case, besides substituting these new values in (43), the real and imaginary parts of $\Upsilon_{k,R}$ and $\hat{\Upsilon}_{k,R}$ are used in the same equation instead of Υ_k and $\hat{\Upsilon}_k$.

IV. Average and Asymptotic SEP of Optimal MLD for a Special Case

In this section, the ASEP is theoretically calculated for the optimal MLD method by assuming that source and relay nodes are connected by a P2P link, which is regarded as one of the transmission scenarios for future 5G networks [37]. Moreover, an asymptotic approximation is derived to have an idea about the effects of key parameters on the system behavior.

Firstly, (43) is rewritten as follows

$$\mathrm{Ps}_o = Q\left(\sqrt{\frac{E_r\Xi}{4(1-\rho^2)\sigma_{\tilde{n}^I}^2}}\right), \tag{44}$$

where $\Xi = \zeta_1^2+\zeta_2^2-2\varrho\zeta_1\zeta_2$, $\zeta_1 = (\Upsilon_k^I-\hat{\Upsilon}_k^I)$ and $\zeta_2 = (\Upsilon_k^Q-\hat{\Upsilon}_k^Q)/\xi_R$. Here, the terms ζ_1 and ζ_2 can be expanded as follows

$$\begin{aligned}\zeta_1 &= (\Lambda^I+\Omega^I)(x_k^I-\hat{x}_k^I)+(\Omega^Q-\Lambda^Q)(x_k^Q-\hat{x}_k^Q)\\ \zeta_2 &= [(\Lambda^Q+\Omega^Q)(x_k^I-\hat{x}_k^I)+(\Lambda^I-\Omega^I)(x_k^Q-\hat{x}_k^Q)]/\xi_R,\end{aligned} \tag{45}$$

where Λ^I and Λ^Q depend on the real and imaginary parts of $(G_1K_1h_1h_2+G_2K_2h_1^*h_2^*)$, while Ω^I and Ω^Q depending on the real and imaginary parts of $(G_2^*K_1h_1h_2+G_1^*K_2h_1^*h_2^*)$. Under the assumption of P2P communication between the source and the relay, since h_1 is a Gaussian link in this case, it is clear that ζ_1 and ζ_2 are independent Gaussian RVs that follow $\mathcal{N}(0,\upsilon_1)$ and $\mathcal{N}(0,\upsilon_2)$, respectively. After this, Ξ can be expressed in a quadratic form of Gaussian RVs as [38]

$$\Xi = \boldsymbol{\delta}^T\boldsymbol{A}\boldsymbol{\delta} = \sum_{l=1}^{2}a_lU_l^2, \tag{46}$$

where $\boldsymbol{\delta}^T = \begin{bmatrix}\zeta_1 & \zeta_2\end{bmatrix}$, U_l represents independently distributed standard normal variables, and a_l are the eigenvalues of the matrix $\boldsymbol{A}$, which has a quadratic form, given as

$$\boldsymbol{A} = \begin{bmatrix}\upsilon_1 & -\rho\sqrt{\upsilon_1\upsilon_2}\\ -\rho\sqrt{\upsilon_1\upsilon_2} & \upsilon_2\end{bmatrix}. \tag{47}$$

Additionally, a_l, in (46), are calculated from

$$a_l = \frac{\upsilon_1+\upsilon_2\pm\sqrt{(\upsilon_1+\upsilon_2)^2-4(1-\rho^2)\upsilon_1\upsilon_2}}{2}. \tag{48}$$

Noting that (46) gives Ξ as a linear combination of two independent central chi-square random variables with one degree of freedom, the moment generation function (MGF) of Ξ can be given by the following expression [39]

$$M_\Xi(t) = \prod_{l=1}^{2}(1-2ta_l)^{-\frac{1}{2}}. \tag{49}$$

Hence, an exact closed form expression of the ASEP is calculated by using (44) and (49) from

$$\overline{\mathrm{Ps}}_o = \frac{1}{\pi}\int_0^{\frac{\pi}{2}}M_\Xi\left(-\frac{E_r}{8(1-\rho^2)\sigma_{\tilde{n}^I}^2\sin^2\theta}\right)d\theta. \tag{50}$$

This integral can be easily solved via numerical methods. Additionally, for high SNR values, i.e., $E_r >> 1$, (49) approximates to $M_\Xi(t)\approx\frac{1}{\sqrt{4a_1a_2t^2}}$. In this case, an asymptotic approximation of ASEP can be obtained as follows

$$\overline{\mathrm{Ps}}_o \approx \frac{1}{\pi}\int_0^{\frac{\pi}{2}}\left(\frac{4\sigma_{\tilde{n}^I}^2\sqrt{1-\rho^2}\sin^2\theta}{E_r\sqrt{\upsilon_1\upsilon_2}}\right)d\theta. \tag{51}$$

In order to avoid the integral operation by considering the maximum value of $\sin^2\theta = 1$ in (51), ASEP can be upper bounded as well [40, p.230]. Under this assumption, substituting $\sigma_{\tilde{n}^I}^2$ in (51), asymptotic ASEP can be written as given in (52), which proves that amplitude IQI at Rx side has no effect on the optimal MLD system performance.

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$$\overline{\text{Ps}}_o \approx \frac{[E_r|h_2|^2\sigma_{n_1}^2 + E_s|h_1|^2(|G_1|^2+|G_2|^2)\sigma_{n_2}^2]\sqrt{1-\rho^2}}{E_rE_s|h_1|^2\sqrt{\upsilon_1\upsilon_2}(|G_1|^2+|G_2|^2)}. \quad (52)$$

V. Power Allocation Optimization

Under the condition of P2P communication between the source and the relay nodes, the PA between them is optimized to minimize the overall system error probability. This optimization problem can be formulated as follows [3]

$$E_s^{\#}, E_r^{\#} = \arg\min_{E_s,E_r} \overline{\text{Ps}}_o$$
$$\text{subject to}: E_t = E_s + E_r \text{ and } E_s, E_r > 0, \quad (53)$$

where E_t is total power consumption at the source and the relay nodes, while $E_s^{\#}$ and $E_r^{\#}$ are the optimal source and relay powers, respectively. In order to find an optimal solution for this problem, first, it should be proven that the objective function in (53), $\overline{\text{Ps}}_o$, is convex. In accordance with this purpose, the second derivative of this function is required to be obtained by rewriting (52) as

$$\overline{\text{Ps}}_o \approx \frac{\kappa}{E_s} + \frac{\varepsilon}{E_t - E_s}. \quad (54)$$

Here, $\kappa = \frac{\sqrt{1-\rho^2}|h_2|^2\sigma_{n_1}^2}{|h_1|^2\sqrt{\upsilon_1\upsilon_2}(|G_1|^2+|G_2|^2)}$ and $\varepsilon = \frac{\sqrt{1-\rho^2}\sigma_{n_2}^2}{\sqrt{\upsilon_1\upsilon_2}}$. Now, taking the second derivative of (54) respect to E_s leads to

$$\frac{\partial^2\overline{\text{Ps}}_o}{\partial E_s^2} = \frac{2\kappa}{E_s^3} + \frac{2\varepsilon}{(E_t-E_s)^3}. \quad (55)$$

It is easy to say that $\partial^2\overline{\text{Ps}}_o/\partial E_s^2$ is positive subject to optimization condition, $E_t > E_s > 0$, beside $\kappa > 0$ and $\varepsilon > 0$. This proves that $\overline{\text{Ps}}_o$ is a strictly convex function of E_s [3]. After this proof, the optimal solution can be found by taking the first derivative of $\overline{\text{Ps}}_o$ respect to E_s, and equating it to zero. Hence, the optimal source power is the root of

$$-\frac{\kappa}{E_s^2} + \frac{\varepsilon}{(E_t-E_s)^2} = 0. \quad (56)$$

Hereby, the optimal source and relay powers are obtained by considering $E_s^{\#} > 0$ and $E_r^{\#} = E_t - E_s^{\#}$ as follows

$$E_s^{\#} = \frac{-\kappa E_t + \sqrt{\varepsilon\kappa E_t^2}}{\varepsilon - \kappa}, \quad E_r^{\#} = \frac{\varepsilon E_t - \sqrt{\varepsilon\kappa E_t^2}}{\varepsilon - \kappa}. \quad (57)$$

VI. Complexity Analysis

The computational complexity is defined as the number of real multiplications required for any algorithm [41]. It is calculated by searching through the all possibilities and evaluating the Euclidean distances. Note that there are M possibilities of the modulated symbol, x_k, depending on the utilized modulation scheme. In addition to this, each complex multiplication requires 4 real multiplications while the square of the absolute value of a complex number requires 2. Considering these, the computational complexity of the proposed detectors are calculated.

For the traditional MLD detector given in (15), 52 real multiplications are required once to obtain Λ and Ω, which are used to find the term Υ_k. Considering the M probabilities of x_k, the computational complexity of traditional MLD is calculated as $C_t = 11M + 52$. Similarly, the complexity of the MLD with utilized compansation methods can be considered as $2M$ by using (39). However, the compensation procedures accompany more complexity burden. Applying compensation to reach y by using weighting method, first, w_1 and w_2 should be found and multiplied by y_d^{IQ} and its conjugate. This process costs 100 extra real multiplications. Since these operations are realized once, total complexity of MLD with weighting compensation is equal to $C_w = 2M + 100$.

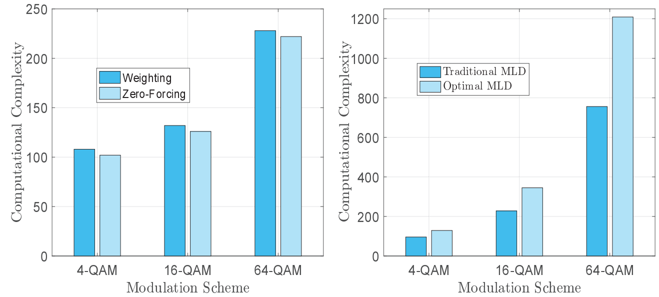

Fig. 3. Computational complexities of the proposed detector designs for different modulation schemes.

On the other hand, the matrix $\mathbf{H}^{-1}$ should be obtained to apply zero-forcing method, and it is multiplied by the vector $\mathbf{y}$. As this procedure contains 94 real multiplications, the complexity of MLD with zero-forcing compensation equals to $C_z = 2M + 94$. Finally, the complexity analysis is conducted by finding the summation of the complexities of each term in (39) for the proposed optimal MLD method. Considering that $8M+52$ real multiplications are necessary to find Υ_k^I and Υ_k^Q, as well as $\sigma_{\tilde{n}^I}^2$, $\sigma_{\tilde{n}^Q}^2$ and $2\rho/\sigma_{\tilde{n}^I}\sigma_{\tilde{n}^Q}$ should be calculated once, total computational complexity of the optimal MLD design is calculated as $18M + 57$.

Fig. 3 is given to compare the computational complexities of the proposed detector designs for 4-QAM, 16-QAM and 64-QAM modulation schemes, which are utilized in this study. It is clear that the compensated MLD is superior to the traditional and the optimal MLD designs in terms of complexity, especially for high modulation orders. Additionally, almost the same complexity with zero-forcing method is achived via the proposed weighting method by using simple math instead of confusing matrix operations. Optimal MLD design has the highest complexity burden, while zero-forcing method provides the minimum number of real computations.

VII. Numerical Results

In this section, the effects of IQI at both Tx and Rx sides on the performance of the AF dual-hop CSI-assisted relaying wireless communication system are evaluated by using the MLD designs given in Section III through computer simulations. Additionally, the analytical derivation is verified under the assumption of P2P transmission between the source and the relay nodes. PA optimization results are provided under this assumption as well. ASEP results are validated with

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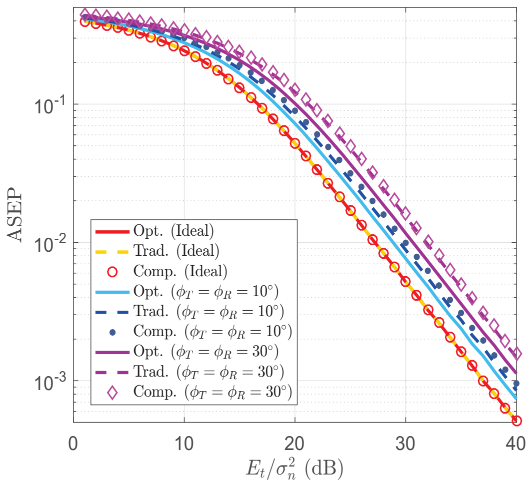

Fig. 4. ASEP simulation results of the proposed MLD designs (optimal (opt.), traditional (trad.) and compensated (comp.)) for AF dual-hop relaying system in the presence of both Tx and Rx phase imbalance with fixed gain imbalance ($\xi_T = \xi_R = 0.6$).

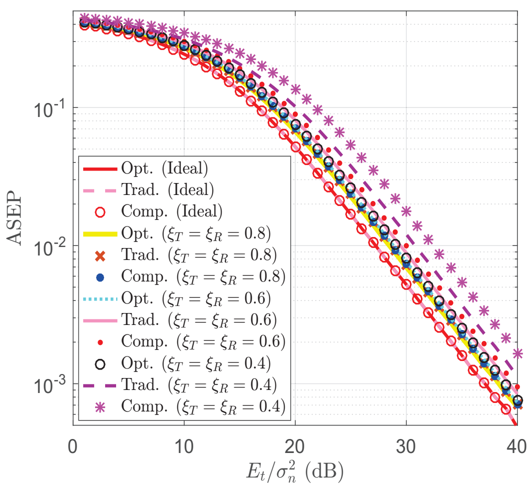

Fig. 5. ASEP simulation results of the proposed MLD designs (optimal (opt.), traditional (trad.) and compensated (comp.)) for AF dual-hop relaying in the presence of both Tx and Rx gain imbalances with fixed phase imbalance ($\phi_T = \phi_R = 10°$).

equal PA ($E_s = E_r = E_t/2$) by using 16-QAM modulation scheme unless otherwise stated. In this case, optimal MLD has approximately 1.5-fold and 2.5-fold more complexity than the traditional MLD and the compansation methods, respectively. It is assumed that at least 10^6 symbols have been transmitted for each SNR value, which is defined as E_t/σ_n^2, between $0-40$ dB.

First of all, AF dual-hop relaying system under the effect of both Tx and Rx phase imbalances is analyzed by the proposed MLD methods. Since weighting and zero-forcing compensation techniques provide the same mathematical equations, the results belong to this two methods are given as one and called compansated MLD. In Fig. 4, the ASEP results are presented with fixed values of amplitude imbalance, $\xi_T = \xi_R = 0.6$, while varying the phase imbalance among $10°$ and $30°$. It is observed that the optimal MLD achieves the best results for all cases. Its superiority is clearer with increasing phase imbalance. On the other hand, traditional

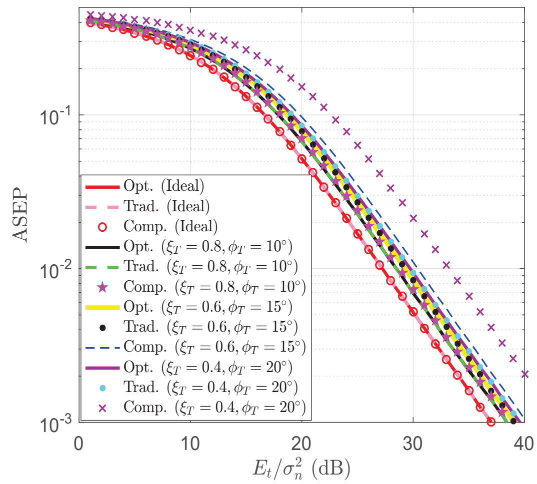

Fig. 6. ASEP results of AF dual-hop relaying in the presence of Tx IQI.

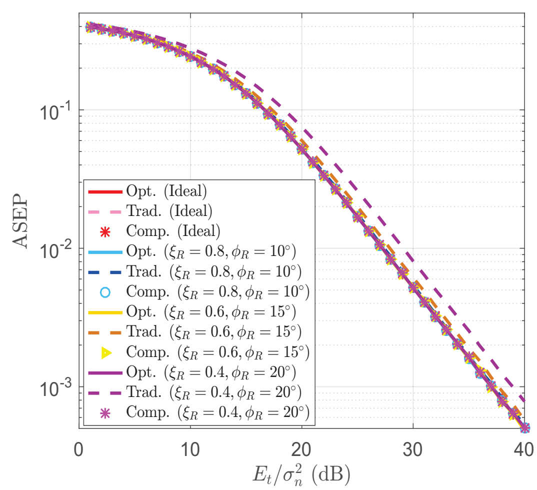

Fig. 7. ASEP results of AF dual-hop relaying in the presence of Rx IQI.

MLD and compensated MLD provide almost the same results. Comparing to the ideal case, the destructive effect of IQI on the overall system performance is quite obvious as well.

Fig. 5 is given to analyze the effects of both Tx and Rx amplitude imbalances on the ASEP performance for fixed values of phase imbalance, $\phi_T = \phi_R = 10°$, while varying the amplitude imbalance among $0.4, 0.6$ and 0.8. It is clear that optimal MLD is the most resistant scheme to amplitude IQI. For instance, increasing amplitude imbalance effect by changing ξ_T and ξ_R from 0.8 to 0.4 has caused approximately 2.5 dB and 4.5 dB worse results with traditional and compensated MLD schemes, respectively, while the optimal MLD technique achives almost the same QoS at $\mathrm{ASEP} = 10^{-2}$.

The proposed methods are also compared by considering Tx- and Rx-side IQI effects individually. The ASEP results of the traditional MLD are as effective as the optimal MLD results, as it was expected in Section III-C, nothwithstanding the severity of the deterioration for only Tx-side IQI case as seen in Fig. 6. This two methods are more effective to make AF dual-hop relaying system robust against to Tx IQI impairments than the compansated MLD design, especially for high level IQI. For example, increasing Tx IQI effect

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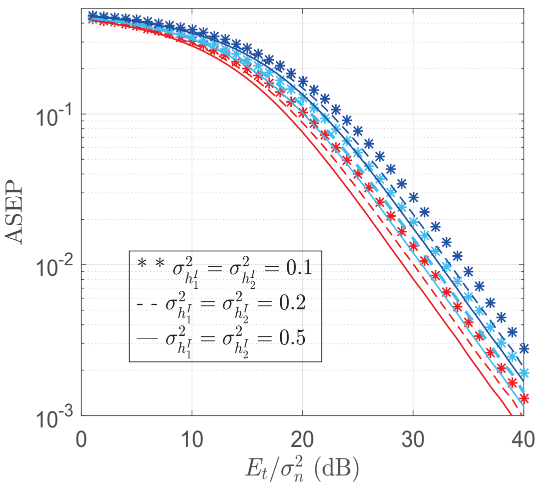

Fig. 8. ASEP simulation results of the proposed MLD designs (red: optimal, light blue: traditional, and blue: compensated) over non-identical Rayleigh fading channels for AF dual-hop relaying in the presence of both Tx and Rx IQI ($\xi_T = \xi_R = 0.4, \phi_T = \phi_R = 10°$).

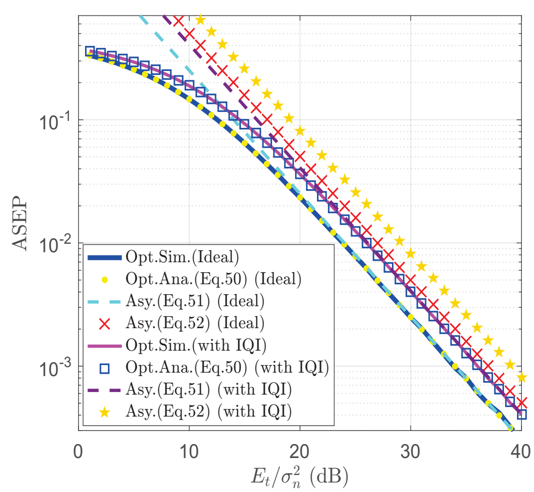

Fig. 9. ASEP results of AF dual-hop relaying with perfect balance and IQI ($\xi_T = \xi_R = 0.6$ and $\phi_T = \phi_R = 10°$).

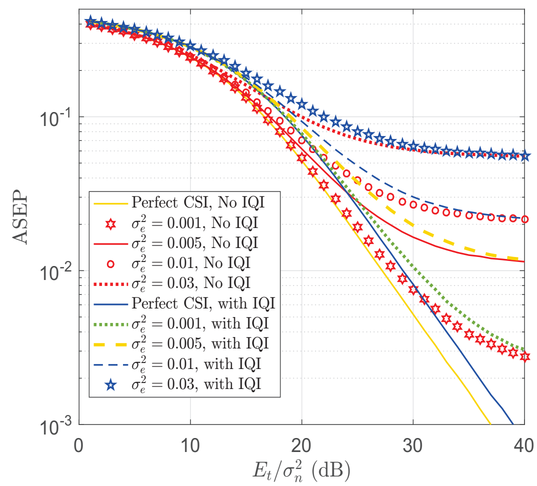

Fig. 10. ASEP results of AF dual-hop relaying system under the joint effect of imperfect CSI and IQI ($\xi_T = \xi_R = 0.4$ and $\phi_T = \phi_R = 10°$).

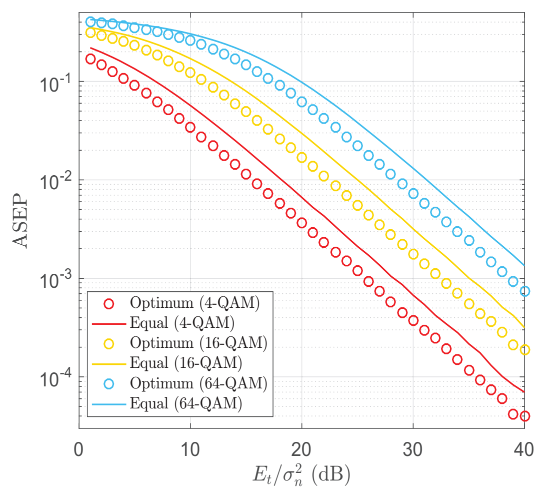

Fig. 11. ASEP results of AF dual-hop relaying in the presence of IQI ($\xi_T = \xi_R = 0.6$ and $\phi_T = \phi_R = 15°$) for optimum and equal PA cases with different modulation schemes.

by changing $\xi_T = 0.6, \phi_T = 15°$ to $\xi_T = 0.4, \phi_T = 20°$ has almost no effect on the QoS of traditional and optimal MLD, however this causes approximately 3 dB worse results for compansated MLD. In Fig. 7, it is observed that optimal and compensated MLD methods, in the presence of only Rx IQI, provide the same QoS, which is also equal to ASEP results of the ideal transceiver. It means that the optimal and compensated MLD methods are totally able to cancel Rx IQI effects. This result is in compliance with (33) and the remarks therein.

The effects of non-identical variance characteristics of the Rayleigh fading channel on the CSI-assisted AF relaying system is investigated with the ASEP results given in Fig. 8. Noting that $\sigma^2_{h^I} + \sigma^2_{h^Q} = 1$ for both h_1 and h_2, it is clear that increasing variance imbalance between the real and imaginary parts of the Rayleigh fading coefficients, decreases the system performance for all detector designs. Hence, the best results are provided under the assumption of identically distributed ($\sigma^2_{h^I} = \sigma^2_{h^Q} = 0.5$) Rayleigh fading channel environment.

Fig. 9 is given to prove that our analytical ASEP analysis for the optimal MLD design based on P2P assumption is accurate. It is seen that computer simulation results perfectly match with the analytical results in both ideal and IQI effected AF communication systems. Besides this, the asymptotic analysis results are successfully achieved in accordance with the analytical results. The asymptotic analysis results, which are upper-bounded with (52), are presented in Fig. 9 as well, and PA is optimized based on the results obtained from this equation.

In order to obtain more realistic results, and give an idea for the future studies, which will focus on channel estimation errors, the effects of imperfect CSI case at the destination node are considered with optimal MLD in Fig. 10. This figure not only proves that even very small values of the estimation error, such as $\sigma_e^2 = 0.001$, substantially reduce the overall QoS, but also shows that imperfect CSI is a crucial impairment on the system performance of AF cooperative relaying. The error floor arising from the imperfect CSI reveals the need for effective algorithms to improve the quality of the estimator.

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Finally, the ASEP performance of the optimal MLD that uses the optimized PA is compared to the performance of the optimal MLD using equal PA. Optimization procedure provides approximately 2.5 dB better results for IQI affected AF dual-hop relaying systems using 4, 16 and 64-QAM modulation schemes as seen in Fig. 11. It is also shown that increasing the spectral efficiency from 2 bps/Hz to 6 bps/Hz, by using 64-QAM insted of 4-QAM, costs almost 13 dB loss in overall AF QoS. This is a key result that points out the importance of considering IQI effects on future AF cooperative communication system designs, which require high data rates.

VIII. Conclusion

The destructive effects of IQI, which are highlighted in terms of SIR, on the performances of dual-hop CSI-assisted AF cooperative communication systems have been addressed via computer simulations in this study. An optimal MLD method has been proposed, and performance results have been compared to the results obtained by using the traditional MLD in terms of ASEP. Furthermore, two compensation algorithms have been used to detect the transmitted signal. The obtained results point out that IQI is a major drawback for AF dual-hop cooperative systems, and the proposed optimal MLD design is the best to mitigate the effects of IQI among these methods, although it has more complexity burden. Moreover, an exact closed form expression has been derived for optimal MLD with P2P connection between the source and the relay nodes, and the effects of imperfect CSI at the destination node is analyzed.

A further performance improvement has been achieved by optimizing the PA parameters through the asymptotic approximations for this case. As increasing data rates cost more performance degradations, the effects of IQI has to be considered seriously while designing next-generation wireless systems. PA algorithms can provide better QoS as well as energy efficiency. Future work can focus to the extention of the presented analysis and compensation developments to more challenging propagation environments and frequency-domain effects of IQI. Also extending the work towards multiple-input multiple-output cooperative systems offers an interesting topic for future studies.

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