Leveraging on the Impact of Imperfect Channel Estimation for MIMO Relaying Systems

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ABSTRACT In this paper, we study a multi-user multiple-input multiple-output (MIMO) dual-hop wireless relaying system, where the decode-and-forward protocol is applied at the intermediate relay node. The well-known MIMO spatial multiplexing technique is employed, where the linear yet efficient zero-forcing detection is utilized at the receiver side of each hop. The practical scenario of channel estimation errors and/or delayed channel response is embraced when the signals experience mutually independent Rayleigh fading. In oppose to the approach followed in the vast majority of the published works, in this manuscript the channel estimation error term is reckoned as a signal rather than interference or noise, which may significantly enhance the overall system performance. Analytical results involving outage probability and average symbol error rate are extracted in closed-form. Capitalizing on the above performance metrics, new engineering insights are manifested highlighting clearly the superiority of the proposed approach against the conventional one when a delayed/imperfect channel response is present.

INDEX TERMS Decode-and-forward (DF), imperfect channel estimation, multiple-antenna transmission, zero-forcing detection.

I. INTRODUCTION

One of the most challenging issues that modern wireless communication systems face is the densely occupied spectrum bandwidth that urges the need for spectrally efficient methods. Multiple-input-multiple-output output (MIMO) communication is already a mature key technology for achieving higher bit rates by employing spatial multiplexing technique [1]. Additionally, the challenging multiuser MIMO (MU-MIMO) setups have recently attracted much research interest in utilizing the spatial multiplexing method so as to enhance the multiuser diversity [2]. In particular, spatial multiplexing independent data streams that combined with both linear and non-linear (e.g., maximum likelihood) detection schemes have been adopted in MU-MIMO systems. Linear detectors (e.g. zero-forcing (ZF)) are frequently used due to less computational complexity, but with suboptimal performance though [3]–[8].

In most realistic networking systems, the acquisition of perfect channel state information (CSI) between the transmitter and receiver links represents an ideal and rather overoptimistic condition, infeasible in practice. This occurs to a large extent due to unknown user mobility, rapid fading variations, and/or lack of feedback signaling. Typically, in order to obtain CSI estimates, training methods are usually utilized via trained pilot signals and/or appropriate signal processing between the involved transceiver pairs [9]–[11]. To date, most literature works conventionally handle the impact of channel estimation error as a noise-like segment upon the detection of the received signal. It was shown in [12], [13] that the channel estimation error can be processed in a more efficient basis, by treating it as a signal rather than noise, since the error term (along with its associated transmitted symbol counterpart) can be demodulated via envelope detection; thereby, it can further enhance the overall system performance.

Motivated by the latter observation, we study spatial multiplexed dual-hop relaying MU-MIMO systems operating over multipath Rayleigh channels and ZF detection at the receiver end. Specifically, the decode-and-forward (DF) protocol is...
adopted at the intermediate relay node. Moreover, prior to data transmission, a training phase occurs for CSI acquisition. In contrary to how the channel estimation term is used in the majority of published works, we leverage the channel estimation error by considering it as a signal rather than interference or noise, which may boost the received signal-to-noise ratio (SNR) and, consequently, the overall system performance [12]. Unlike [12] (the spatial diversity was analyzed therein), we focus on the spatial multiplexing mode of operation over quasi-static Rayleigh fading channels i.e., the Γ(·) denotes the Gamma function [14, Eq. (8.310.1)].

The impact of imperfect CSI onto the system performance is explicitly analyzed and numerically evaluated. Insightfully, some new engineering insights are highlighted, such as the superiority of the introduced approach in the presence of CSI imperfection and/or intense user mobility. Capitalizing on the latter result, a new dual-hop system configuration is proposed when the nodes in one of the two communications hops remain static. In fact, a new ‘mixed’ detection scheme is presented, which appropriately combines the conventional and proposed approach, preserving the end-to-end diversity order; even in the presence of delayed/imperfect CSI.

Notation: Matrices are denoted by uppercase bold typeface letters, whilst I_v stands for the v × v identity matrix. Superscript (·)^H denotes Hermitian transposition and |·| represents absolute (scalar) value. X_{ij} denotes the element at the i^{th} row and j^{th} column of matrix X. E[·] is the expectation operator and symbol d denotes equality in distribution. f_X(·) and F_X(·) represent the probability density function (PDF) and cumulative distribution function (CDF) of a random variable (RV) X, respectively. Re{·} and Im{·} denote the real and imaginary parts of a complex-valued x, respectively. C_N(μ, σ^2) and N(μ, σ^2) define, respectively, a complex-valued and real-valued Gaussian RV with mean μ and variance σ^2, and X^2(u, w) denotes a non-central chi-squared RV with u degrees of freedom (DoF) with a non-centrality parameter u and variance w. Q_1(·, ·) is the first-order Marcum-Q function and I_p(·, ·) is the Kummer’s confluent hypergeometric function [14, Eq. (9.210.1)]. Also, (·)_p is the Pochhammer symbol with p ∈ N [14, p. xliii] and Γ(·) denotes the Gamma function [14, Eq. (8.310.1)].

II. SYSTEM AND SIGNAL MODEL

A dual-hop wireless relaying system with M single-antenna transmitters, a relay node equipped with N_r ≥ M antennas, and a destination node equipped with N_d ≥ N_r antennas operating over quasi-static Rayleigh fading channels i.e., the fading remains fixed for the duration of a given frame transmission, while they may change independently amongst various frames. The distributed MU-MIMO scenario analyzed in this paper can be considered equivalent to the classical single-user MIMO system with M co-located antennas at the transmitter side. At each consecutive frame and before the actual data transmission phase, a training phase is executed at the receiver side for CSI acquisition. This is implemented through pilot symbols whereas the classical minimum mean squared error (MMSE) channel estimation is being implemented at the receiver side (i.e., at the relay and destination), while full-blind transmitters are assumed. In the current work, the spatial multiplexing technique is deployed, where M independent data streams are simultaneously transmitted by the corresponding nodes. During the data communication phase, the efficient ZF detection scheme is applied at the receivers’ side. Also, the direct link between transmit/source and destination nodes is assumed to be absent due to the strong propagation attenuation of the end-to-end received signal [15], [16].

More specifically, at the end of training phase, the relation between the actual channel status at the (n + 1)^{th} time instance and its estimated counterpart at the n^{th} time instance at the relay node reads as

\[ H[n+1] = \hat{H}[n] + E[n+1], \quad (1) \]

where H ∈ C_{N_r}^{N_r × M}, \hat{H} ∈ C_{N_r}^{N_r × M} and E ∈ C_{N_r}^{N_r × M} denote the actual channel, estimated channel and channel estimation error (incorporating errors caused by both imperfect and delayed CSI), respectively. In what follows, the time-instance index n is dropped for ease of presentation, since all the involved random vectors are mutually independent. Also, without delving into details of this well-investigated subject, the interested reader may refer to, e.g., [7] for a thorough analysis of the derivation in (1). For the i^{th} column of H, say h_i, it holds that \( h_i \overset{d}{=} C\mathcal{N}(0, g_i I_{N_r}) \) where g_i stands for the large-scale fading between the i^{th} transmitter and the relay. Also, \( \hat{h}_i \overset{d}{=} C\mathcal{N}(0, \sigma^2_{h_i} I_{N_r}) \) and \( e_i \overset{d}{=} C\mathcal{N}(0, \sigma^2_{e_i} I_{N_r}) \), where

\[ \sigma^2_{h_i} = \frac{\alpha^2 p_i g_i^2}{1 + p_i g_i}, \quad (2) \]

and

\[ \sigma^2_{e_i} = g_i - \frac{\alpha^2 p_i g_i^2}{1 + p_i g_i}, \quad (3) \]

with \( p_i \overset{d}{=} M \omega_m, p, \) and \( \alpha \) denoting the total transmit power (from all the subscribed users) at the training phase, transmit power per user, and temporal correlation parameter reflecting the severity of delayed CSI, respectively. Typically, by considering the fakes fading model, \( \alpha = J_0(2 \pi f_{D} T_s) \), where \( J_0(·) \) is the first-kind Bessel function, \( T_s \) is the channel sampling duration, while \( f_{D} \) is the maximum Doppler frequency shift determined by the nodes’ velocity and transmission frequency. It holds that \( 0 \leq |\alpha| \leq 1 \), where \( |\alpha| = 1 \) denotes non-delayed CSI, while a smaller \( |\alpha| \) value indicates higher CSI versatility and stronger CSI mismatch due to the time delay. Notably, using MMSE channel estimation, \( \hat{h}_i \) and \( e_i \) are mutually independent due to the orthogonality principle.

Subsequently, the data transmission phase is started and the received signal can be expressed as

\[ y_t = \sqrt{p} H_s + w, \quad (4) \]
where \( y_r \in \mathbb{C}^{N_c \times 1}, s \in \mathbb{C}^{M \times 1}, w \in \mathbb{C}^{N_c \times 1} \) denote, respectively, the received signal, transmitted signal vector (where \( s^H s = I_M \)), and white Gaussian noise at the data phase. ZF detection via the efficient QR decomposition is applied at the receiver and the detected signal becomes

\[
r = \hat{Q}^H y_r = \sqrt{p} \hat{Q}^H (\hat{H} + \hat{E}) s + \hat{Q}^H w
\]

where \( \hat{Q} \) and \( \hat{R} \) are the \( N_c \times N_r \) unitary matrix (with its columns representing the orthonormal ZF nulling vectors) and \( N_r \times M \) upper triangular matrix, respectively, given \( \hat{H} \) (i.e., \( \hat{H} \cong \hat{Q} \hat{R} \)). Consequently, assuming that \( w \cong \mathcal{CN}(0, I_{N_c}) \) (and hence \( p \) reflects the transmit SNR), the received SNR of the \( i \)-th stream (\( 1 \leq i \leq M \)) at the first communication hop, defined as \( \gamma_i^{(1)} \), reads as

\[
\gamma_i^{(1)} = \frac{p |\hat{R} + \hat{Q}^H \hat{E}|_i^2}{E|\hat{Q}^H w|_i^2} = \frac{p |\hat{R} + \hat{Q}^H \hat{E}|_i^2}{E|\hat{Q}^H w|_i^2}
\]

where the last equality due to the fact that \( w \) is a zero-mean RV and \( E|\hat{Q}^H w|_i^2 = E(|\hat{Q}^H w|_i^2) = |\hat{Q}^H E[|w^H \hat{Q}|] = |\hat{Q}^H Q| \), while \( Q \) is a unitary matrix. Notice that the channel estimation error term within (5) is treated as a signal (since it carries transmit symbol information); hence, it may further augment the received SNR.\(^1\)

To this end, the derived SNR expression in (6) is accurate and reflects a clear physical meaning. On the other hand, the conventional approach is to treat the channel estimation error as a noise-like term, yielding the classical SINR expression

\[
\gamma_{i, \text{conventional}}^{(1)} = \frac{p |\hat{R}|_i^2}{p |\hat{Q}^H E|_i^2 + 1}
\]

Proceeding to the second-hop (regarding the relay-to-destination link), the aforementioned process is repeated in quite a similar basis. Therefore, the said system model applies also for the second communication hop by simply substituting \( N_c \) with \( N_d \). It is noteworthy to state that the relay uses \( M \) out of \( N_t \) antennas for transmission\(^2\) during the second hop (whenever \( N_t > M \)).

### III. PERFORMANCE METRICS

#### A. OUTAGE PROBABILITY

Given \( \hat{H} \) (and thus \( \hat{R} \)), the SNR expression in (6) introduces a non-central chi-squared RV, namely, \( \gamma_i^{(1)} \cong \chi^2_2(p |\hat{r}_{i,i}|^2, \sigma_i^2) \).

\(^1\)In practice, this process can be implemented by adopting the classical coherent detector (given \( \hat{H} \)), followed by a simple envelope detector along with a low-pass filter, so as to capitalize on the fluctuations of the unknown parameter \( E_s \). It turns out that although the latter parameter is unknown, it may be impactful since it carries out the actual symbol information.

\(^2\)Since there is no CSI available at the transmit side (i.e., the relay and destination acquire CSI of the transmitters/source nodes and relay node, respectively, during the training phase), a fixed antenna allocation is utilized when enabling \( M \) out of \( N_t \) antennas for transmission.
Hence, $F_{\gamma_i}(\cdot)$ can be easily, accurately and rapidly computed, especially in the case when $R_k \gg T_k > 1$. Also, for the special case of equal number of transmit and receive antennas (i.e., $N_t = M$) and with the aid of [20, Eq. (07.20.03.0026.01)], (13) reduces to

$$F_{\gamma_i}(\cdot) = 1 - \exp\left(-\frac{2M + \frac{2g_i}{p} - M\alpha^2}{1 + g_ipM(1 - \alpha)^2}\right).$$

(15)

Moreover, it is straightforward to show that the corresponding CDF of SNR at the destination hop during the second transmission hop is given in (14) by replacing $N_i$ with $N_d$. Capitalizing on the DF relayed protocol, the end-to-end SNR scenario, i.e., by setting $\alpha = 1$, for the non-delayed CSI and asymptotically high $\gamma$ performance can be greatly enhanced. In the presence of nodes’ mobility (i.e., $\alpha < 1$), the performance of the proposed approach is being enhanced for smaller values of $\alpha$ (i.e., stronger mobility; higher users’ velocity). This should not be surprising since, according to (6) and given the instantaneous status of $\mathbf{R}$, the contribution of channel estimation error onto the received effective channel gain is being increased for smaller $\alpha$ values (note that the channel estimation error is a zero-mean complex Gaussian RV with variance $g_1 - \sigma^2_1g_1^2 \approx|\hat{g}_1|^2/g_1^2$). Quite the opposite outcome yields by adopting the conventional approach since the contribution of channel estimation error is placed at the denominator of (7).

### B. ASYMPTOTICALLY HIGH SNR REGIME

To obtain more impactful insights, we proceed with the analysis of the considered scheme when an asymptotically high transmit SNR is being applied. Based on (13), using the identity $\lim_{x \to 0} \gamma_{e2e}(1 + 1, 2, x) \to 1$, assuming that $\alpha < 1$, and taking $p \to +\infty$, it stems that

$$F_{\gamma_{e2e}(1,p\to+\infty)}(\cdot) \approx 1 - \exp\left(-\frac{2x}{g_2(1 - \alpha)^2p}\right) \approx \frac{2x}{g_2(1 - \alpha)^2p} \to 0^+, \quad j \in \{1, 2\},$$

(17)

where the expansion $\exp(-x) \approx 1 - x$, for $x \to 0^+$ was used in the second equality of (17). Using the latter approximation into (16), the asymptotic outage probability of the $i$th stream tends to zero, as expected, since the SNR expression in (6) grows without bound for $p \to +\infty$. In fact, the asymptotic end-to-end outage probability can be tightly approximated by adding the asymptotic CDF of each hop given in (17), i.e., $F_{\gamma_{e2e}(1,p\to+\infty)}(\cdot)$, yielding a corresponding decay $\propto 1/p$; hence preserving a diversity order of one. For the non-delayed CSI and asymptotically high transmit SNR scenario, i.e., setting $\alpha = 1$ and then $p \to +\infty$ in (14), the CDF of SNR for the $j$th hop becomes $F_{\gamma_{e2e}(j,p\to+\infty)}(\cdot) \approx 1 - \exp(-Mx)$, reflecting a (fixed) outage floor proportional to the number of simultaneously transmitted streams $M$.

On the other hand, by adopting the conventional approach given in (7) and taking $p \to +\infty$, it yields a bounded expression for the $j$th hop, such that

$$\gamma_{1j,\text{conventional}} \approx |\hat{R}_{i,j}^2|/\hat{Q}^M_E_{i,j}^2.$$

(18)

The latter expression introduces an irreducible outage/error floor, which is independent of $p$ and is related only to the channel fading severity. Furthermore, recall that $e_i \sim \mathcal{C}N(0, \sigma^2_1I_{N_c})$ with $\sigma^2_1 \equiv g_1 - \sigma^2_p/g_1^2$, which implies that the detrimental impact of channel estimation error entirely vanishes when $\alpha = 1$ (i.e., for a non-delayed CSI). Only for the latter ideal case the conventional received SNR per hop grows unbounded for increasing values of $p$, which represents a key difference from the proposed approach.

Summarizing on the aforementioned observations, the following remarks are extracted:

- **Remark 1:** In the presence of nodes’ mobility (i.e., $\alpha < 1$), the performance of the proposed approach is being enhanced for smaller values of $\alpha$ (i.e., stronger mobility; higher users’ velocity). This should not be surprising since, according to (6) and given the instantaneous status of $\mathbf{R}$, the contribution of channel estimation error onto the received effective channel gain is being increased for smaller $\alpha$ values (note that the channel estimation error is a zero-mean complex Gaussian RV with variance $g_1 - \sigma^2_1g_1^2 \approx|\hat{g}_1|^2/g_1^2$). Quite the opposite outcome yields by adopting the conventional approach since the contribution of channel estimation error is placed at the denominator of (7).

- **Remark 2:** In the case when there is no mobility ($\alpha = 1$), i.e., whenever the involved nodes are static, the proposed approach provides a certain outage performance floor, irrespective of the transmit SNR. In this application scenario, on the other hand, the conventional approach seems more appealing, whereby the system performance can be greatly enhanced.

### C. AVERAGE SYMBOL ERROR RATE

Errors at the symbol decoding stage occur either when the source-to-relay transmission is received correctly and the relay-to-destination transmission is received in error, or vice versa. Hence, for any given modulation, we get

$$P_s,i(\gamma_{1i}^{(1)}), \gamma_{1i}^{(2)} \triangleq \left[1 - P_s,i(\gamma_{1i}^{(1)})\right]P_s,i(\gamma_{1i}^{(2)}) + \left[1 - P_s,i(\gamma_{1i}^{(2)})\right]P_s,i(\gamma_{1i}^{(1)}),$$

(19)

where $P_s,i(x, y), P_s,i(\cdot)$ and $P_s,i(\cdot)$ represent the end-to-end symbol error rate, symbol error rate of the first hop and symbol error rate of the second hop, respectively, regarding the $i$th stream and given the instantaneous SNR values per hop $\{x, y\}$. Therefore, ASER is computed as

$$\overline{P}_{s,i} = \mathbb{E}_{\gamma_{1i}^{(1)}, \gamma_{1i}^{(2)}}[P_s,i(\gamma_{1i}^{(1)}, \gamma_{1i}^{(2)})].$$

(20)
Since \( \{\gamma_{i}^{(1)}, \gamma_{i}^{(2)}\} \) are mutually independent RVs, it holds that
\[
\overline{P}_{s,i} = \left(1 - \overline{P}_{s,i}^{(1)}\right) \overline{P}_{s,i}^{(2)} + \left(1 - \overline{P}_{s,i}^{(2)}\right) \overline{P}_{s,i}^{(1)},
\]
where \( \overline{P}_{s,i}^{(j)} \) denotes ASER of the \( i^{th} \) stream for the \( j^{th} \) transmission hop. Therefore, it suffices to compute \( \overline{P}_{s,i}^{(j)} \). It holds that [21]
\[
\overline{P}_{s,i}^{(j)} = \frac{A\sqrt{B}}{2\sqrt{\pi}} \int_{0}^{\infty} x^{-\frac{3}{2}} \exp(-Bx) F_{\gamma_{i}^{(j)}}(x) dx,
\]
where \( A \) and \( B \) are specific constants that define the modulation type. Inserting (14) into (22), with the aid of [14, Eq. (3.351.3)], we arrive at the following closed-form expression:
\[
\overline{P}_{s,i}^{(j)} = \frac{A}{2} \left(1 - \left(\frac{B}{B + \frac{2M - \frac{2}{g} - \frac{M}{g} - \frac{M}{g}}{1 + g/pM(1-\alpha^2)}\right)^{\frac{1}{2}} \right)
- \sum_{l=1}^{N_j - M} \sum_{k=0}^{l-1} \frac{\Gamma(k + \frac{1}{2})}{\Gamma(l + 1)} \cdot \delta_{k \frac{1}{2}} \left[1 - \alpha^2 \left(1 - \frac{1}{1 + g/pM}\right)^l \right]
\times \frac{(1 - l)k(-1)^{k-1}}{k!(2)_k} \left[\frac{\alpha^2 M}{1 + g/pM(1-\alpha^2)}\right]^{k+1},
\]
where \( N_j = \{N_r, N_d\} \) for \( j = \{1, 2\} \), correspondingly. Then, using (23) into (21), ASER can be computed in an exact closed formula.

IV. NUMERICAL RESULTS, DISCUSSION AND ENGINEERING INSIGHTS

In this section, the derived analytical results are corroborated via Monte-Carlo simulations. Also, without loss of generality and for the sake of clarity, hereinafter, we assume an identical statistical profile for each transmitted stream (i.e., \( g_i \triangleq g = 1 \) and \( \gamma_{i}^{(1)} \triangleq \gamma_{i}^{(2)} \) \( \forall i; \) corresponding to a co-located multi-antenna transmitting source node) in order to evaluate more concretely the overall system performance (on average) and obtain more impactful insights. Also, line-curves and solid dot-marks denote the analytical and simulation results, respectively.

Figure 1 demonstrates the end-to-end system performance in terms of outage probability vs. various values of the transmit SNR. Obviously, the performance of the proposed approach is being enhanced for smaller \( \alpha \) values (i.e., a more intense nodes’ mobility). Conversely, the performance of the conventional approach is getting worse for smaller \( \alpha \) values. This is in accordance to the previous analysis and the remarks provided in subsection III-B, due to the different treatment regarding the impact of channel estimation errors. Moreover, the asymptotic curves of Fig. 1 indicate the tightness of the corresponding asymptotic expression, as per (17), in the high SNR regime.

In a similar basis, Fig. 2 illustrates the end-to-end ASER when a binary phase shift-keying (BPSK) modulation scheme and moderate transmit SNR of 10dB are being applied. Various system configuration setups are cross-compared with an emphasis on the large receive antenna scale (tending to the massive antenna principle). Obviously, the spatial multiplexing gain provided by the increasing number of receive antenna elements, given a fixed number of transmit antennas, is beneficial only for low mobility environments. Whenever the channel estimation errors are quite severe and strong user mobility is present, the proposed approach outperforms the conventional one for all the provided antenna scales. Also, it is worthy to state that the spatial multiplexing gain of the proposed approach is almost negligible as \( \alpha \) tends to zero (i.e., extremely high user mobility, quite a bad received CSI). This effect is reasonable in such an extreme condition; whereby even an increasing number of receive antenna elements cannot provide an effective spatial multiplexing gain. The latter observation is verified by (17), since even in the asymptotically high SNR regime the CDF of SNR is convex with respect to \( \alpha \), i.e., \( \frac{d^2}{d\alpha^2} F_{\gamma_{i}^{(j,p+\infty)}}(\cdot) > 0 \).

Capitalizing on the analysis and derived outcomes ob-
obtained from the previous section, we define a system configuration where the relay adopts the proposed approach (due to the presence of nodes’ mobility at the transmit side) during the stream detection at the first communication hop, while the conventional approach is being applied at the destination side. According to the analysis and derived outcomes obtained from the previous section, hereinafter, we propose and evaluate a dual-hop DF system where the relay and destination are static nodes. For instance, consider fixed base-stations and/or access points, which play the role of relay and destination nodes in a typical uplink networking system. Also, the transmitted streams are being generated from mobile source nodes with arbitrary mobility, placed within a reference coverage area (e.g., a classical macro- or micro-cell of a wireless cellular infrastructure). Then, in the first communication hop between source and relay, \( \alpha < 1 \) should hold, whereas \( \alpha = 1 \) occurs for the second communication hop regarding the static relay-to-destination channel link. It is noteworthy that the presence of imperfect channel estimates is always present at both communication hops; yet, the delayed received CSI restriction is only removed from the second hop. According to the analysis and derived outcomes obtained from the previous section, we define a system configuration where the relay adopts the proposed approach (due to the presence of nodes’ mobility at the transmit side) during the stream detection at the first communication hop, while the conventional approach is being applied at the destination during the second hop since the involved nodes remain static.

The latter scheme is named ‘mixed’ and abbreviated as ‘mix’ in Fig. 3. Also, the performance of mixed scheme is compared with the performance of the conventional approach, when it is being applied for both hops, regardless of the presence of nodes’ mobility. It is clear that the performance of the conventional scheme is getting worse for a more intense nodes’ mobility (i.e., smaller \( \alpha \) value), whereas an unavoidable error floor (independent of the transmit SNR) is introduced. Interestingly, the performance of the mixed scheme is greatly enhanced in comparison to the former scheme for smaller \( \alpha \) values, as expected. Last yet not least, it is worthy to note that the error floor, present in the conventional approach, is entirely vanished in the mixed scheme. This occurs because the proposed scheme preserves the diversity gain, even in the presence of delayed/imperfect CSI (in contrast to the conventional approach). Doing so, the performance difference of the two schemes is much more impactful in the high SNR regime, which is usually the case in most practical application scenarios.

V. CONCLUSION

A new multiuser MIMO dual-hop DF relaying approach was analytically studied under Rayleigh faded channels and imperfect/delayed CSI conditions. The ZF detection was used at the receiver side, whereas the channel estimation error was treated as a signal rather than noise as conventionally utilized so far. New exact closed-form expressions were derived with respect to the end-to-end outage probability and ASER. Further, some useful engineering insights were revealed, such as the superiority of the proposed approach in the presence of delayed CSI. Moreover, a novel method for dual-hop setups was presented where the system diversity order is being preserved; even in the presence of imperfect/delayed CSI.

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