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Designs of MIMO Amplify-and-Forward Wireless Relaying Networks: Practical Challenges and Solutions Based on MSE Decomposition

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ABSTRACT Amplify-and-forward (AF) relaying is an efficient way to extend radio range and improve link reliability with a low implementation cost. While the research on the AF relaying in a single-antenna scenario has matured, its application to broadband radio transmission utilizing multiple-input multiple-output techniques still needs more effort to overcome several practical challenges, such as large overheads for channel estimation and vulnerability to the channel estimation errors. This paper reviews some recently proposed linear transceiver designs based on the minimum mean squared error (MSE) criterion and shows that the proposed MSE decomposition and relaxation method can lead to an efficient solution for those challenges. Insightful observations and comprehensive discussions are also made on both analytical and numerical results from practical implementation perspectives.

INDEX TERMS Survey, MIMO, relay, AF, MMSE, transceiver design, MSE-decomposition, robust design, performance analysis.

I. INTRODUCTION

In next generation radio communication systems, cellular concepts in which one base station handles all users within its coverage area may no longer support exponentially increasing subscribers and data rate requirements. In particular, the spectrum released for broadband mobile systems will typically be located well above the 2 GHz band in which radio propagation significantly suffers from high power decaying especially in urban cellular networks. To address this problem, relaying techniques have garnered significant interests over the last several years, thanks to its advantages of improved reliability and radio range extension with low implementation complexity.

In the relaying systems, the link between a base-station and a relay is often referred to as the *backhaul link*, while the link between the relay and the mobile user is called the *access link*. Depending on the radio frequency (RF) assignment for two-hop links, relays are either inband or outband. The inband relay uses the same RF for the backhaul link and the access link, whereas the outband relay utilizes different RFs on both links. Relays can also be classified into full-duplex and half-duplex, depending on whether the relay can simultaneously transmit and receive or not.

For general relay channels, the capacity achieving relaying protocol is still unknown [1]. Among several forwarding strategies at the relay, two schemes are popular. First, a *decode-and-forward* (DF) scheme decodes and re-encodes the received signal prior to forwarding it to the receiver. This type of relay allows separate rate adaptation and scheduling for the backhaul and access links. However, a large processing delay and high implementation complexity remains as issues.

In contrast, an *amplify-and-forward* (AF) scheme simply retransmits the linearly amplified version of the received signal. As the additive noise is also amplified along with the desired signal, the output signal-to-noise ratio (SNR) at the destination may deteriorate. Nevertheless, the AF relay

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exhibits advantages of having low cost and low processing delay, which are usually the most important points when deploying network nodes. For this reason, in this article, we focus on the AF relaying systems.

In the meantime, it has been well recognized that multiple-input multiple-output (MIMO) techniques provide considerably improved spectral efficiency. One simple approach for realizing such a capacity advantage is spatial multiplexing (SM) which transmits multiple data streams in parallel for a given time and frequency. As for the receiver designs against the SM transmission, the minimum mean squared error (MMSE) criterion has commonly been adopted, attributed to good performance with low decoding complexity [2]. The MMSE approach can also be used for information rate maximization by exploiting the relation between MMSE and mutual information [3], [4], and thus is a useful performance metric for both the practical and theoretical perspectives.

The SM transmission may operate in two different modes, namely, *open-loop SM* and *closed-loop SM*. The open-loop SM transmits data without channel state information at the transmitter (CSIT). Since the CSIT is mostly based on the feedback or sounding signals from the receivers, the open-loop scheme is suitable for high-mobility environment where reliable feedback is not feasible. In contrast, to achieve a full throughput gain of MIMO systems, we can apply the precoding schemes at the transmitters when the accurate CSIT is available, which is referred to as the closed-loop SM.

Compared to conventional point-to-point (P2P) MIMO systems, a transmitter and receiver (transceiver) design in MIMO AF relaying is much more challenging due to the coupled transmission links and the noise propagation. Regarding the open-loop SM relaying without the CSIT at the source, the authors in [5] and [6] have first proposed the optimal joint designs on the linear transceivers between the relay and the destination with respect to the information rate and the MMSE, respectively. Then, extending them to the closed-loop SM relaying with full CSIT at the source, [4], [7], and [8] suggested the optimal source-relay-destination joint transceivers in terms of max-rate, MMSE, and quality-of-service (QoS) criteria, respectively. The work [9] would be a good tutorial for an introduction to the linear transceiver designs in MIMO-AF relaying systems. The transceiver design was also investigated in [10]–[12] considering the direct path between the source and the destination. Some of the results have been further extended to the multi-user relaying scenarios in [13]–[15] for the downlink and in [16] and [17] for the uplink scenarios.

Due to non-convexity of the transceiver design problems, the optimal solutions are mostly attainable from iterative algorithms having high complexity. To reduce the complexity to the level of implementable efficiency, several suboptimal closed-form designs have been developed by exploiting the MSE decomposition property [18]–[23]. The MSE decomposition method also facilitates the error performance analysis of the systems. Thus, several analytical results was then reported in [24]–[28] to evaluate the uncoded and coded bit error rate (BER) performance, which provide a helpful guideline for designing the MIMO AF relaying systems. To achieve accurate channel estimation at each node, up/downlink training and channel estimation methods have also been treated as important issues for implementing the relay nodes as discussed in [29]–[38]. A robust transceiver design is also being investigated in terms of both the stochastic [39]–[43] and the worst-case [44], [45] philosophies to address the CSI mismatch due to the channel estimation error at each node of the MIMO relaying systems.

In this article, we provide a comprehensive discussion on the MMSE-based SM MIMO AF relaying systems which have been considered as a powerful class of practical relaying systems. It is worth noting that although the AF relaying in single antenna scenarios is now realizable, its application to MIMO broadband transmission still needs more effort to overcome several practical challenges such as resource burden for channel estimation and a countermeasure about the estimation errors. The goal of this article is to highlight these challenges and suggest an efficient solution. Through an in-depth discussion on the analytical performance and the computational complexity of various relaying strategies, we also provide key insights into the design of SM MIMO AF relaying systems, thereby helping readers to fully understand practical ways of their implementation.

This paper is organized as follows: Section II describes the channel model for MIMO AF relay systems. In Section III, we present conventional designs and practical challenges. Then, the MSE decomposition-and-relaxation method and its applications are introduced in Section IV. Also, some analytical and numerical results are explained in Section V and Section VI, respectively. Finally, the paper is terminated with conclusions in Section VII.

II. MIMO AF RELAY CHANNEL MODELING

In this article, we consider an SM MIMO AF relaying system as presented in Fig. 1, where the source, the relay, and the destination are equipped with N_t, N_r, and N_d antennas, respectively. Then, assuming the narrow-band flat fading, the baseband equivalent channels can be represented by complex matrices $\mathbf{H} \in \mathbb{C}^{N_r \times N_t}$ for the backhaul link and $\mathbf{G} \in \mathbb{C}^{N_d \times N_r}$ for the access link, where $\mathbb{C}$ denotes a set of complex numbers. When the frequency selectivity occurs in the wide-band channels, multi-carrier processing such as the orthogonal frequency division multiplexing (OFDM) can be adopted in each hop to transform the frequency selective channel into a set of parallel flat faded subchannels.

We normalize the range of the access link as a unit distance while setting the distance of the backhaul link as a variable α (i.e., the distance ratio equals $\alpha : 1$), and assume that both links experience the same pathloss exponent ζ. The transmit signal at the source is denoted by $\mathbf{x} \in \mathbb{C}^{N_s \times 1}$, which means that N_s parallel sub-streams are transmitted simultaneously. Here, N_s is normally adjusted to be lesser than or equal to the minimum of N_t, N_r,

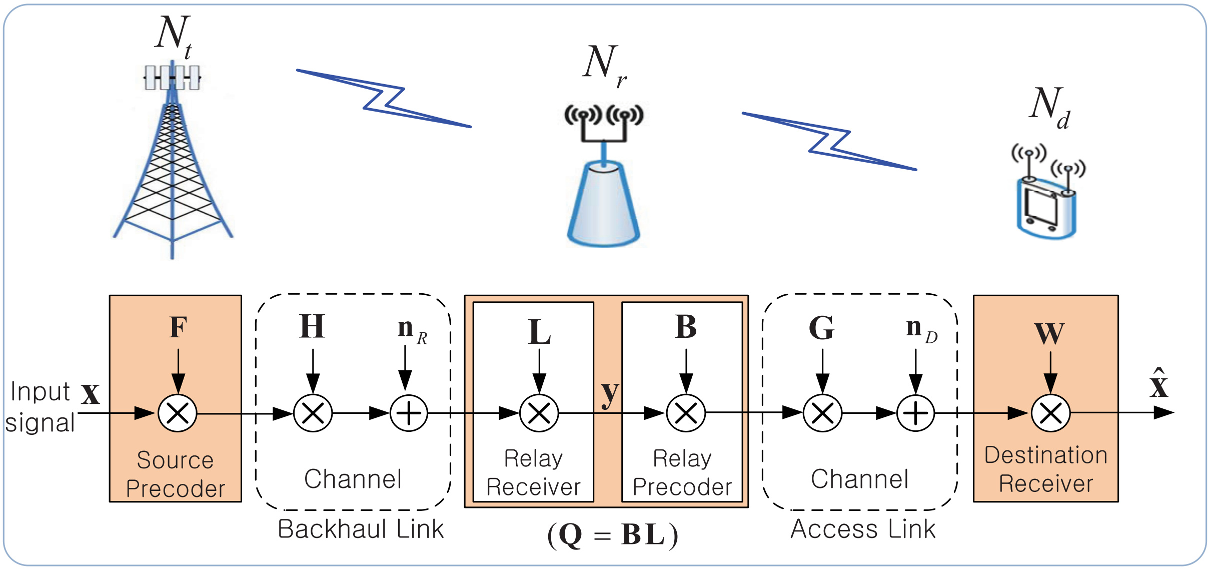

FIGURE 1. **Baseband signal model of the MIMO AF relaying systems based on the MSE decomposition and relaxation method.**

and N_d. Throughout the article, we ignore the direct path between the source and the destination due to the large pathloss.

As shown in Fig. 1, in our system model, the relay can be either half-duplex inband or full-duplex outband,[1] where a signal transmission typically occurs over two different phases. In the first phase, the input signal $\mathbf{x}$ is transmitted to the relay after being precoded by the source precoder $\mathbf{F} \in \mathbb{C}^{N_t \times N_s}$. Then, in the second phase, the received signal at the relay is multiplied by relay amplifying matrix $\mathbf{Q} \in \mathbb{C}^{N_r \times N_r}$ and retransmitted to the destination. Then, final observation vector $\hat{\mathbf{x}} \in \mathbb{C}^{N_s \times 1}$ at the destination is given by

$$\hat{\mathbf{x}} = \mathbf{W}\Big\{\mathbf{GQ}(\mathbf{HFx} + \mathbf{n}_R) + \mathbf{n}_D\Big\}, \tag{1}$$

where $\mathbf{W} \in \mathbb{C}^{N_s \times N_d}$ designates the linear receiver matrix at the destination, and $\mathbf{n}_R \in \mathbb{C}^{N_r \times 1}$ and $\mathbf{n}_D \in \mathbb{C}^{N_d \times 1}$ indicate additive noise vectors at the relay and the destination, respectively.

When the channel estimation errors are taken into account, each of the true channels $\mathbf{H}$ and $\mathbf{G}$ is further expressed by $\mathbf{H} = \overline{\mathbf{H}} + \Delta\mathbf{H}$ and $\mathbf{G} = \overline{\mathbf{G}} + \Delta\mathbf{G}$, where $\overline{\mathbf{H}}$ and $\overline{\mathbf{G}}$ indicate estimated channels, and $\Delta\mathbf{H}$ and $\Delta\mathbf{G}$ represent uncertainty matrices. As the instantaneous values of the channel uncertainty matrices are unknown for all nodes, the signals multiplied by $\Delta\mathbf{H}$ and $\Delta\mathbf{G}$ will be considered as background noise. Thus, the system model in (1) also holds for the case of the channel estimation errors, although the statistical property of the noise vectors $\mathbf{n}_r$ and $\mathbf{n}_d$ may be changed [39], [40].

[1] A full-duplex inband relay is not considered in this article since the loop interference between the incoming and outgoing signals at the relay may incur a significant performance degradation without proper interference cancellation techniques which are beyond the scope of this article.

III. CONVENTIONAL DESIGNS AND CHALLENGES

A. OPTIMAL DESIGNS WITH GLOBAL CSI

The MSE between the input signal $\mathbf{x}$ and the estimated signal $\hat{\mathbf{x}}$ is expressed as

$$MSE = \mathbb{E}\left[\|\hat{\mathbf{x}} - \mathbf{x}\|_2^2\right], \tag{2}$$

where $\mathbb{E}[\cdot]$ and $\|\cdot\|_2$ denote the expectation and the 2-norm operations, respectively. Under the MMSE strategy, the goal is to minimize the MSE metric in (2) by customizing the linear filters $\mathbf{F}$, $\mathbf{Q}$, and $\mathbf{W}$ to the channels. In order to optimize these filter matrices, each node must be informed of the full CSI of both the backhaul link and the access link, which is referred to as *"global CSI"*. For example, the optimal receiver at the destination is given by the Wiener filter solution [19], which requires the global CSI at the destination.

For the single-input single-output cases, the output SNR is mostly determined by the receiver operation at the destination, and thus the CSIT is of little interest. In contrast, as for the MIMO relaying, the precoding operations at the source or the relay greatly affect the performance, and thus should be properly optimized utilizing the CSIT. With the MMSE criterion and the Wiener equalizer at the destination, the precoding strategies of the relaying systems are generally classified into three cases:

- *Optimal joint precoding (Opt-JP) [7]* jointly optimizes the source precoder $\mathbf{F}$ and the relay forwarding matrix $\mathbf{Q}$ assuming the global CSI at all nodes, and thus is corresponding to the closed-loop SM relaying. Opt-JP performs an iterative algorithm with multiple initial points to find a solution close to the true optimum. Opt-JP is not strictly *"optimal"* because the global optimality is not always guaranteed due to non-convexity of the design problem. Nevertheless, it often serves as a performance limit of the MMSE-based MIMO AF relaying systems.

- *Optimal relay only precoding (Opt-RP) [6]* realizes the open-loop SM relaying, where no CSIT is needed at the source. We note that the global CSI is still compulsory at the relay and the destination to compute the relay matrix $\mathbf{Q}$ and the Wiener filter at the destination. This scheme is casted into a special case of the Opt-JP with $\mathbf{F} = \gamma_f \mathbf{I}$ where $\mathbf{I}$ and γ_f denotes an identity matrix and a power normalizing coefficient, respectively. Unlike the Opt-JP, we can find a unique optimal solution as a closed-form, since the design problem is reformulated into an equivalent convex form.
- *Naive AF* is the simplest scheme in which no precoding operation is employed at both the source and the relay [28], and thus is also associated with the open-loop SM relaying with no CSI at both the source and the relay, i.e, $\mathbf{F} = \gamma_f \mathbf{I}$ and $\mathbf{Q} = \gamma_q \mathbf{I}$ where γ_q denote the power normalizing coefficient at the relay. However, we note again that the destination still needs to know the global CSI to compute the destination receiver $\mathbf{W}$.

B. ROBUST DESIGNS WITH CHANNEL ESTIMATION ERRORS

In practical communication systems, channel estimation errors are inevitable, due to the limited length of training sequences and/or the time-varying nature of wireless channels. The channel estimation error typically incurs serious performance degradation due to the mismatch between the channels and the filters. Therefore, a robust design which mitigates such a performance loss is also an essential part of the transceiver designs. Recently, the robust designs for MIMO AF relaying systems have widely been investigated in literature [39]–[45]. Similar to the optimal designs, they are simply divided into two classes:

- *Robust Opt-JP [40]* jointly optimizes all nodes taking the channel uncertainty into consideration. This scheme amounts to a robust version of Opt-JP. As the problem is still non-convex, a solution is found by iterative methods assuming the global CSI (possibly imperfect) at all nodes.
- *Robust Opt-RP [39]* jointly determines the relay and the destination assuming no CSIT at the source, but the global CSI (possibly imperfect) at both the relay and the destination. This scheme is a robust version of Opt-RP, and several of efficient solutions.

C. CHALLENGES FOR CHANNEL ESTIMATION

A key distinction of the two-hop MIMO AF relaying compared to the one-hop P2P MIMO is that two channel matrices in the backhaul and access links are interlinked with each other Therefore, the conventional channel estimation methods developed in the P2P MIMO systems are not immediately applied to the AF relaying systems. In particular, to compute Opt-JP, Opt-RP, robust Opt-JP, and robust Opt-RP at each node, it is required to estimate the global CSI, which is difficult (even if it is imperfect) to attain in practice.

1) CHANNEL ESTIMATION AT THE DESTINATION

As mentioned previously, to implement the optimal Wiener filter at the destination, global CSI is essential. One way to obtain the global CSI at the destination is that the relay first estimates the backhaul link channel, and then forwards it to the destination together with its own pilots. In practice, however, forwarding the estimated CSI to another node over the wireless channel is undesirable, because it may cause an additional distortion by the quantization error and channel noise during the second transmission. In particular, such a forwarding process will severely impair the performance when detection failure occurs at the destination. The channel forwarding process may also cause an inefficient bandwidth usage, because extra resource should be allocated in advance at the source, which the relay uses to forward its estimated CSI to the destination.

To combat the problems, one may envision a way to estimate the compound channel, i.e., the effective channel from the source to the destination [29]–[33]. However, not to mention the performance loss due to ignorance of the noise correlation, the compound channel estimation approach may be even unrealizable in broadband wireless systems, because the compound channel generates a large number of unknown parameters that have to be estimated at the destination. For example, if each link experiences multi-path fading with the L-tap delay profile, the total number of unknown parameters that we should estimate becomes $(2L-1)N_sN_rN_d$, which is typically larger than the number of occupied subcarriers per an OFDM symbol, e.g., $76 \sim 1201$ in LTE-A [2]. The result implies that several OFDM symbols must be used for estimating a single MIMO channel, which makes the system vulnerable to the time-varying nature of wireless channels.

Striving to overcome such challenges, recently the authors in [34]–[36] have proposed individual channel estimation method based on the segment training with the aid of an appropriately designed relay matrix. Also, the tensor and compressed sensing based algorithms have been proposed in [37] and [38], respectively. However, the channel estimation complexity is still incomparably high over the P2P MIMO systems.

2) CHANNEL ESTIMATION AT TRANSMITTERS

At the transmitter sides, obtaining the global CSI is even more challenging, since perfect channel reciprocity or a large amount of feedback information are additionally required.

D. CHALLENGES FOR ROBUST DESIGNS

Considering the global CSI, if any of the two channels fails in accurate channel estimation, it may incur a critical mismatch between a filter and the true channels at each node. Therefore, a system should be designed to be strong to the estimation errors of both channels, which increases implementation complexity. A more important thing here is that in real environment, the amount of estimation error of a channel matrix is likely to be different between the relay and the destination, due to the different mobility and path loss over the two hops, and possibly additional distortion that may occur during the CSI feedforward. Therefore, we may need to address different levels of channel uncertainty between the backhaul and access links. Nevertheless, an efficient robust design for such cases is still open since it is difficult to address different levels of channel uncertainty at once.

IV. SOLUTIONS: MSE DECOMPOSITION AND RELAXATION

In this section, we propose that some suboptimal designs based on the MSE decomposition and relaxation can be a solution to the aforementioned challenges. To this end, we first explain some useful features hidden behind the MMSE problem in (2), and then demonstrate the efficiency of the proposed designs.

A. MSE DECOMPOSITION

We start by noting that the optimal relay matrix is without loss of generality expressed as a product of two matrices, i.e., $\mathbf{Q} = \mathbf{BL}$ with relay precoder $\mathbf{B} \in \mathbb{C}^{N_r \times N_s}$ and relay receiver $\mathbf{L} \in \mathbb{C}^{N_s \times N_r}$ [19], [20], [28], [46] as shown in Fig. 1. The relay precoder is arbitrary, and thus unknown as of yet, but the relay receiver is given by an Wiener equalizer for the backhaul link channel $\mathbf{H}$.

Now, we define $\mathbf{y}$ as the output signal of the relay receiver $\mathbf{L}$, i.e., the MMSE (or Wiener) estimate of $\mathbf{x}$ at the relay as

$$\mathbf{y} = \mathbf{L}(\mathbf{Hx} + \mathbf{n}_R).$$

Then, by the orthogonality principle [47], $\mathbf{y}$ is orthogonal to the error vector $\mathbf{y} - \mathbf{x}$. Next, let us reformulate the MSE in (2) as $E[\|\hat{\mathbf{x}} - \mathbf{y} + \mathbf{y} - \mathbf{x}\|_2^2]$. From the orthogonality principle again, it is seen that for a given $\mathbf{Q} = \mathbf{BL}$, the error vector $\mathbf{y} - \mathbf{x}$ becomes orthogonal to $\hat{\mathbf{x}}$ as well as $\mathbf{y}$, because $\hat{\mathbf{x}}$ in (1) is also a function of $\mathbf{y}$ and independent noise $\mathbf{n}_D$ as

$$\hat{\mathbf{x}} = \mathbf{W}(\mathbf{GBy} + \mathbf{n}_D).$$

Therefore, the total MSE in (2) is equivalently decomposed as

$$\begin{aligned} MSE &= \mathbb{E}\left[\|\hat{\mathbf{x}} - \mathbf{y}\|_2^2] + \mathbb{E}[\|\mathbf{y} - \mathbf{x}\|_2^2\right] \\ &= \mathbb{E}\left[\|\mathbf{W}(\mathbf{GBy} + \mathbf{n}_D) - \mathbf{y}\|_2^2\right] \\ &\quad + \mathbb{E}\left[\|\mathbf{L}(\mathbf{HFx} + \mathbf{n}_R) - \mathbf{x}\|_2^2\right]. \end{aligned} \tag{3}$$

The above MSE decomposition highlights several fundamental aspects of the MIMO AF relaying systems. First, it allows us to consider the system as a simple concatenation of two MIMO P2P systems: one between the precoder $\mathbf{F}$ and the receiver $\mathbf{L}$ with the input signal $\mathbf{x}$, and the other one between the precoder $\mathbf{B}$ and the receiver $\mathbf{W}$ with the input signal $\mathbf{y}$. In principle, the two MSE terms in (3) are still related to each other, because the source precoder $\mathbf{F}$ in the second MSE term is also affiliated with $\mathbf{y}$ in the first MSE term. However, it can be verified that at both the low and high SNR regimes, the first and second order statistics of $\mathbf{y}$ is closely approximated to that of $\mathbf{x}$, which means that the first MSE term becomes independent of the source precoder $\mathbf{F}$ [19]. Thus, by an assumption that $\mathbf{x}$ and $\mathbf{y}$ are statistically equivalent, i.e., $\mathbf{y} \sim \mathbf{x}$, two MSE terms in (3) are separable from each other. Although the independence is not yet clear for the medium SNR range, extensive computer simulations demonstrate that the MSE decomposition and relaxation method exhibits little performance loss over all SNR region.

TABLE 1. Required CSI for relaying systems.

Schemes		Source	Relay	Destination
Global CSI based designs	Opt-JP	$\mathbf{H}$, $\mathbf{G}$	$\mathbf{H}$, $\mathbf{G}$	$\mathbf{H}$, $\mathbf{G}$
	Opt-RP	-	$\mathbf{H}$, $\mathbf{G}$	$\mathbf{H}$, $\mathbf{G}$
	Naive AF	-	-	$\mathbf{H}$, $\mathbf{G}$
Local CSI based designs	FP / FP	$\mathbf{H}$	$\mathbf{H}$, $\mathbf{G}$	$\mathbf{G}$
	LP / LP	k_F-bits	$\mathbf{H}$, k_B-bits	$\mathbf{G}$
	NP / FP	-	$\mathbf{H}$, $\mathbf{G}$	$\mathbf{G}$
	NP / LP	-	$\mathbf{H}$, k_B-bits	$\mathbf{G}$
	NP / NP	-	$\mathbf{H}$	$\mathbf{G}$

We recognize from (3) that the transceiver design paradigm of MIMO AF relaying systems is now shifted from finding $\mathbf{F}$ and $\mathbf{Q}$ for a given receiver $\mathbf{W}$ as described in Section III-A to identifying precoders $\mathbf{F}$ and $\mathbf{B}$ for given receivers $\mathbf{L}$ and $\mathbf{W}$, respectively. This new strategy offers flexibility towards the precoder designs, because conventional beamforming and power allocation strategies in the P2P systems [48], [49] can be applied to the source and the relay in a distributed manner without requiring the joint design. It is important to notice that in this case, the source and the destination need to estimate only the channel that is connected to their own link, i.e., the backhaul link channel for the source and the access link channel for the destination, which is referred to as the *"local CSI"*. This gives rise to a great reduction in the estimation burden of AF relaying systems.

B. SUBOPTIMAL DESIGNS WITH LOCAL CSI

Assuming that $\mathbf{y} \sim \mathbf{x}$, we can construct the relay and destination receivers $\mathbf{L}$ and $\mathbf{W}$ as the Wiener equalizers customized to fit their corresponding CSIs at the receivers (CSIR), i.e., $\mathbf{H}$ and $\mathbf{G}$, respectively. Then, for given structures of $\mathbf{L}$ and $\mathbf{W}$, the remaining work is to find the precoders $\mathbf{F}$ and $\mathbf{B}$ to minimize each MSE term in (3). Traditionally, the precoding strategy in one-hop MIMO channels is categorized into three schemes according to the level of CSIT: full-CSI-feedback precoding (FP) [49], limited-feedback precoding (LP) with shared codebooks [50], and non-feedback precoding (NP) which is associated with the open-loop SM. Applying these to the source and relay precoders separately, we finally arrive at the desired precoders with the local CSI. We note that although not discussed here in detail, many other precoding and codebook designs can be employed for FP and LP according to the situation.

For readers' better understanding, the amount of CSI required for each design scheme is summarized in Table 1

TABLE 2. Existing works for precoder designs using the MSE decomposition in MIMO AF relaying systems.

Categories	S/ R/ D	Direct Link	Transceiver Features	Main Objectives	Main Results	Ref.
Two-Hop	1/ 1/ 1	×	Linear receiver	Source, relay, and destination optimization for MMSE/MBER s.t. individual power constraints	Closed-form solution; Limited feedback design	[19]
		○	Linear receiver	Relay and destination optimization for MMSE/MBER s.t. individual power constraints	Closed-form solution	[22]
		×	Decision feedback receiver	Source, relay and destination designs for MMSE/MBER s.t. individual power constraints	Closed-form solution; Mode selection	[51]
		○	Linear receiver	Relay and destination optimization for MMSE/MBER s.t. shaping power constraints	Closed-form solution; Limited feedback design	[52]
	M/ 1/ M	×	Linear receiver	Source, relay and destination optimization for MMSE s.t. individual and sum power constraints	Iterative solution	[15]
	M/ 1/ 1	○	Linear receiver	Source, relay and destination optimization for MMSE s.t. individual power constraints with imperfect channel infromation	Two-step / Three-step solutions	[43]
	1/ M/ 1	×	Linear receiver	Source, relay and destination optimization for MMSE s.t. individual power constraints	Closed-form solution; Iterative solution	[53]
Multi-Hop	M/ M/ 1	×	Linear receiver	Source, relay and destination optimization for MMSE and BER s.t. individual power constraints	Iterative solution; Closed-form solution	[20]
	1/ M/ M	×	Multi-casting/ Linear receiver	Source, relay and destination optimization for Min-Max MSE s.t. individual power constraints	Closed-form solution	[14]
	M/ M/ M	×	Multi-casting/ Linear receiver	Source, relay and destination optimization for Min-Max MSE s.t. individual power constraints	Closed-form solution	[54]

where $\mathcal{F}/\mathcal{B}$ denotes the local CSI based suboptimal designs with precoding strategies $\mathcal{F}$ and $\mathcal{B}$ at the source and the relay, respectively. For example, NP/NP indicates a suboptimal scheme having precoders $\mathbf{F} = \gamma_f \mathbf{I}$ and $\mathbf{B} = \gamma_b \mathbf{I}$ without the CSIT at both the source and the relay. Unlike the naive AF scheme, however, the NP/NP still requires the relay to know the CSIR $\mathbf{H}$ to compute the relay receiver $\mathbf{L}$. We note again that for all suboptimal designs, the channel estimation at the source or the destination is immediate from the classical methods in P2P MIMO systems, since only one-hop channel estimation is sufficient.

We may utilize codebook based designs such as NP/LP and LP/LP considering a common practice in current wireless communications, i.e., the CSIRs at the relay ($\mathbf{H}$) and the destination ($\mathbf{G}$) are obtained via training signals from the source and the relay, respectively, while the CSITs at the source ($\mathbf{H}$) and the relay ($\mathbf{G}$) are computed by the limited feedback from the relay and the destination, respectively. As for the LP/LP in Table 1, the relay and the destination independently feeds back k_F and k_B-bit quantized CSIs to the source and the relay for precoders $\mathbf{F}$ and $\mathbf{B}$, respectively. The amount of feedback bits k_F and k_B are not necessarily equal as long as the relay holds two separate codebooks. One interesting point here is that the relay matrix $\mathbf{Q}$ is built upon a combination of the quantized precoder $\mathbf{B}$ and the unquantized receiver $\mathbf{L}$. Our observation implies that the suboptimal design affords the relay to address different levels of CSI between $\mathbf{H}$ and $\mathbf{G}$, which cannot be addressed by the optimal design.

C. SUBOPTIMAL ROBUST DESIGNS WITH IMPERFECT LOCAL CSI

As long as the relay receiver is properly designed taking the channel uncertainty $\Delta\mathbf{H}$ into account, the MSE decomposition and relaxation in (3) still hold in a system with channel estimation errors. Therefore, simple robust designs in P2P MIMO systems [55] can directly be applied to both links separately with the local CSI at the source and the destination. This advantage not only enforces robustness to the errors due to the reduced CSI requirements at each node, but also provides improved flexibility in the transceiver designs, because the channel uncertainty of the link on the other side becomes petty considerations.

We note that unlike the cases without the channel estimation errors, the MSE decomposition and relaxation method does not ensure the optimality even in the high SNR regime, because the channel uncertainty boosts the relay noise along with the signal power. Nevertheless, it can be verified from computer simulations that our approach shows good performance for all SNRs regardless of the amount of channel uncertainty.

TABLE 3. Existing works for performance analysis using the MSE decomposition in MIMO AF relaying systems.

Categories	S/ R/ D	Direct Link	Tranceiver Features	Main Objectives	Main Results	Ref.
Two-Hop	1/ 1/ 1	◯	Linear receiver	DMT analysis of the closed-form solutions for MMSE s.t. individual power constraints	DMT analytical expressions	[22]
		×	Linear receiver	Uncoded BER analysis of the closed-form solutions for MMSE/MBER s.t. individual power constraints	Uncoded BER analytical expressions	[24]
		×	Linear receiver	DMT and DMR analysis of the closed-form solutions for MMSE/MBER s.t. individual power constraints	DMT and DMR analytical expressions	[28]
		×	Channel-coded OFDM/ Beamforming	Pairwise error probability analysis for channel-coded single-stream beamforming in MIMO-OFDM relaying systems	Channel-coded diversity gain analytical expressions	[25]
		×	Channel-coded/ Linear receiver	Pairwise error probability analysis for channel-coded SM relaying systems	Channel coded diversity gain analytical expressions	[26]

D. EXISTING WORKS BASED ON THE MSE DECOMPOSITION

The MSE decomposition method has been used in a number of recent papers to get optimal precoding solutions and analytical results for various MIMO AF relay networks. Table 2 and 3 summarize some of existing works based on the MSE decomposition. In each table, the second column indicates the number of the source, relay, and destination nodes in the system.

V. ANALYTICAL PERFORMANCE

In this section, we investigate analytical performance of the aforementioned relaying schemes in terms of the diversity gain and the shifting gain to attain more insight into the designs. In order to highlight the gains collected from the use of a multiple-antenna relay node, we restrict our focus on uncoded systems with the Rayleigh fading channels, but the result can be applied to more general cases.

A. DIVERSITY GAIN PERSPECTIVES

It is noted that since the diversity order characterizes the high SNR performance (specifically the decaying rate of a log-scale BER curve at high SNR [1]), each of the suboptimal designs and its corresponding optimal design achieves the same diversity order. The achievable diversity order of each relaying scheme is summarized in Table 4 [24], in which we obtain some interesting interpretations.

For some of open-loop SM relaying such as Opt-RP and NP/FP, we observe that the diversity order is dependent on the backhaul link channel only. Thus, a number of antennas at the destination may not be efficient in terms of the diversity order. The table also reveals that if we implement the NP/FP relaying such that the number of relay antennas is greater than that of destination antennas, a higher diversity order is attainable over the $N_t \times N_d$ P2P MIMO without the CSIT, while the diversity of the NP/NP never goes beyond that. Therefore, we can conclude that the CSIT, i.e., a proper knowledge on $\mathbf{G}$ and a sufficiently large number of antennas at the relay are essential to harvest a proper diversity advantage by deploying a relay node between a transmitter and a receiver.

In the meantime, the diversity order of the closed-loop relaying, i.e., Opt-JP and FP/FP is determined by the minimum of the backhaul and access links. Thus, increasing N_r would be a more efficient way to realize spatial diversity rather than increasing either N_t or N_d. It is also worthwhile noting that as long as N_r is greater than the maximum of N_t and N_d, the closed-loop relaying obtains a diversity gain over the P2P MIMO systems with the full CSIT.

B. SHIFTING GAIN PERSPECTIVES

In practical environments, it is also of interest to design the system so that the shifting (or coding) gain is maximized. This is because as the wireless propagation becomes more dispersive and the channel coding is employed over different frequency bands, the system becomes more likely to contain enough diversity [25]. In this case, the shifting gain plays an important role for improving performance rather than the diversity gain. In this subsection, we investigate the achievable shifting gain in the MIMO AF relaying.

It was shown for the closed-loop SM relaying such as FP/FP and Opt-JP that when the relay is located closer to the destination, using an additional antenna at the source provides a significant shifting gain over the case where N_t equals N_d [24]. Note that in this case, there is no diversity gain, since the diversity order is determined by the minimum of N_t and N_d. Let us set the case with $N_t = N_d$ as an initial state with initial state diversity $\mathcal{D}$. Then, it is verified that by increasing N_t, one can achieve a $\frac{10}{\mathcal{D}}\log(1+\alpha^{\zeta\mathcal{D}})$ dB shifting gain over the initial state in terms of the transmit power.

Consider that the relay is halfway between the source and the destination, i.e., $\alpha = 1$. Then, the achievable shifting

TABLE 4. Achievable diversity and shifting gains.

Schemes		Diversity Gain	Shifting Gain with initial state diversity $\mathcal{D}$	
			$N_t \uparrow$	$N_d \uparrow$
Closed-loop SM	FP/FP	$(N_r - N_s + 1)(\min(N_t, N_d) - N_s + 1)$	$\frac{10}{\mathcal{D}} \log(1 + \alpha^{\zeta\mathcal{D}})$	$\frac{10}{\mathcal{D}} \log(1 + \alpha^{-\zeta\mathcal{D}})$
	Opt-JP			
Open-loop SM	NP/NP	$\min(N_r, N_d) - N_t + 1$	None	None
	NP/FP	$N_r - N_s + 1$	None	$\frac{10}{\mathcal{D}} \log\left(1 + \frac{\alpha^{\zeta\bar{\mathcal{D}}} A}{N_s^{\bar{\mathcal{D}}}}\right)$
	Opt-RP			
P2P MIMO ($N_t \times N_d$)	NP	$N_d - N_s + 1$	None	None
	FP	$(N_t - N_s + 1)(N_d - N_s + 1)$		

TABLE 5. Number of floating point operations ($N_t = N_r = N_d = N$).

Schemes	Source ($\mathbf{F}$)	Relay ($\mathbf{Q}$)	Destination ($\mathbf{W}$)
Opt-JP	$24N^3 + 2N^2 + 14J_{\text{ini}}J_{\text{alt}}J_{\text{bis}}N$	$26N^3 + 2N^2 + 14J_{\text{ini}}J_{\text{alt}}J_{\text{bis}}$	$33N^3 + (14J_{\text{ini}}J_{\text{alt}}J_{\text{bis}} + 10)N$
Opt-RP	-	$26N^3 + 2N^2 + 7J_{\text{bis}}N$	$40N^3 + 4N^2 + 7J_{\text{bis}}N$
Naive AF	-	-	$10N^3 + 2N^2$
FP/FP	$12N^3 + 2N^2 + 7J_{\text{bis}}N$	$15N^3 + 2N^2 + (7J_{\text{bis}} + 2)N$	$N(N+1)^2$
NP/FP	-	$31N^3 + 2N^2 + 7J_{\text{bis}}N$	$4N^3 + N^2$
NP/NP	-	$4N^3 + N^2$	$4N^3 + N^2$

gain equals simply $3/\mathcal{D}$ dB meaning that the maximum gain equals 3 dB. On the contrary, when the relay is located closer to the destination or the destination moves towards the relay ($\alpha > 1$), a considerable performance gain is expected by employing an additional antenna at the source. For example, when $\alpha = 2$, $\mathcal{D} = 1$, and $\zeta = 4$, the gain equals 12.3 dB over the initial state. This result suggests that when designing the SM AF relaying system, the relay location must be considered as an important factor as well as the antenna configuration. This is also applicable to an efficient relay selection problem. More specifically, for given antenna configurations, we can choose the best relay utilizing the geometric information α in multi-relay networks. Note that for the case of open-loop SM relaying, no shifting gain is expected even if we increase the source antennas with $\alpha > 1$ due to the absence of a precoding operation at the source. By the symmetry of the systems, the same performance is achievable by locating the relay closer to the source with an additional antenna at the destination as shown in Table 4.

C. COMPLEXITY COMPARISON

Table 5 measures the computational complexity of various precoding schemes in terms of the number of floating point operations (flops). For ease of exposition, we assumed that the number of antennas at each node equals N. Here, J_{ini} and J_{alt} indicate the number of initial points and iterations required to obtain the optimal designs, respectively, and J_{bis} designates the numbers of iterations required in a bisection process to find a power allocation matrix, which is a common factor for both the optimal and suboptimal designs. We used the rules in [56] to compute the required number of flops in each operation.[2]

Our complexity analysis reveals that the total amount of computations at each node mostly determined by the singular value decomposition (SVD) and the matrix multiplication of the channel matrices. Therefore, the suboptimal designs based on the local CSI could be advantageous in terms of the reduced matrix computations especially at the source and the destination. According to the table, we can notice that for $N = 4$, NP/FP needs to compute only the 272 number of flops at the destination, while Opt-RP requires $2152 + 56J_{\text{ini}}J_{\text{alt}}J_{\text{bis}}$ number of flops which is typically more than 10 times the computation of NP/FP. A similar observation can be made between Opt-JP and FP/FP.

VI. NUMERICAL PERFORMANCE

In this section, we provide a discussion on the numerical performance of MIMO SM AF relaying systems. To clarify the effect of the channel estimation and provide more insightful observations, the computer simulations were performed by considering a simple uncoded transmission over independent Rayleigh fading channels, but trends may appear similarly in practical communication systems [2]. For fair comparison, we normalize the overall distance of the source-relay-destination link as 2 units. Thus, each channel coefficient

[2]For $N \times N$ matrices, the computational complexities of the SVD only for singular values and the right singular matrix, the total SVD, the matrix multiplication, and the matrix inversion are calculated as $12N^3$, $21N^3$, $2N^3$, and N^3, respectively. Also, the computational complexity of N-dimensional power loading matrix is given by $7J_LN$.

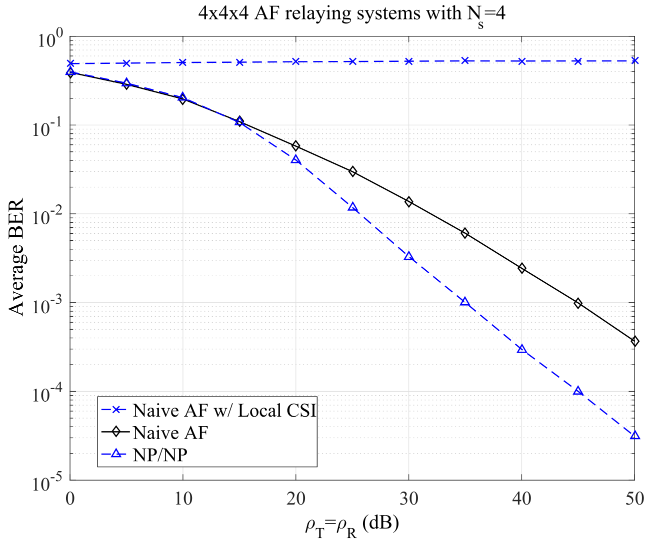

FIGURE 2. BER comparison of precoding schemes with perfect CSI and $\alpha = 1$ for 4-QAM.

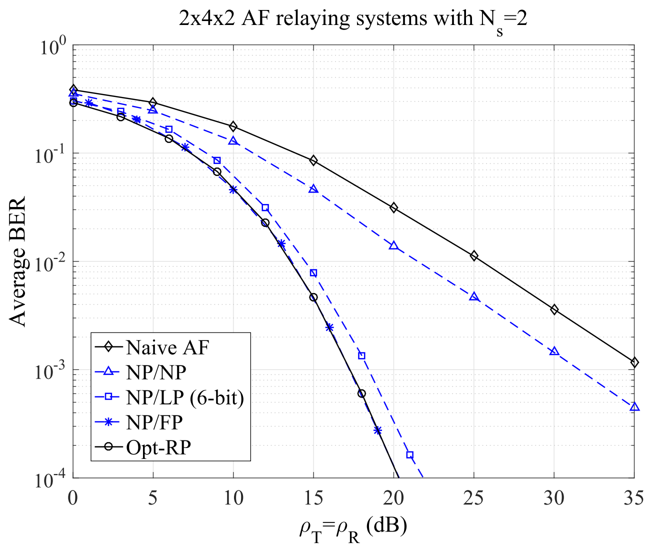

FIGURE 3. BER comparison of precoding schemes with perfect CSI and $\alpha = 1$ for 4-QAM.

of $\mathbf{H}$ and $\mathbf{G}$ are assumed to be complex Gaussian $\mathcal{CN}\big(0, (\frac{2\alpha}{1+\alpha})^{-\zeta}\big)$ and $\mathcal{CN}\big(0, (\frac{2}{1+\alpha})^{-\zeta}\big)$, respectively. Assuming the urban propagation, we set the pathloss exponent to be $\zeta = 4$.

When there exist channel estimation errors, each coefficient of uncertainty matrices $\Delta\mathbf{H}$ and $\Delta\mathbf{G}$ is also modeled by i.i.d. complex Gaussian $\mathcal{CN}(0, e_{h,i}^2)$ and $\mathcal{CN}(0, e_{g,i}^2)$, respectively, with the node index $i \in \{\text{source}(s), \text{relay}(r), \text{destination}(d)\}$. For example, $e_{h,d}^2$ denotes the uncertainty level of $\mathbf{H}$ at the destination. All simulation results are averaged over 10^5 independent realizations of the true channel matrices $\mathbf{H}$ and $\mathbf{G}$. All noise vectors are assumed to be standard complex Gaussian, i.e, $\mathcal{CN}(0, 1)$.

We define ρ_T and ρ_R as the transmit power at the source and at the relay, respectively. We adopt a notation $N_t \times N_r \times N_d$ to denote a system with N_t-source, N_r-relay, and N_d-destination antennas. All iterative schemes (e.g., Opt-JP [7] and robust Opt-JP [40]) are simulated with 100 random initial points, among which the best one is chosen. As for the suboptimal schemes FP and LP, we adopt the precoding method in [49] and the Grassmannian codebook method in [50], respectively. For ease of exposition, we assume an equal rate transmission scheme across the antennas using 4- and 16-QAM modulations. For all figures, the solid and dashed lines indicate the global and local CSI based designs, respectively.

In Figs. 2 and 3, we compare BER performance of various optimal and suboptimal designs assuming perfect channel estimation for both the local and global CSIs. Fig. 2 considers $4 \times 4 \times 4$ MIMO relaying systems with 4 data streams, assuming no CSIT at both the source and the relay. *"Naive AF w/ Local CSI"* denotes the naive AF relaying where only the local CSI $\mathbf{G}$ is allowed at the destination. Several interesting observations are made by comparing the NP/NP and the two naive schemes. First, we confirm that unlike the NP/NP, the naive scheme must have the global CSI at the destination to attain a reliable signal detection, The result illustrates that the relay receiver $\mathbf{L}$ at the relay in NP/NP is essential for reducing the CSI requirement at the destination. It is also interesting to observe that the NP/NP achieves a considerable shifting gain over the naive AF scheme with reduced CSI requirement at the destination. Although computing $\mathbf{L}$ at the relay may require an additional effort, the gain will far outweigh the cost.

Fig. 3 depicts the simulation plot for $2 \times 4 \times 2$ MIMO AF relaying with two data streams ($N_s = 2$). We first recognize that two precoding schemes at the relay, i.e., NP/FP and Opt-RP, exhibit almost equivalent performance over all SNR range, which confirms the optimality of the MSE decomposition and relaxation method. For NP/LP, we assume that a Grassmannian codebook [50] is shared between the relay and the destination. Interestingly, the performance of NP/LP is within 2 dB of the NP/FP with only six feedback bits. While all precoding schemes, e.g., Opt-RP, NP/FP, and NP/LP having CSIT at the relay achieve considerable diversity gains, the naive and NP/NP without the CSIT cannot. Therefore, a certain amount of CSIT at the relay is essential to harvest a proper diversity advantage when deploying a relay node.

Fig. 4 illustrates the numerical performance of FP/FP according to the relay location factor α and the number of source antennas N_t considering that $N_r = N_d = N_s = 2$ with the 16-QAM modulation. As predicted from Table 4, all curves yield an initial state diversity $\mathcal{D} = 1$ regardless of N_t. For the case of $N_t = N_d$, we observe that the shifting gain is maximized at $\alpha = 1$ and gradually decreases as the relay moves towards either of the source or the destination. It is worth noting that for the case of no CSIT at the source, i.e, NP/FP and Opt-RP, the optimal relay location which maximizes the shifting gain may not occur at $\alpha = 1$, due to

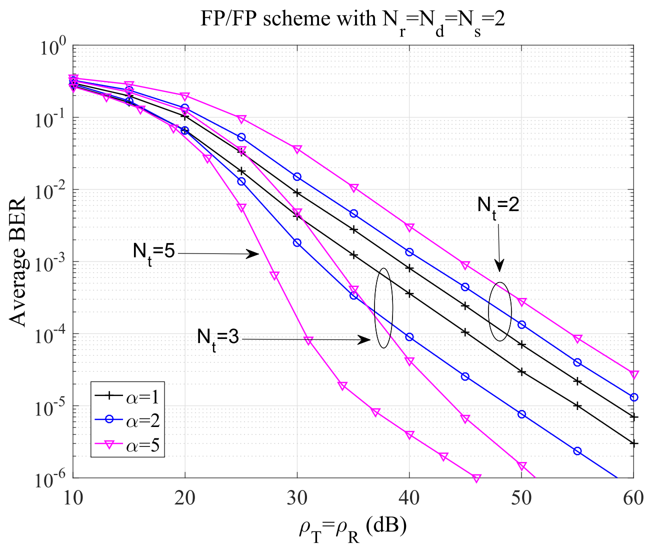

FIGURE 4. BER comparison according to various α and N_t for 16-QAM.

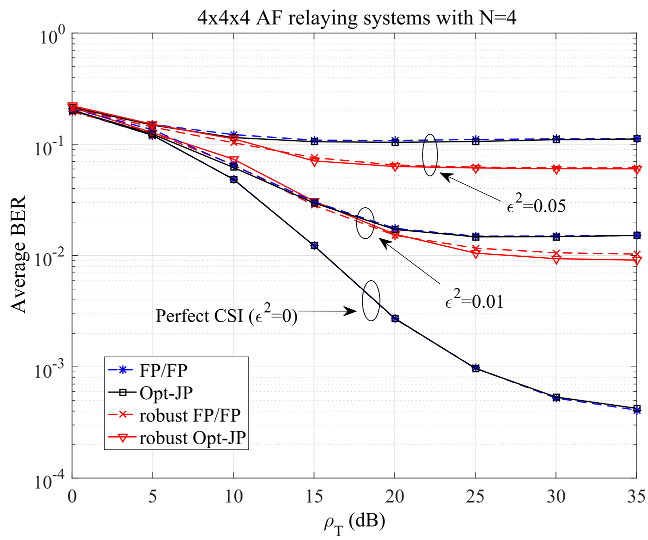

FIGURE 6. BER comparison of joint precoding schemes with imperfect CSI, $\alpha = 1$, and $\rho_R = 25$ dB for 4-QAM.

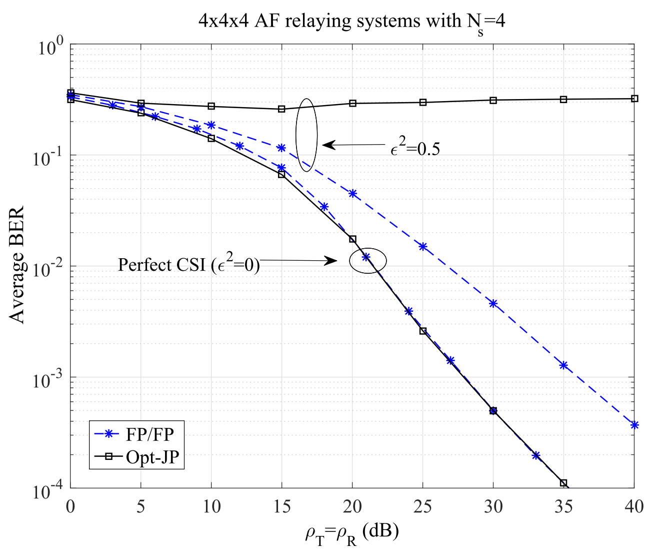

FIGURE 5. BER comparison of joint precoding schemes with $e^2_{h,s} = e^2_{h,d} = e^2_{g,s} = e^2_{g,r} = \epsilon^2$, $e^2_{h,r} = e^2_{g,d} = 0$, and $\alpha = 1$ for 4-QAM.

the asymmetry in the precoding strategy [24]. This figure also confirms our observation made in Section V-B that when the relay is positioned closer to the destination with $N_t > N_d$, a substantial shifting gain is attainable. For instance, the curves with $\alpha = 2$ and 5 illustrate that the use of an additional antenna at the source enables us to obtain gains of 12.3 dB and 28 dB over the initial state, respectively. It is also interesting to observe that the more we increase N_t, the faster the system approaches its theoretical shifting gain. For cases where α equals $1/2$ and $1/5$, i.e., the relay is located closer to the source, the same result can be drawn by increasing N_d from the initial state.

In Figs. 5 and 6, we examine the closed-loop SM relaying with imperfect CSI at each node. In practice, it is often the case that the channel estimate at the receiver sides is sufficiently accurate thanks to the pilot signals from the transmitters, whereas the CSIs available through the channel feedback or feedforward process are likely to be imperfect [55]. In Fig. 5, we consider such a practical situation where the relay and the destination perfectly know their corresponding CSIRs, i.e., $e^2_{h,r} = e^2_{g,d} = 0$, but all other CSIs based on the feedback or feedforward remains imperfect. Interestingly, we see that although the performance of the Opt-JP deteriorates by the channel uncertainties, the NP/NP is insensitive to such errors. This is because the destination of the FP/FP makes no use of the backhaul link CSI $\mathbf{H}$, and thus is not affected by such a CSI mismatch.

In Fig. 6, we compare the performance of several robust designs assuming that channel estimation errors occur for all channels at all nodes as $e^2_{h,i} = e^2_{g,i} = \epsilon^2, \forall i$. Since all nodes experience the equal amount of uncertainty for each channel matrix, the robust design methods based on the global CSI [40] is well defined, and thus the system can be fully optimized. Nevertheless, we observe that the local CSI based robust design still achieves the performance close to the global CSI based designs with reduced channel estimation complexity. Thus, we can conclude that the MSE decomposition and relaxation technique maintains the MSE optimality even with the channel estimation errors, and thus could be an appropriate solution for implementing the SM MIMO AF relaying systems.

VII. CONCLUSION

In this article, we have studied challenges and solutions for implementing SM AF relaying techniques. An overview of conventional designs and major challenges for their practical implementation were provided. Through in-depth discussion on the transceiver designs and their analytical performance, we have demonstrated that the MSE decomposition and a slight optimality relaxation provide an efficient solution for those challenges. It is particularly interesting to observe that the MSE decomposition based approach exhibits good performance in a practical system with channel estimation errors, and thus could be an attractive solution for implementing the SM AF relaying systems. Finally, simulation results were used to demonstrate the ideas in our discussions.

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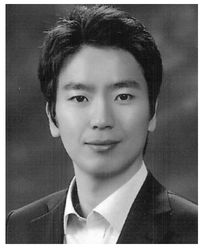

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